Lecture 12
2024/2025

Microwave Devices and Circuits for Radiocommunications

2024/2025

- 2C/1L, **MDCR**
- **Attendance at minimum 7 sessions (course or laboratory)**
- Lectures- **associate professor Radu Damian**
  - Tuesday 16-18, ~~Online~~, P8
  - E – 50% final grade
  - **problems** + (2p atten. lect.) + (3 tests) + (bonus activity)
    - first test L1: 20-27.02.2024 (t2 and t3 not announced, lecture)
    - 3att.=+0.5p
  - **all materials/equipments authorized**

2024/2025

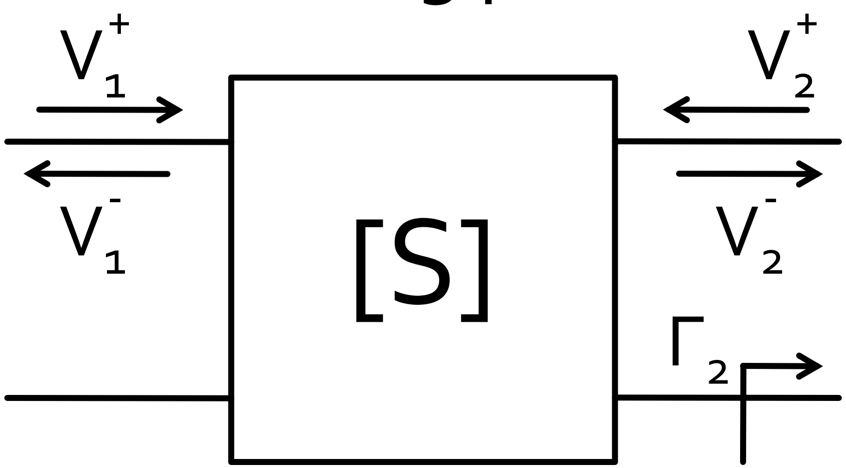
- Laboratory – **associate professor Radu Damian**
  - Tuesday 10-12, II.13 / (even weeks)
  - L – 25% final grade
    - ADS, 4 sessions
    - Attendance + **personal results**
  - P – 25% final grade
    - ADS, 3 sessions (-1? 25.02.2025)
    - personal homework

General theory

Microwave Network Analysis

Scattering matrix – S

- Scattering parameters



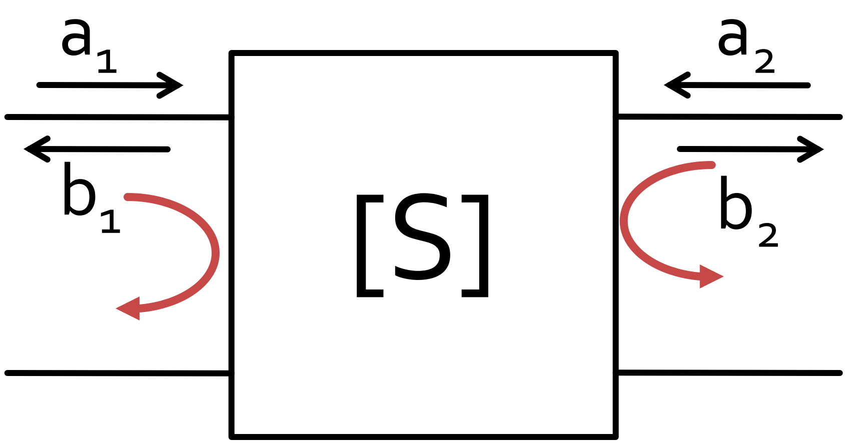
$$\begin{bmatrix} V_1^- \\ V_2^- \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \cdot \begin{bmatrix} V_1^+ \\ V_2^+ \end{bmatrix}$$

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

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

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

Scattering matrix – S

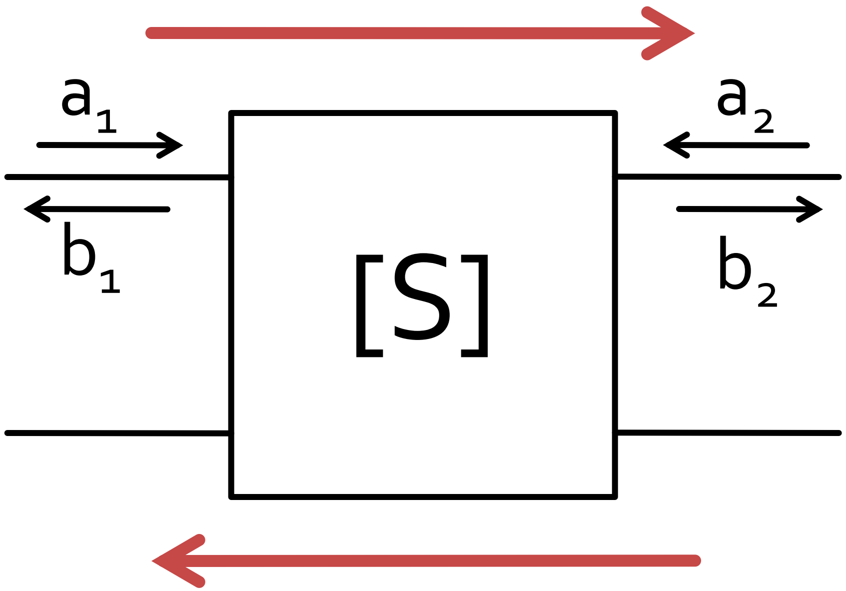


$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \cdot \begin{bmatrix} a_1 \\ a_2 \end{bmatrix}$$

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

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

Scattering matrix – S

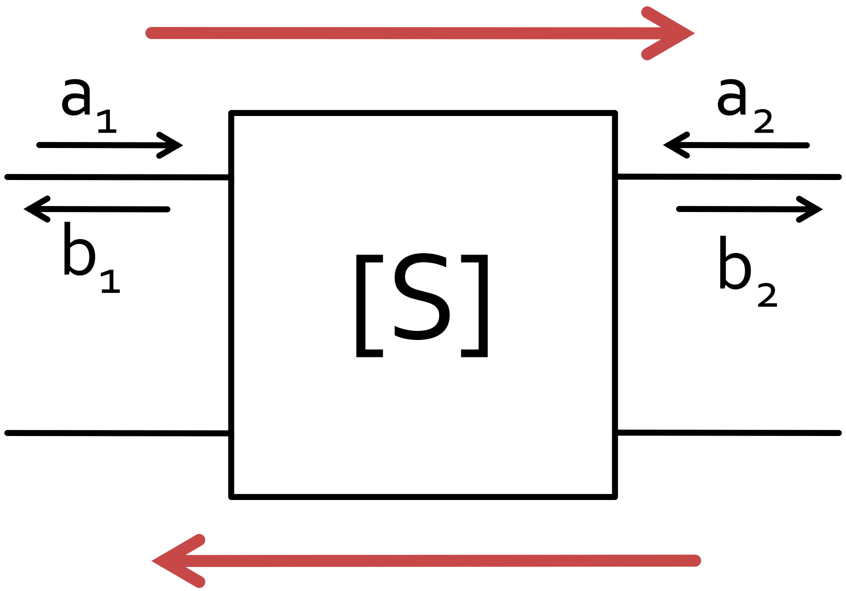


$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \cdot \begin{bmatrix} a_1 \\ a_2 \end{bmatrix}$$

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

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

Scattering matrix – S



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

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

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

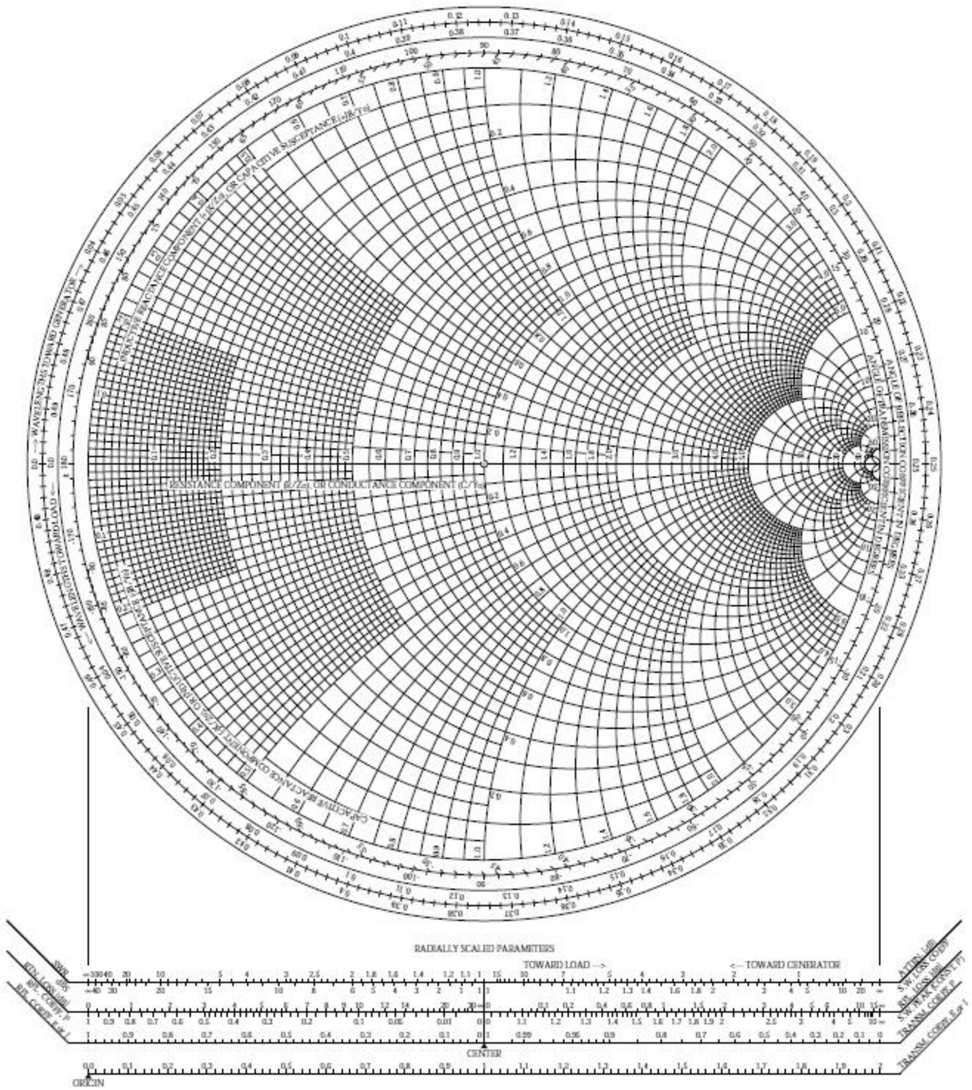
Impedance Matching

The Smith Chart

Course Topics

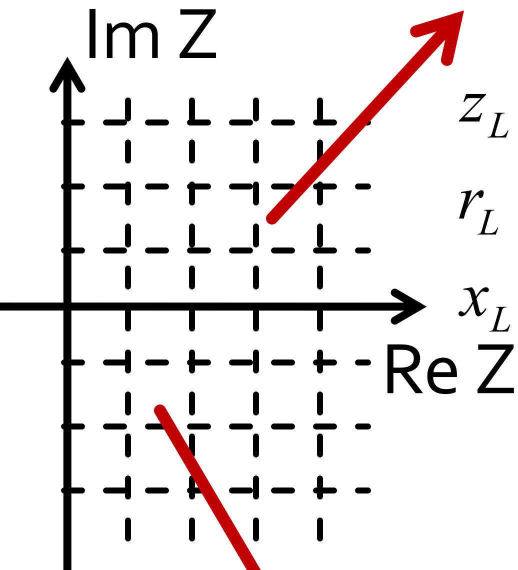
- Transmission lines
- Impedance matching and tuning
- Directional couplers
- Power dividers
- Microwave amplifier design
- Microwave filters
- ~~Oscillators and mixers~~ ?

The Smith Chart



The Smith Chart

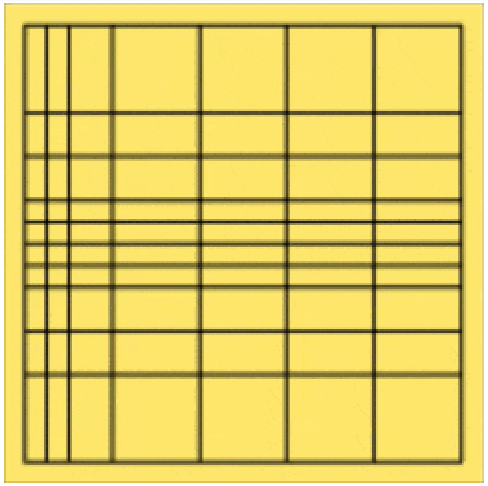
$$\Gamma = \frac{Z_L - Z_0}{Z_L + Z_0} = \frac{z_L - 1}{z_L + 1}$$

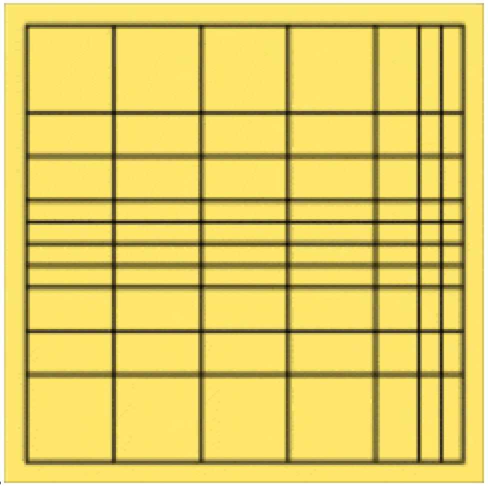


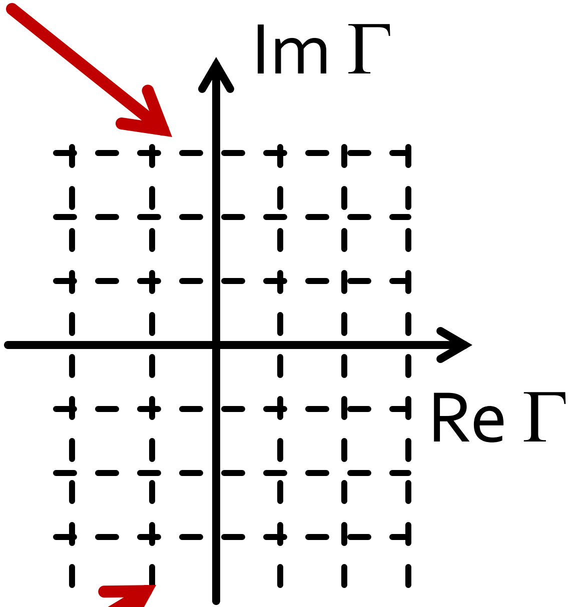
$$z_L = r_L + j \cdot x_L$$

$$r_L = ct.$$

$$x_L = ct.$$

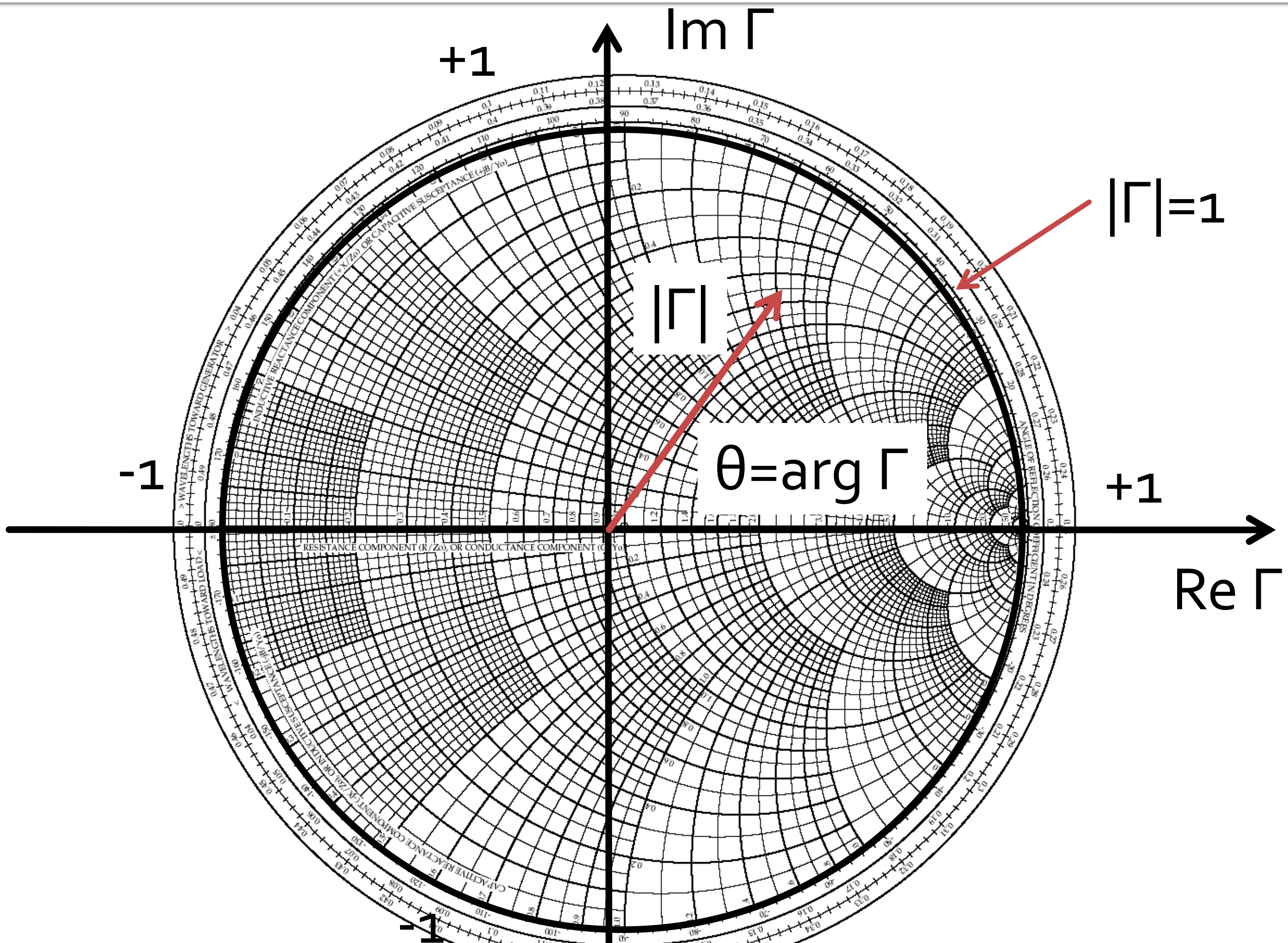




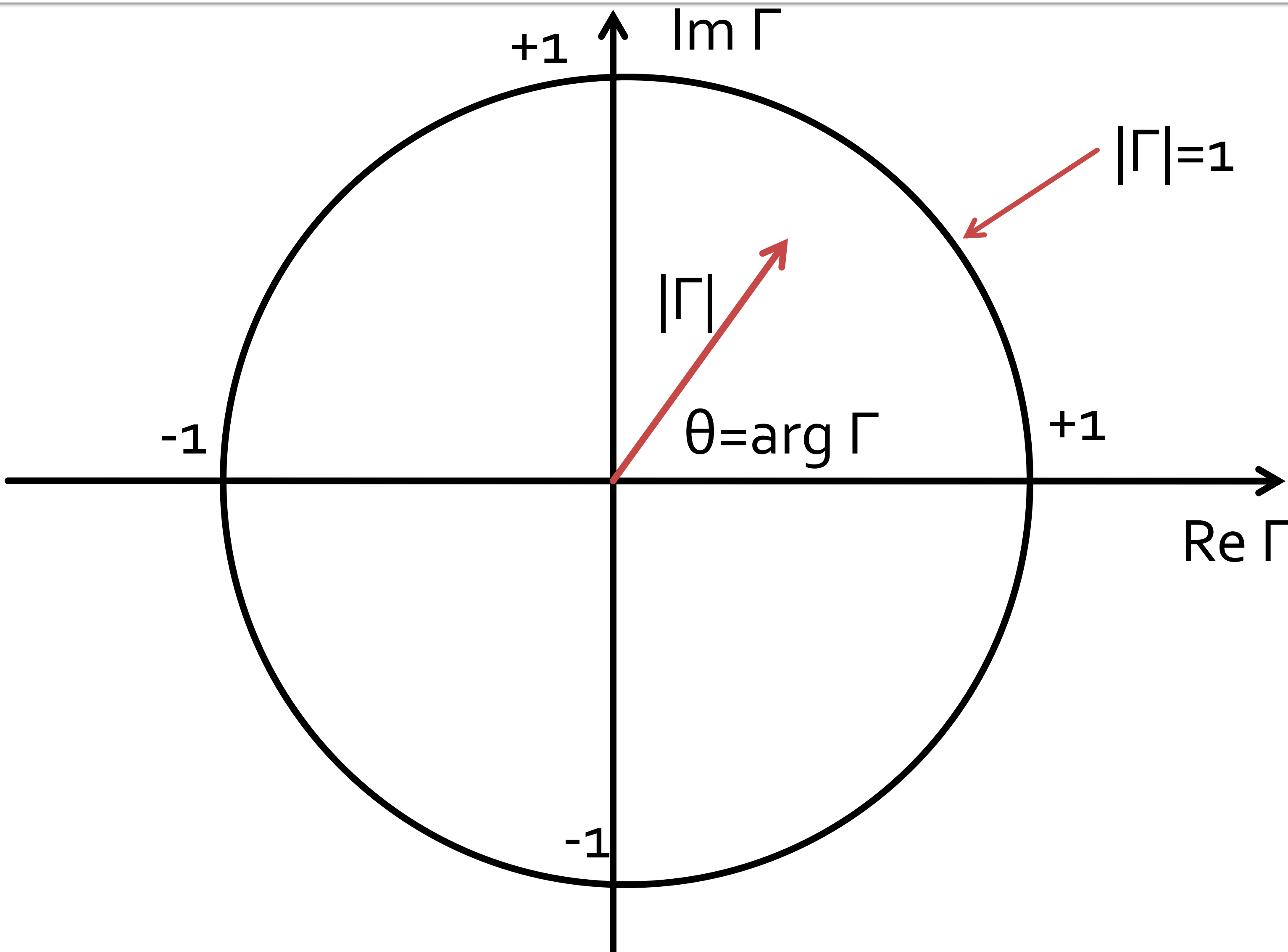


$$\Gamma = \frac{Z_L - Z_0}{Z_L + Z_0} = \frac{Y_0 - Y_L}{Y_0 + Y_L} = \frac{1 - y_L}{1 + y_L}$$

The Smith Chart



The Smith Chart



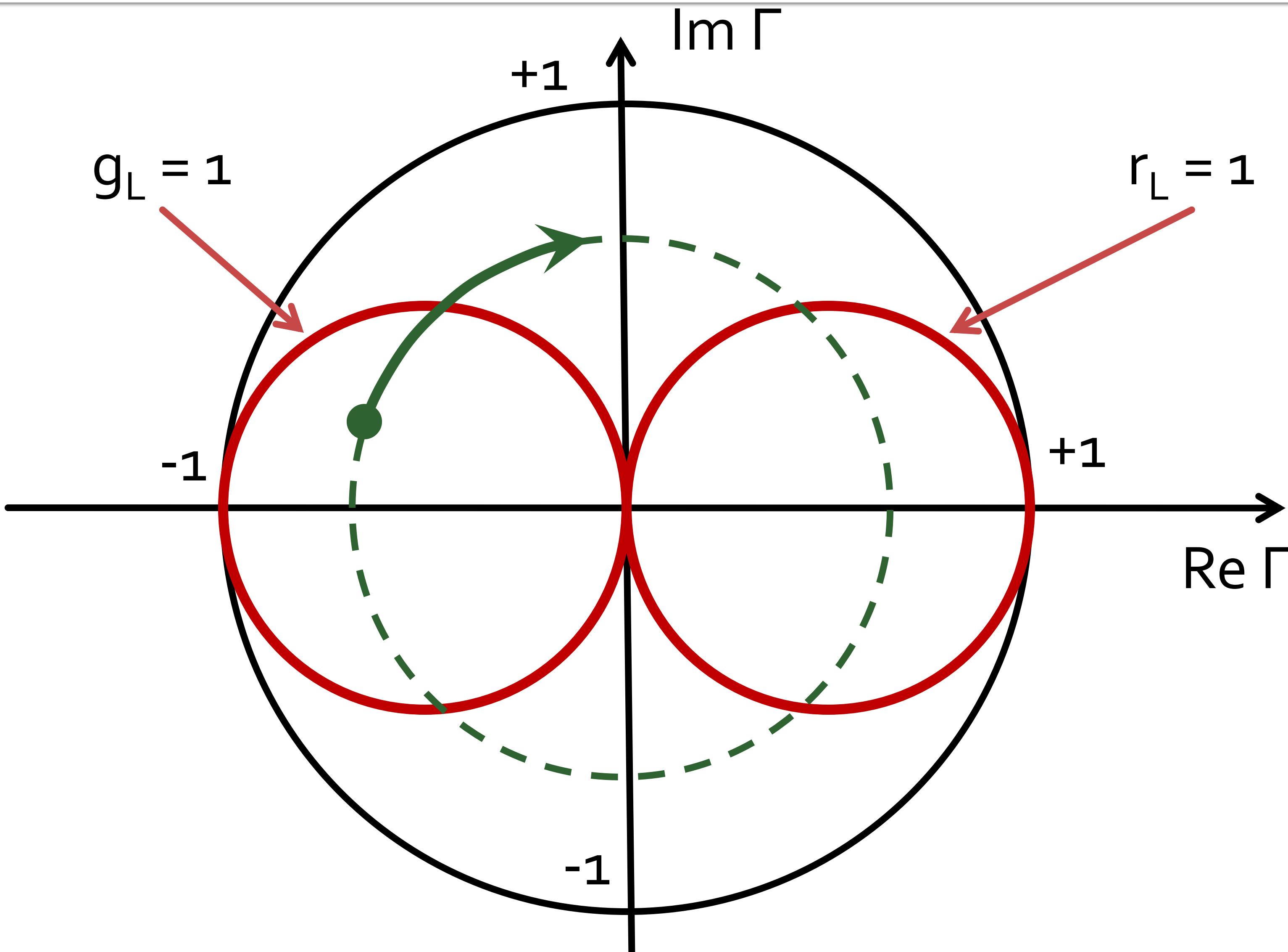
Impedance Matching

Impedance Matching with Stubs

Course Topics

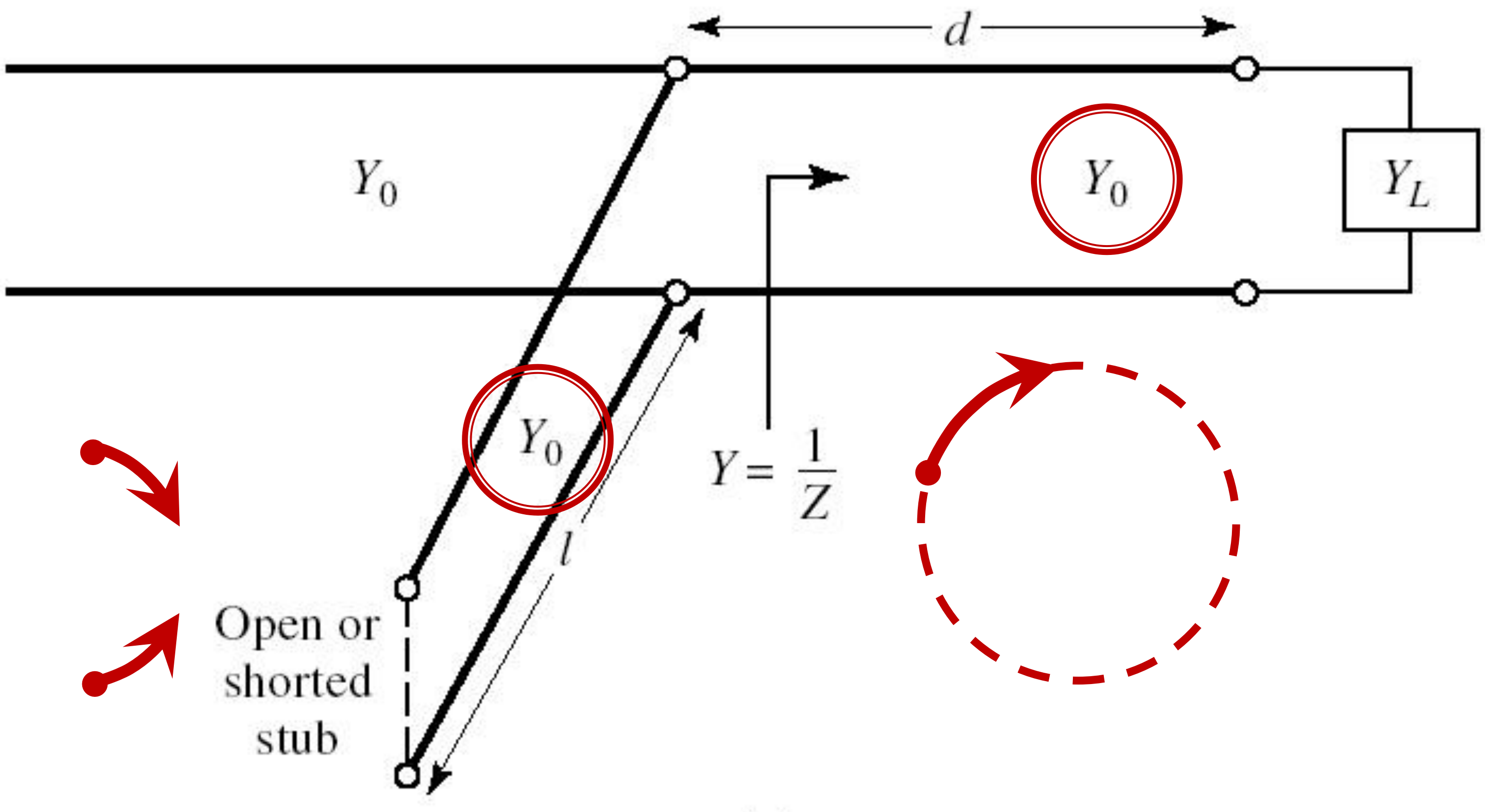
- Transmission lines
- Impedance matching and tuning
- Directional couplers
- Power dividers
- Microwave amplifier design
- Microwave filters
- ~~Oscillators and mixers~~ ?

Smith chart, r=1 and g=1



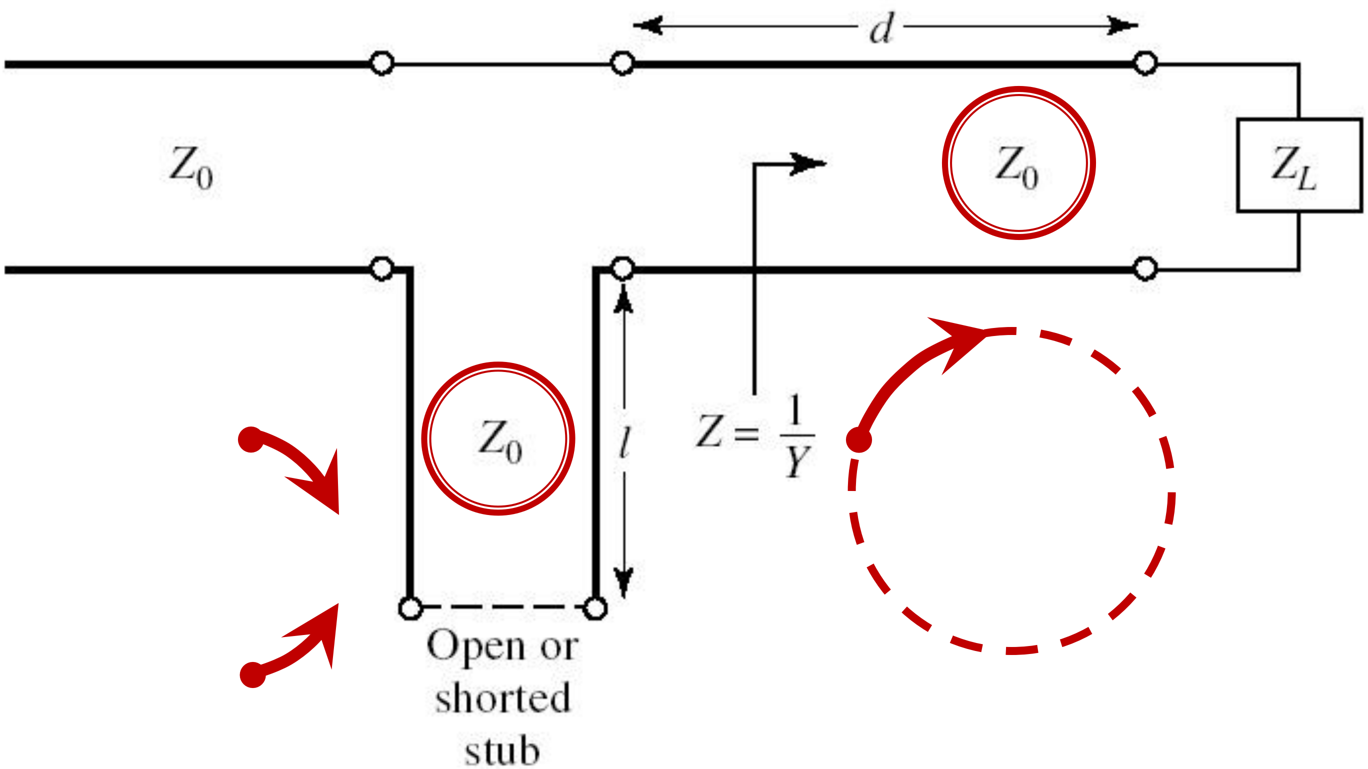
Single stub tuning

- Shunt Stub



Single stub tuning

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

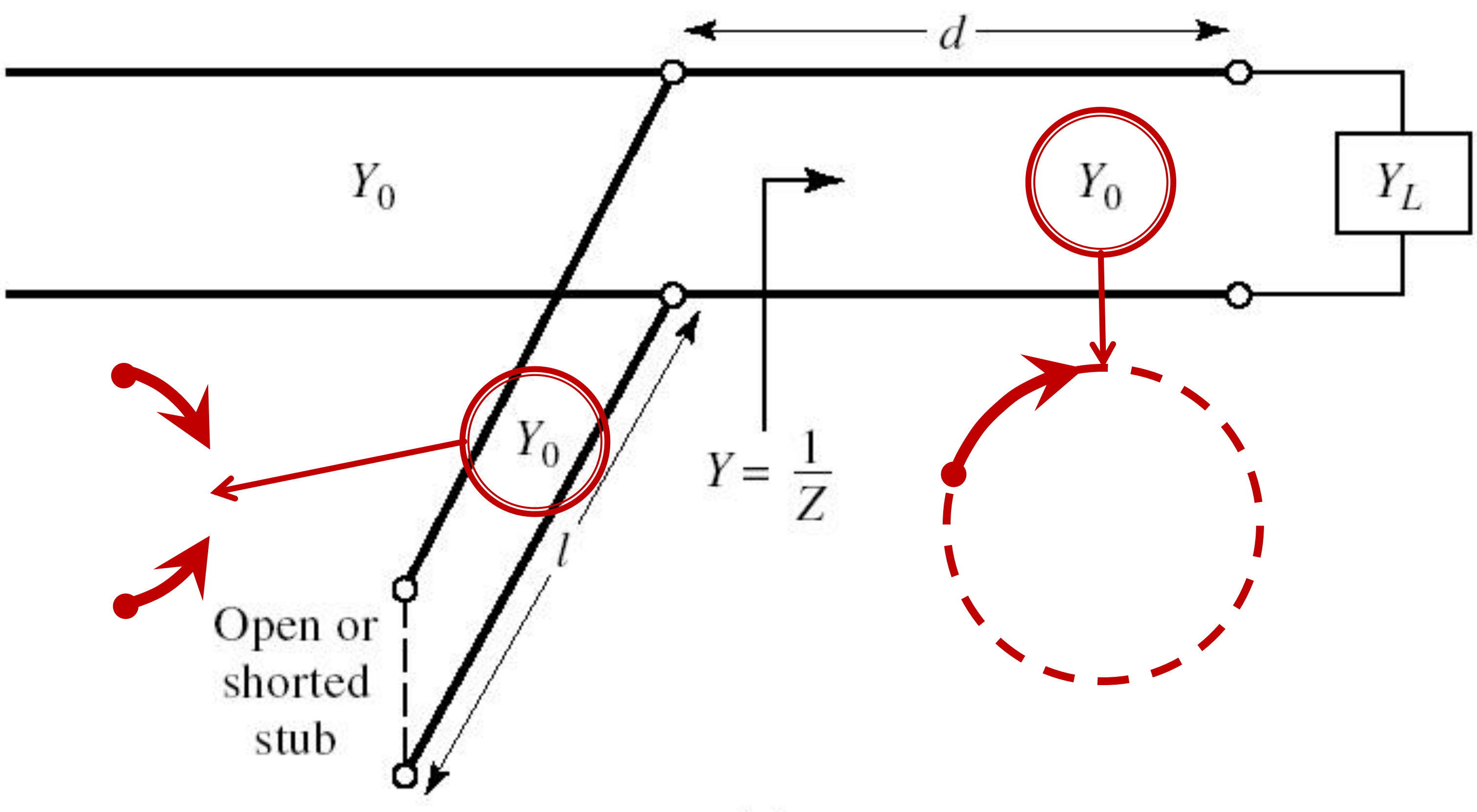


Analytical solutions

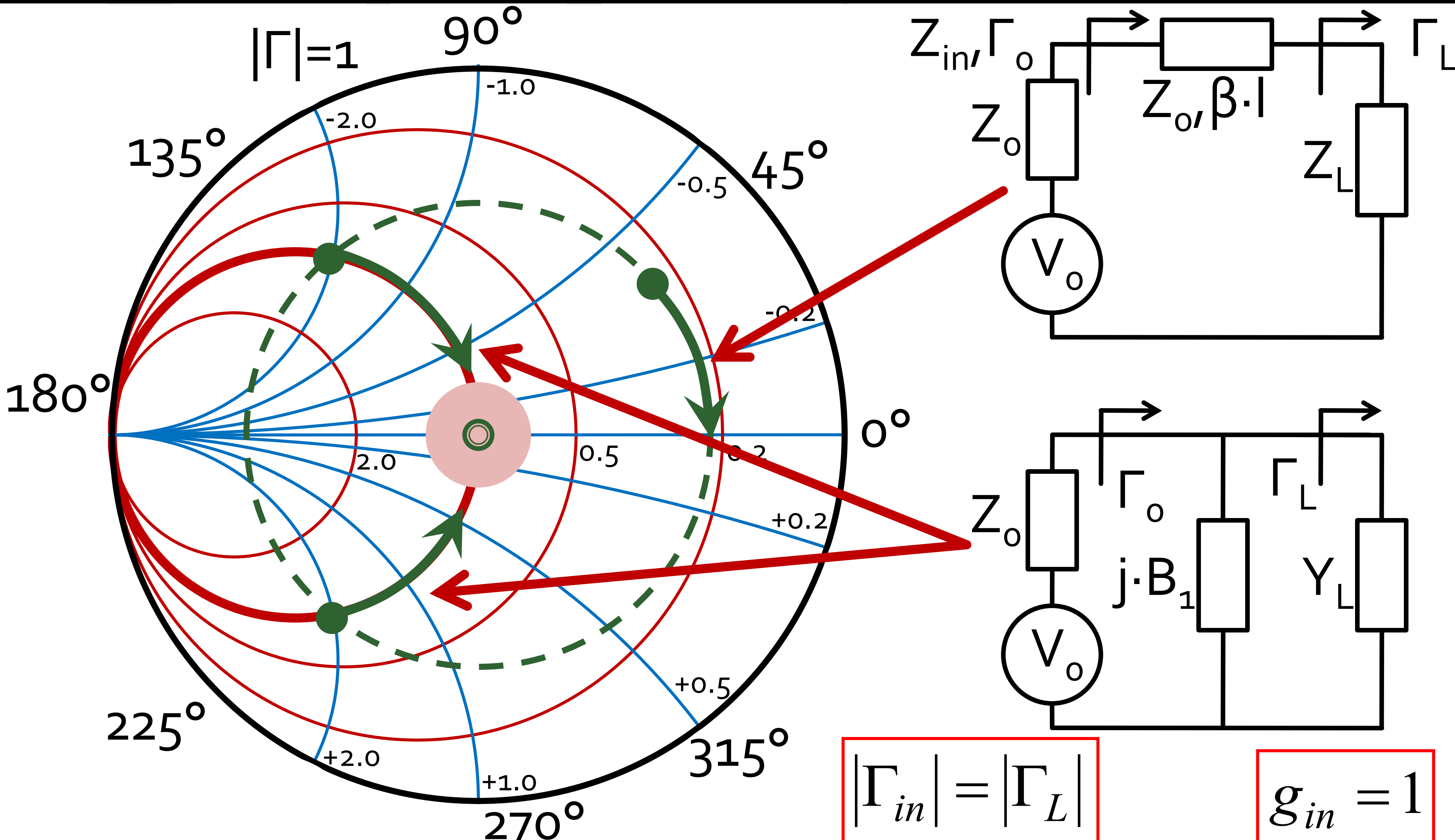
Exam / Project

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

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

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

$$\Gamma_S=0.593\angle 46.85°$$

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

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

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

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

  - **"-" solution**

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

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

Analytical solution, usage

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

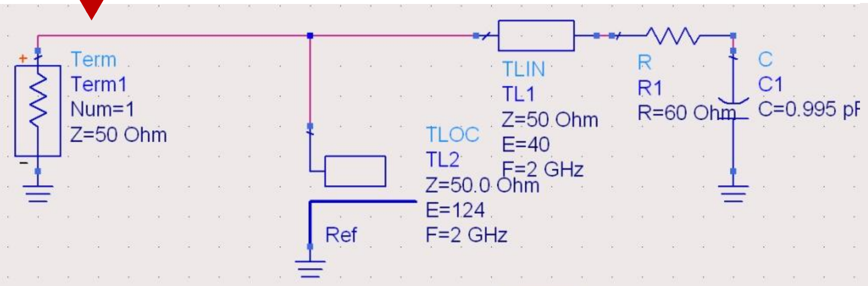
- We choose **one** of the two possible solutions
- The **sign** (+/-) chosen for the **series line** equation imposes the **sign** used for the **shunt stub** equation

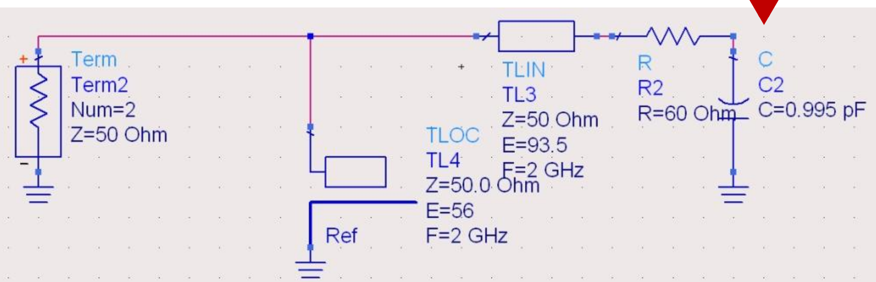
$$l_1=\frac{39.7^\circ}{360^\circ}\cdot\lambda=0.110\cdot\lambda$$

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

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

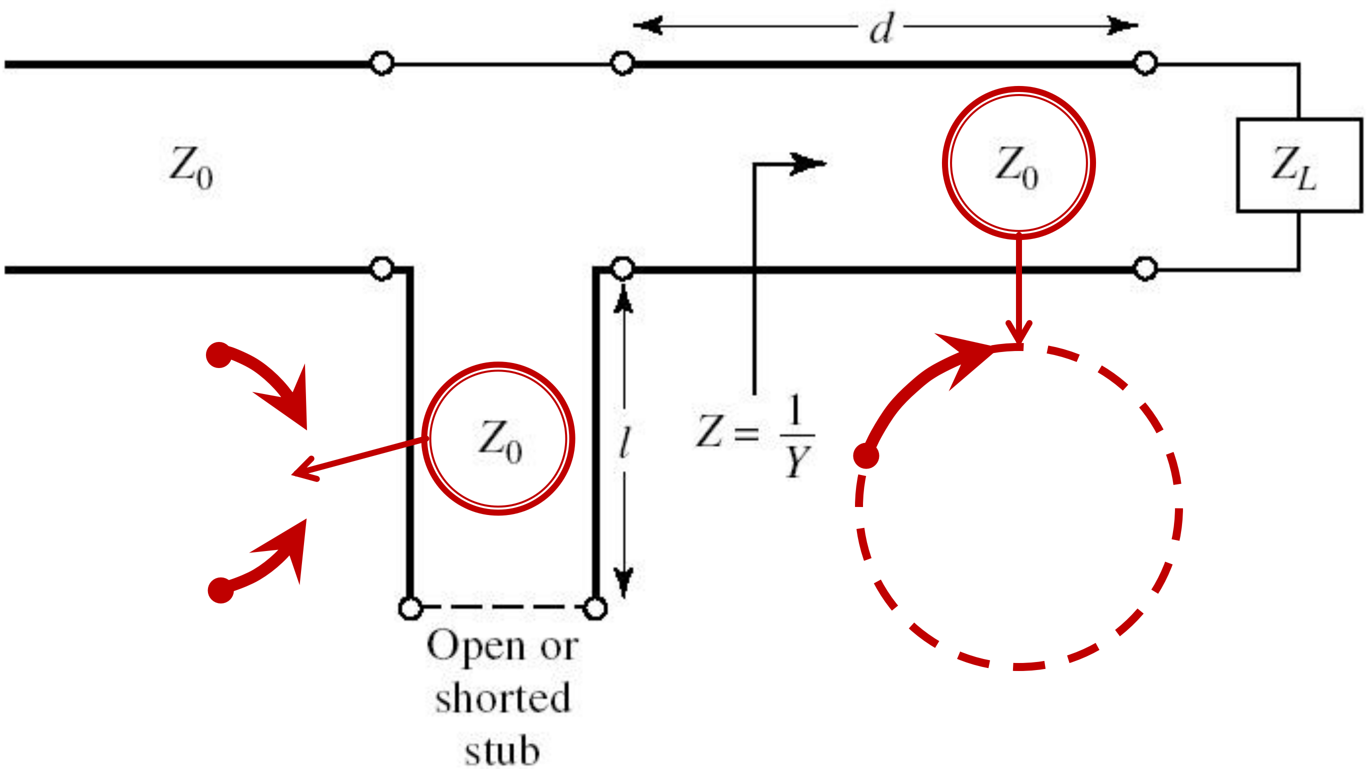
$$l_2=\frac{55.8^\circ}{360^\circ}\cdot\lambda=0.155\cdot\lambda$$



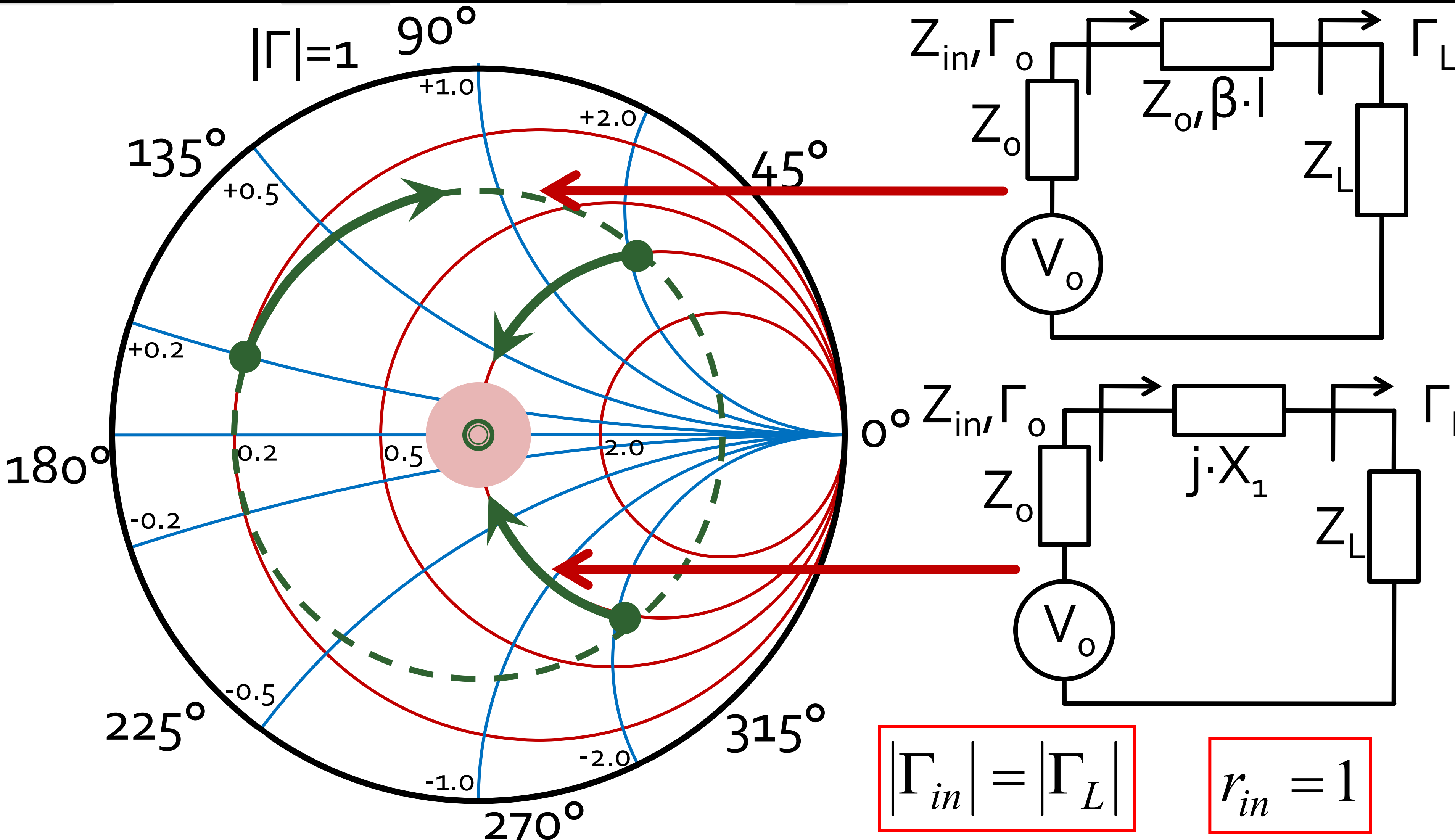


Case 2, Series Stub

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



Matching, series line + series reactance



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

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

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

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

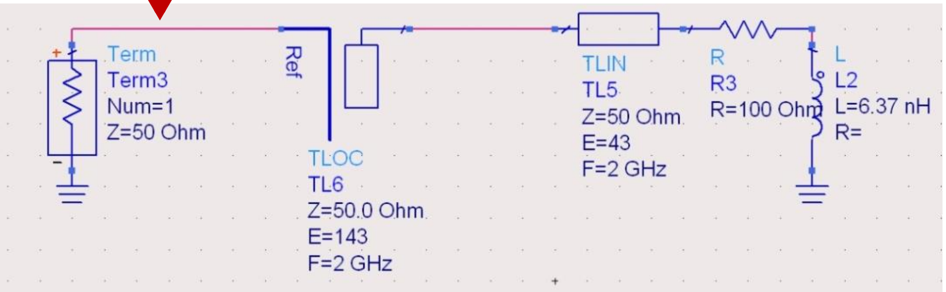
- We choose **one** of the two possible solutions
- The **sign** (+/-) chosen for the **series line** equation imposes the **sign** used for the **series stub** equation

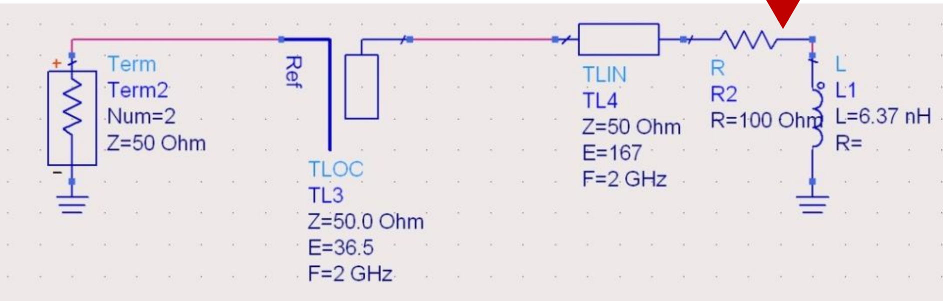
$$l_1 = \frac{43.1^\circ}{360^\circ} \cdot \lambda = 0.120 \cdot \lambda$$

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

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

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





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)

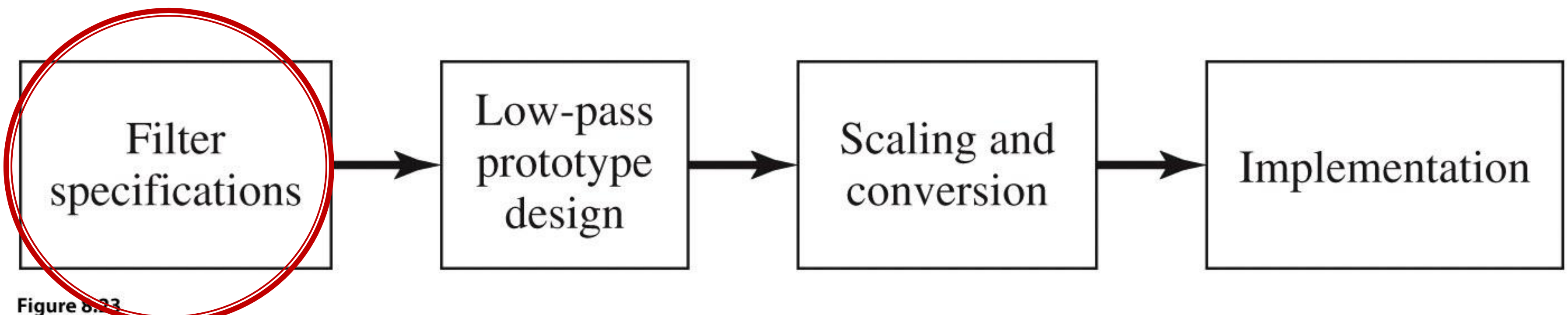
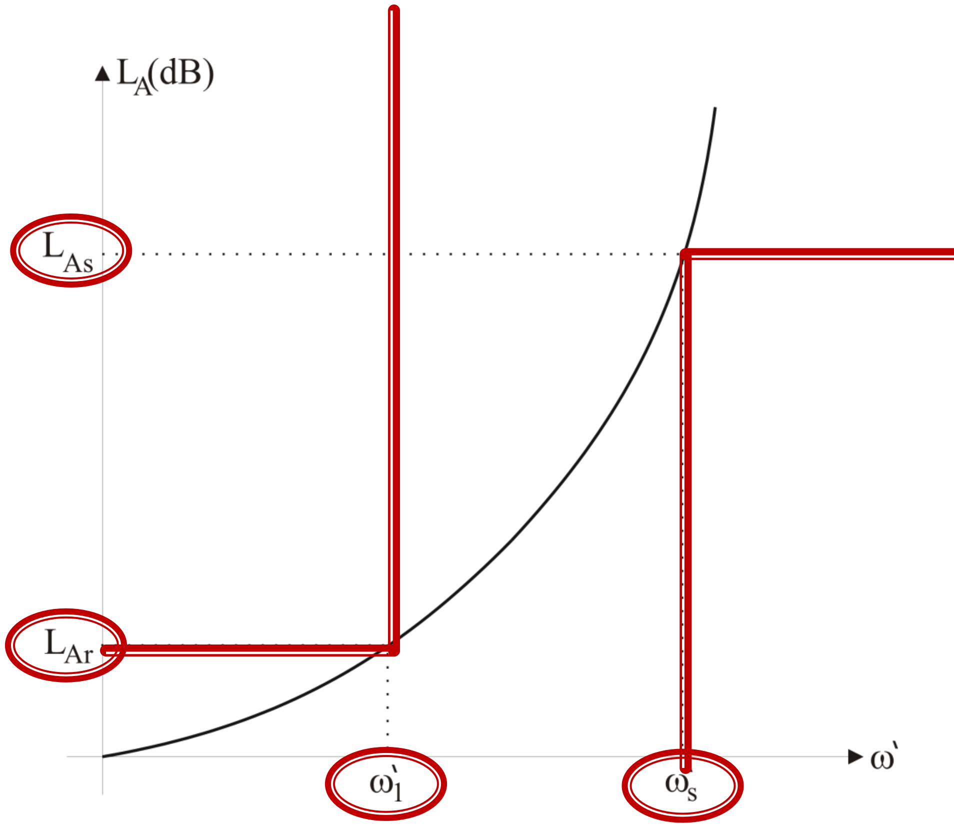


Figure 8.23

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

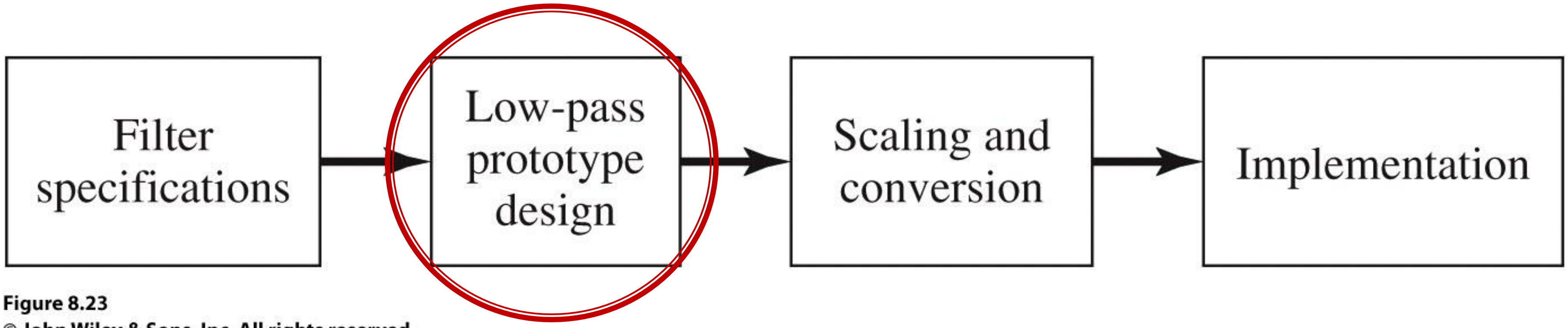
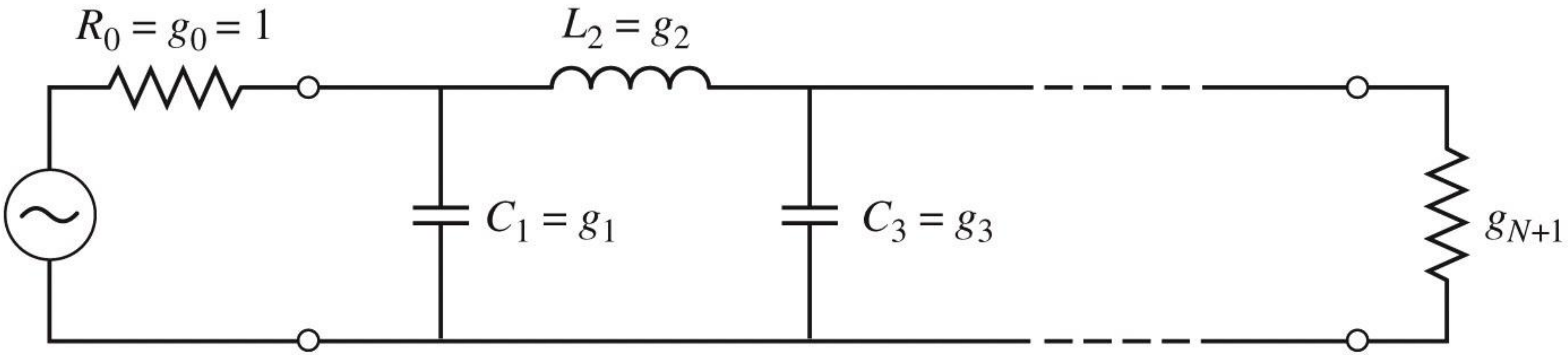
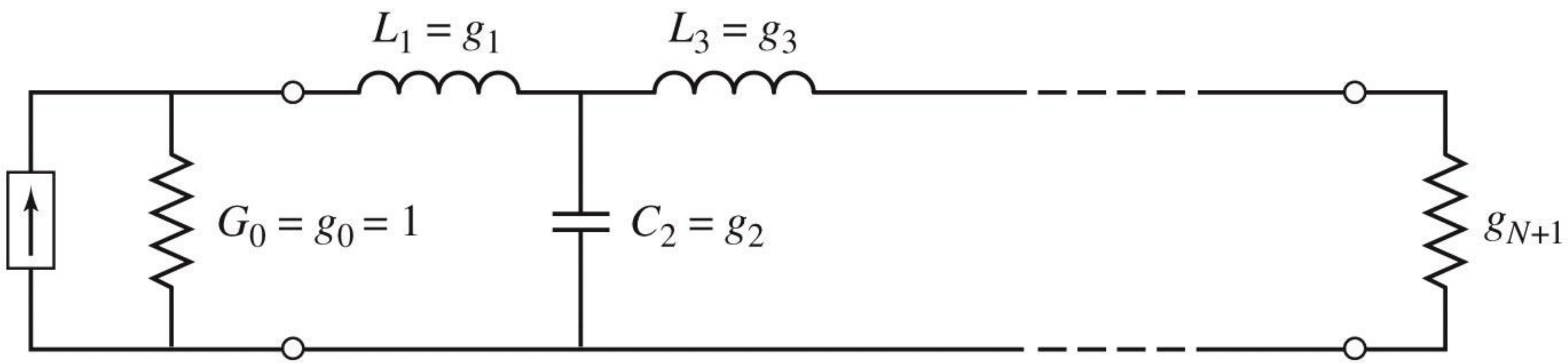


Figure 8.23

Prototype Filters



(a)



(b)

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)**

0.5 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	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

- After computing prototype filter's elements:
  - Low-Pass Filters (**LPF**)
  - cutoff frequency **ω_o = 1 rad/s** (f_o = 0.159 Hz)
  - connected to a source with **R = 1Ω**
- component values can be scaled in terms of impedance and frequency

Impedance and Frequency Scaling

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

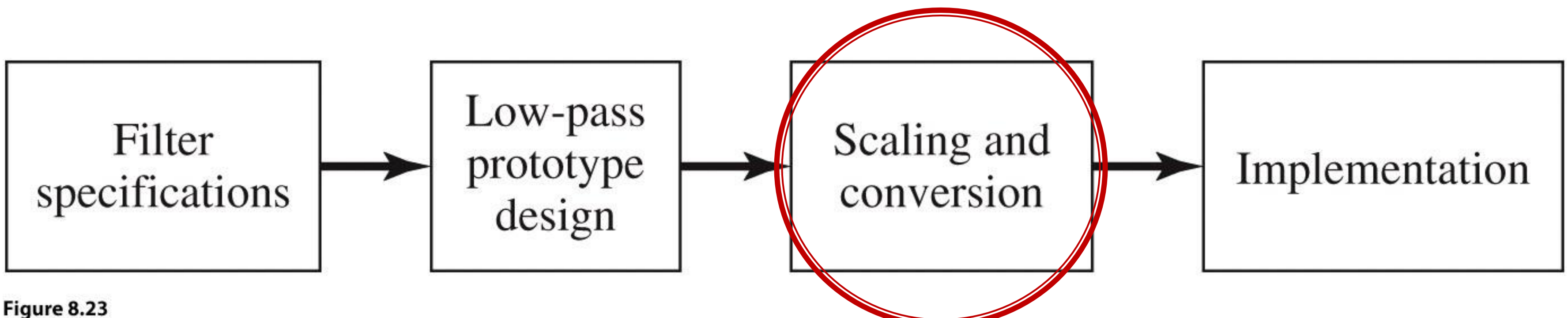


Figure 8.23

Summary of Prototype Filter Transformations

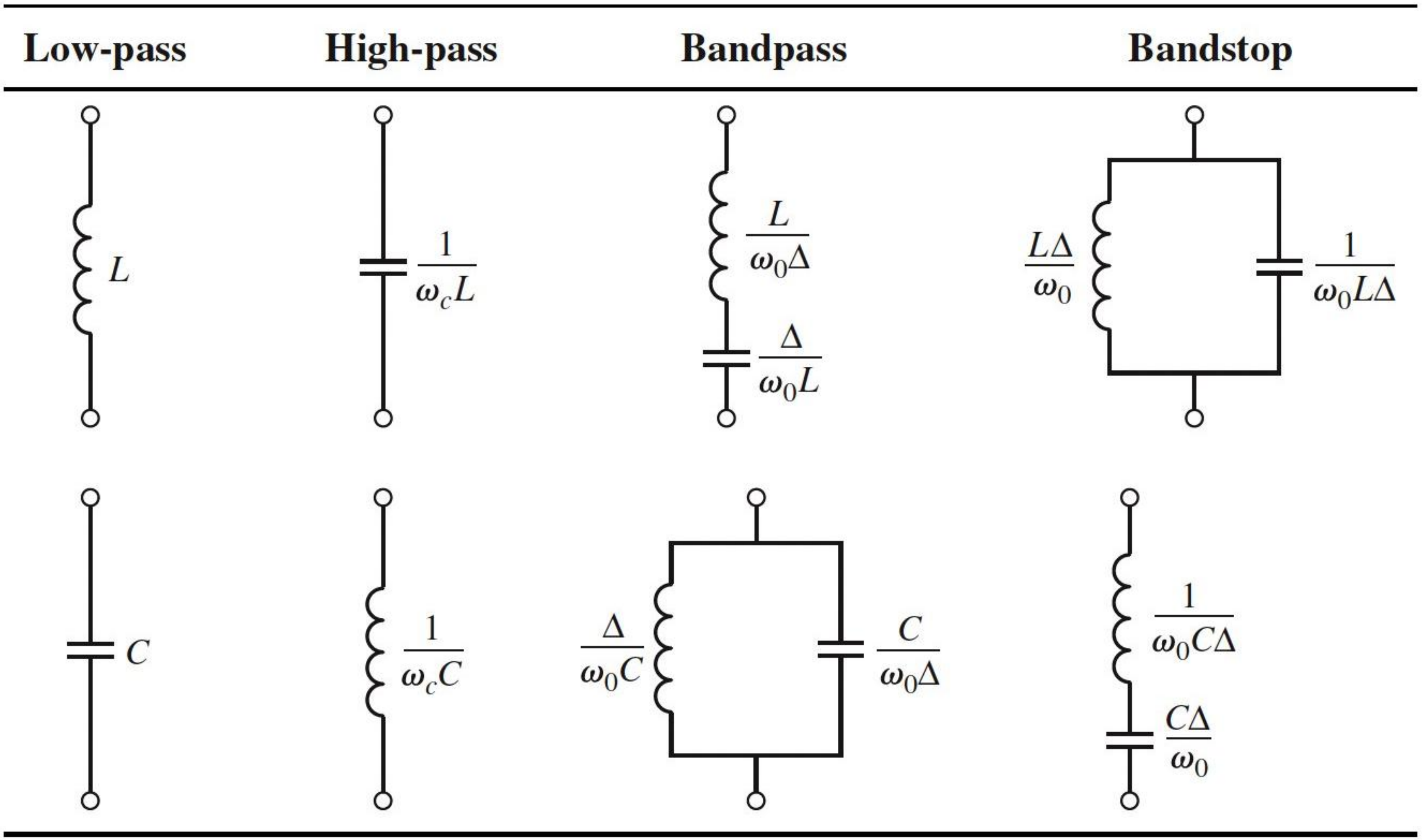


Table 8.6

Microwave Filters Implementation

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

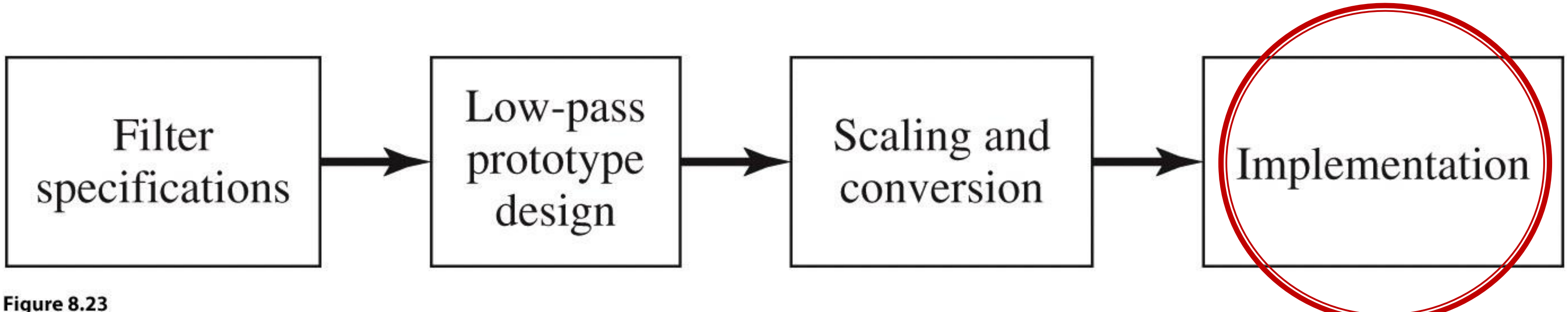
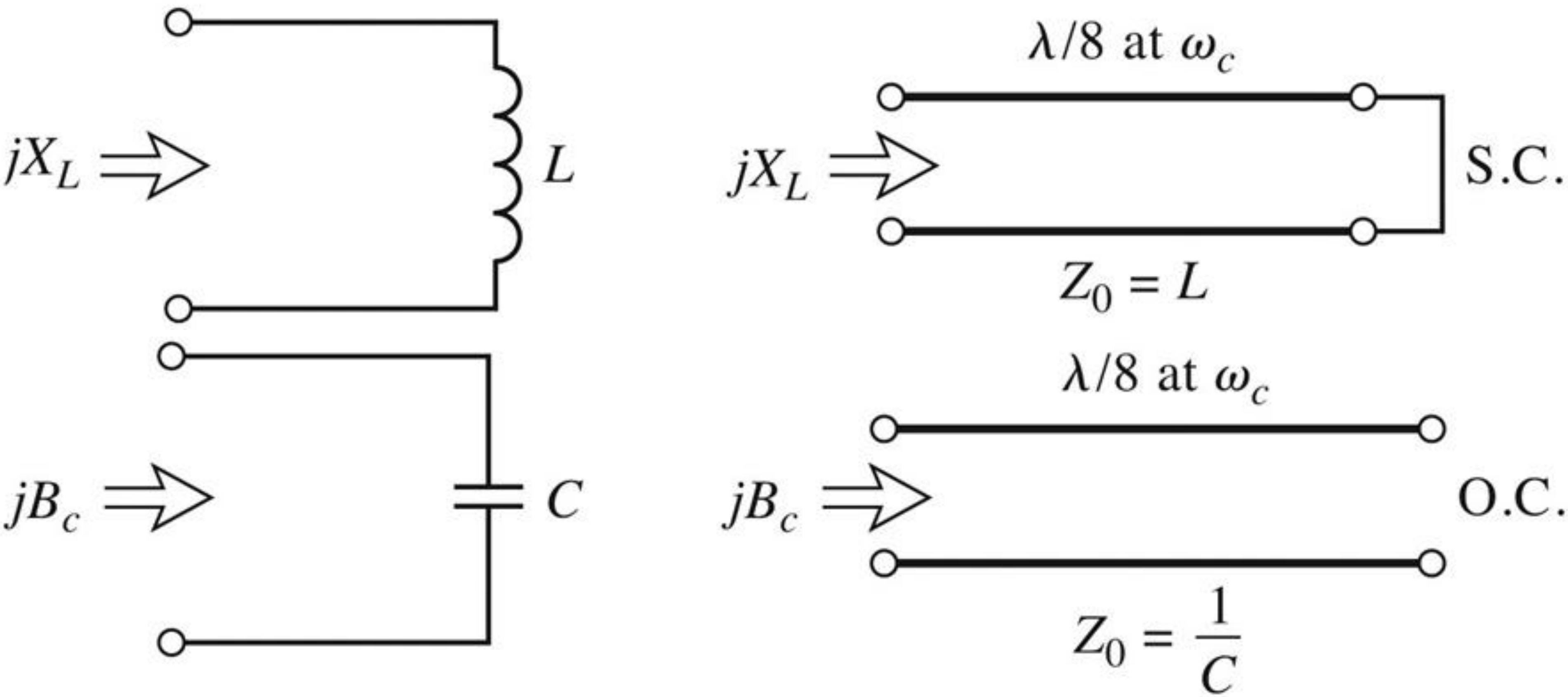


Figure 8.23

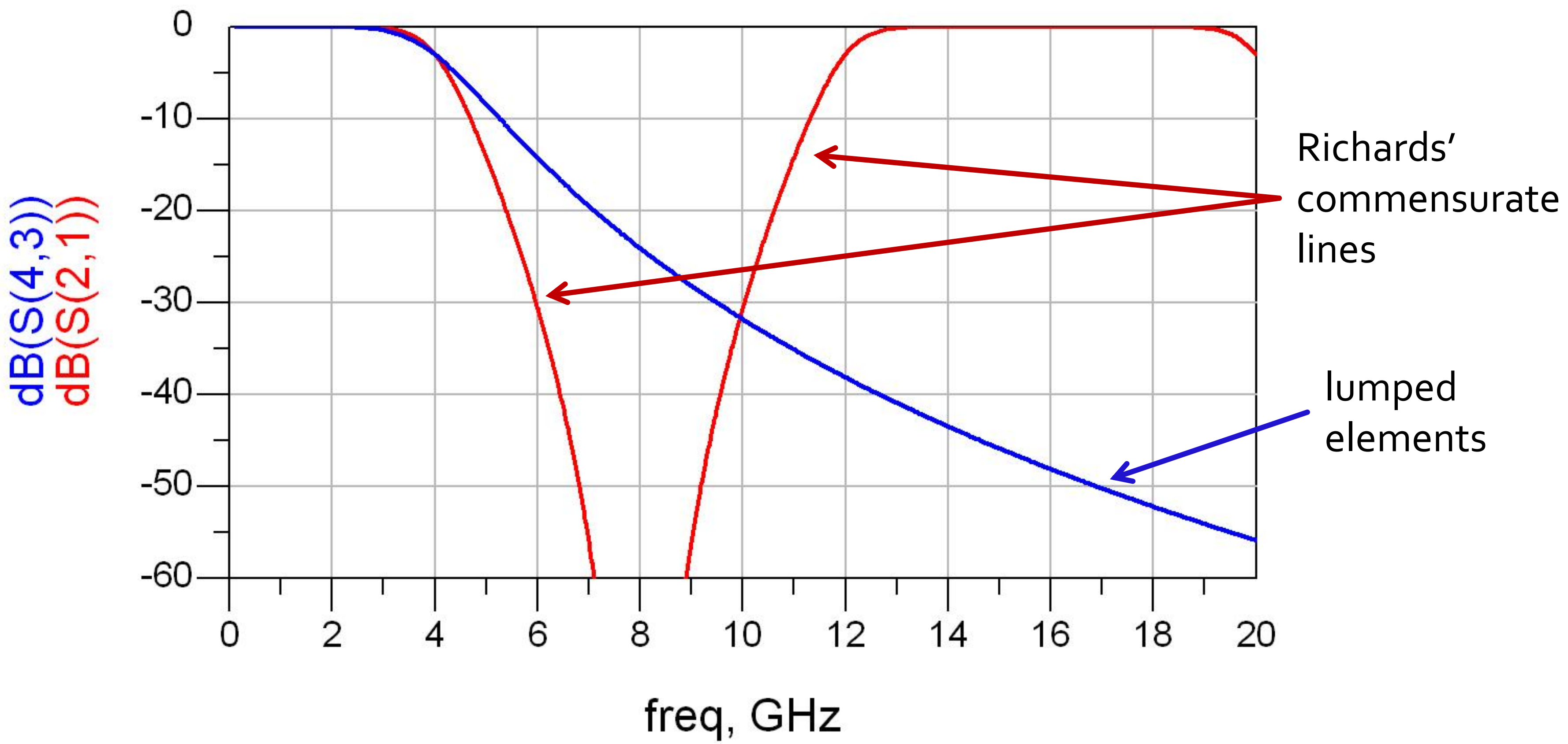
Richards' Transformation

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



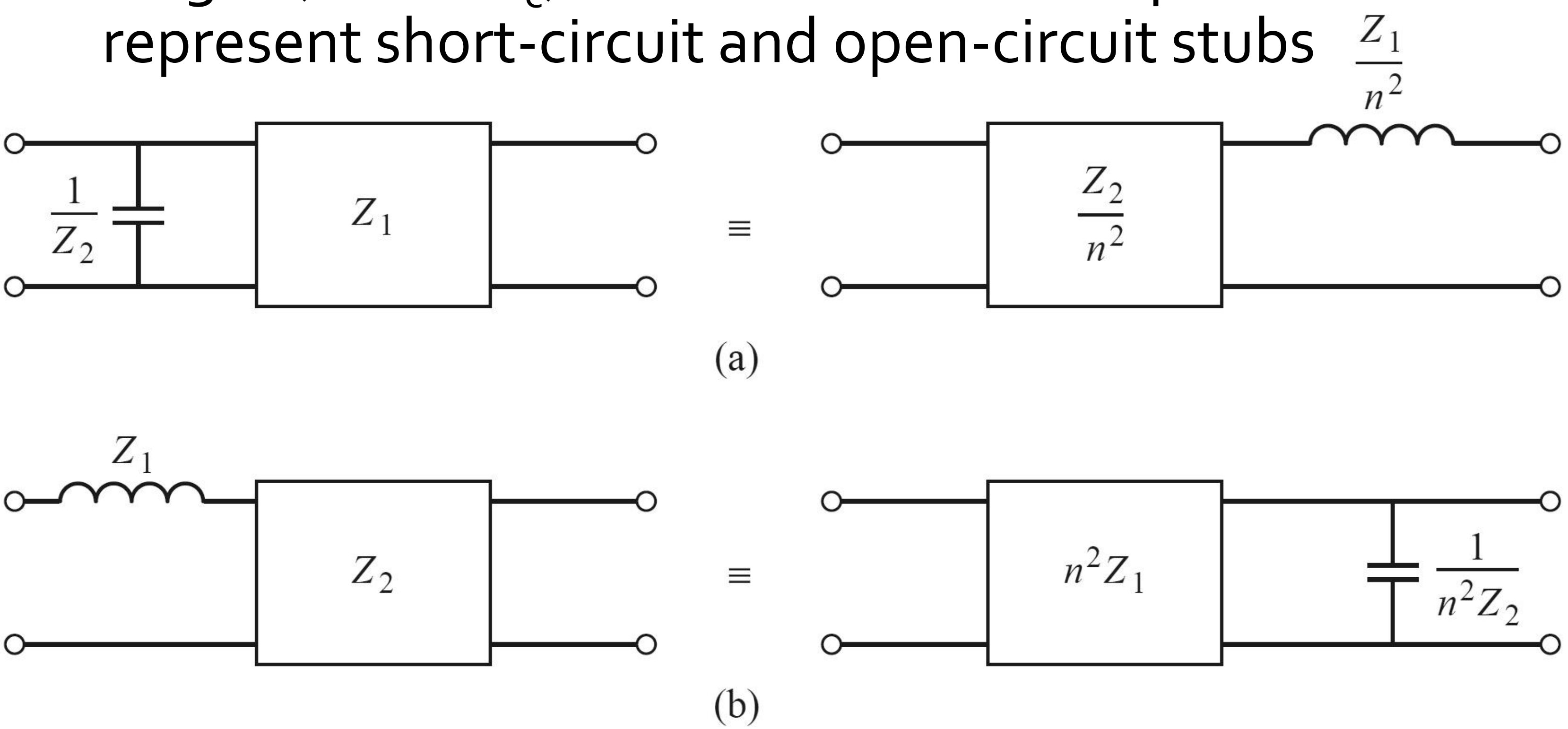
Richards' Transformation

- Filters implemented with Richards' Transformation
  - beneficiate from the supplemental pole at $2 \cdot \omega_c$
  - have the major disadvantage of frequency periodicity, a supplemental non-periodic LPF must be inserted if needed



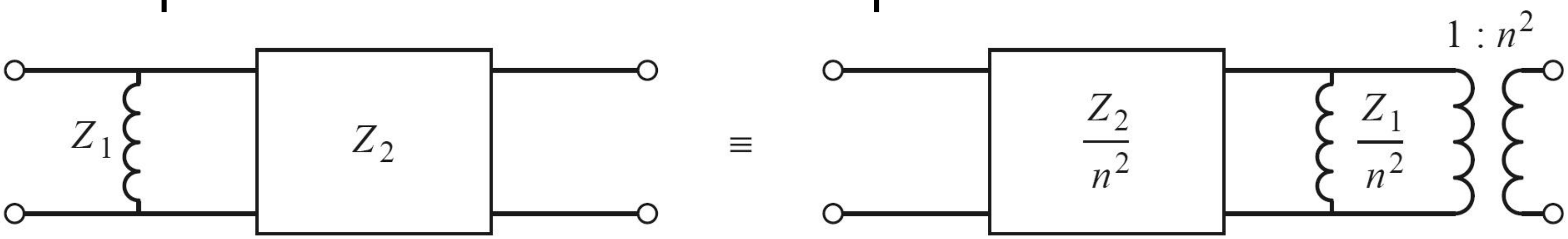
Kuroda's Identities

- 4 circuit equivalents (a,b)
  - each box represents a unit element, or transmission line, of the indicated characteristic impedance and length ($\lambda/8$ at ω_c). The inductors and capacitors represent short-circuit and open-circuit stubs

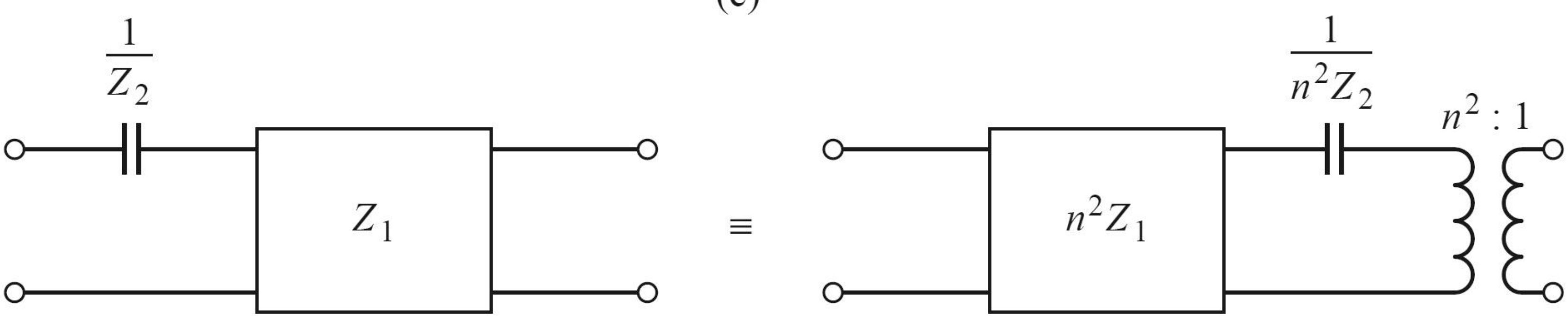


Kuroda's Identities

- 4 circuit equivalents (c,d)
  - each box represents a unit element, or transmission line, of the indicated characteristic impedance and length ($\lambda/8$ at ω_c). The inductors and capacitors represent short-circuit and open-circuit stubs



(c)



(d)

First Kuroda's Identity

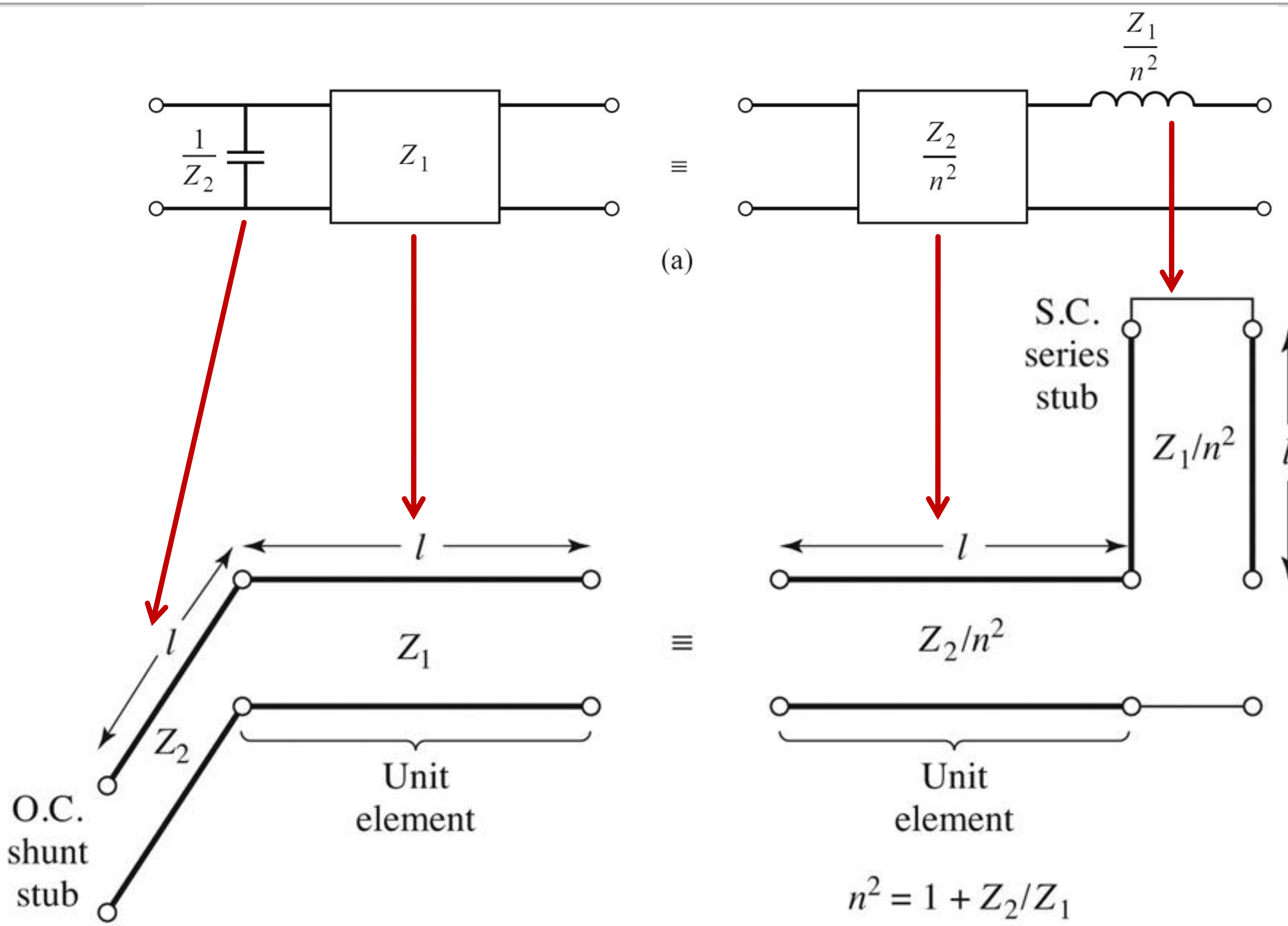
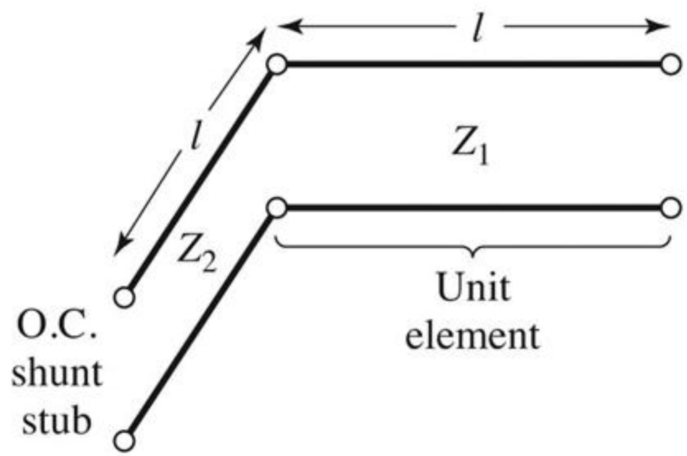
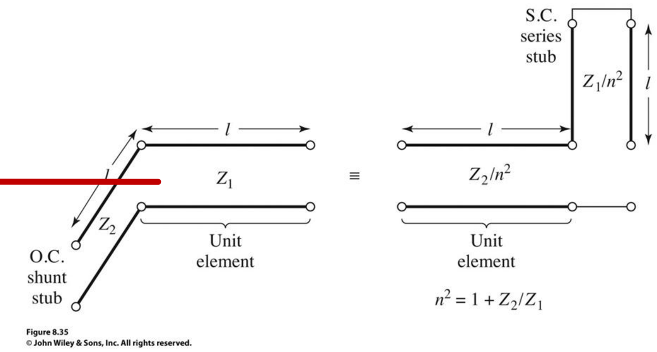


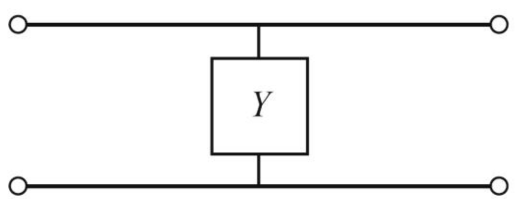
Figure 8.35

First Kuroda's Identity – Proof

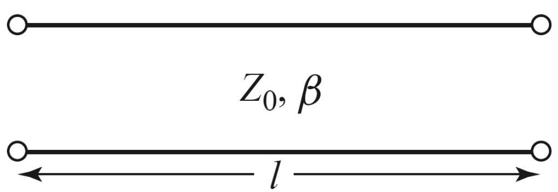




- ABCD matrices, L4



+

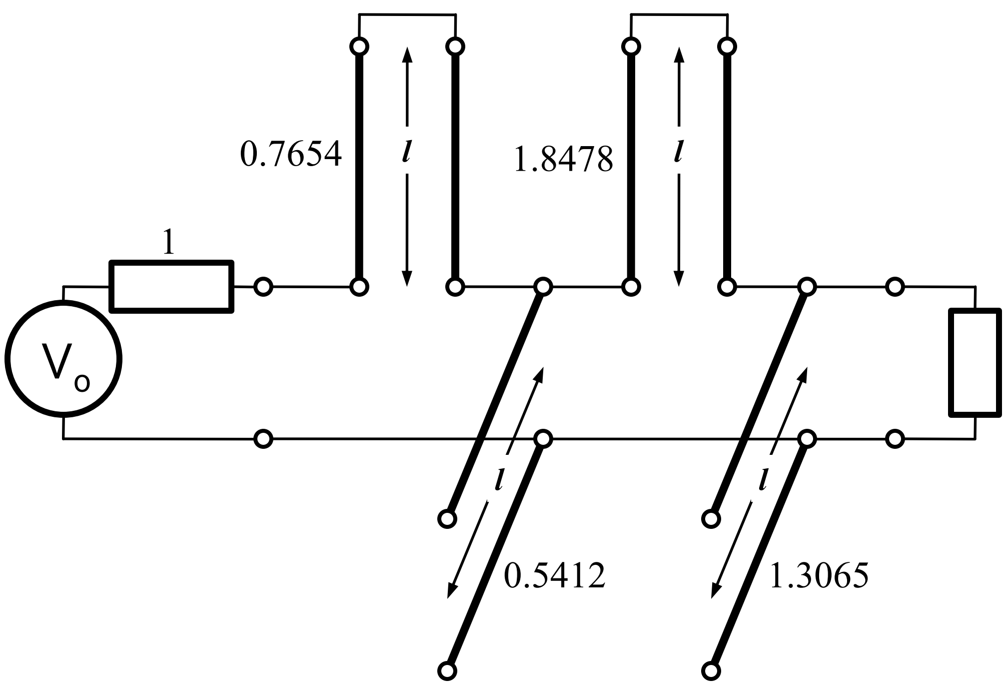


$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ Y & 1 \end{bmatrix}$$

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \cos\beta\cdot l & j\cdot Z_0\cdot\sin\beta\cdot l \\ j\cdot Y_0\cdot\sin\beta\cdot l & \cos\beta\cdot l \end{bmatrix}$$

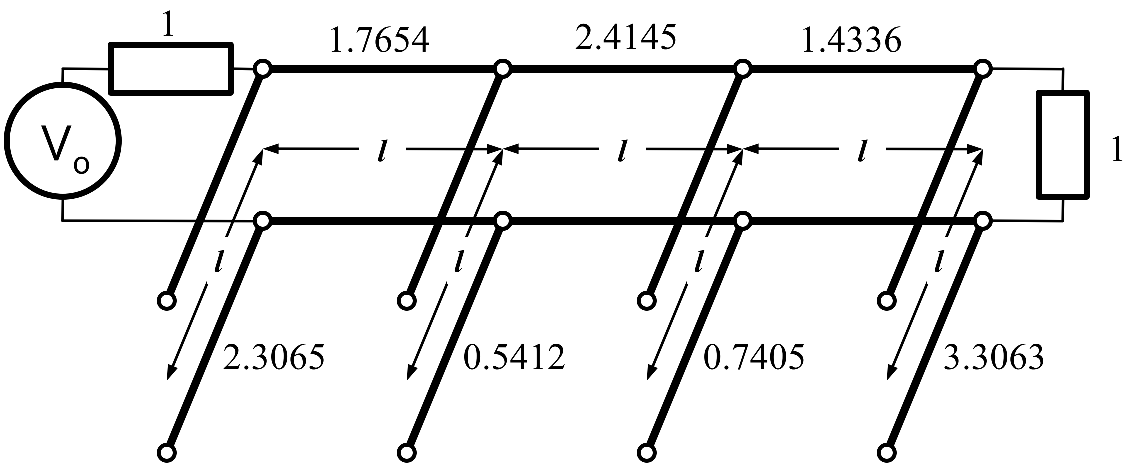
Example

- Apply Richards's transformation

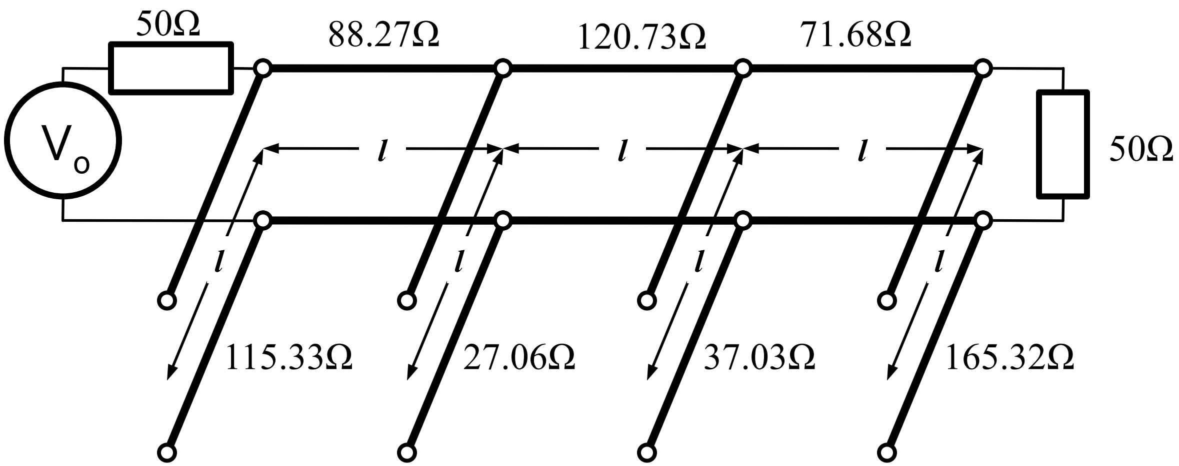


- Problems:
  - the series stubs would be **very difficult** to implement in microstrip line technology
  - in microstrip technology it is preferable to have open-circuit stubs (short-circuit requires a **via-hole** to the ground plane)
  - the 4 stubs are physically connected at the same point, an implementation that eliminates/reduces the **coupling** between these lines is impossible
  - not the case here, but sometimes the normalized impedances are much different from 1. Most circuit technologies are designed for 50Ω lines

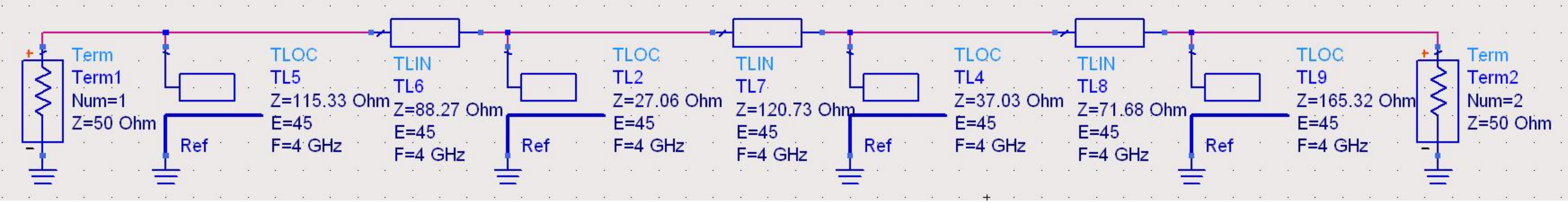
Example

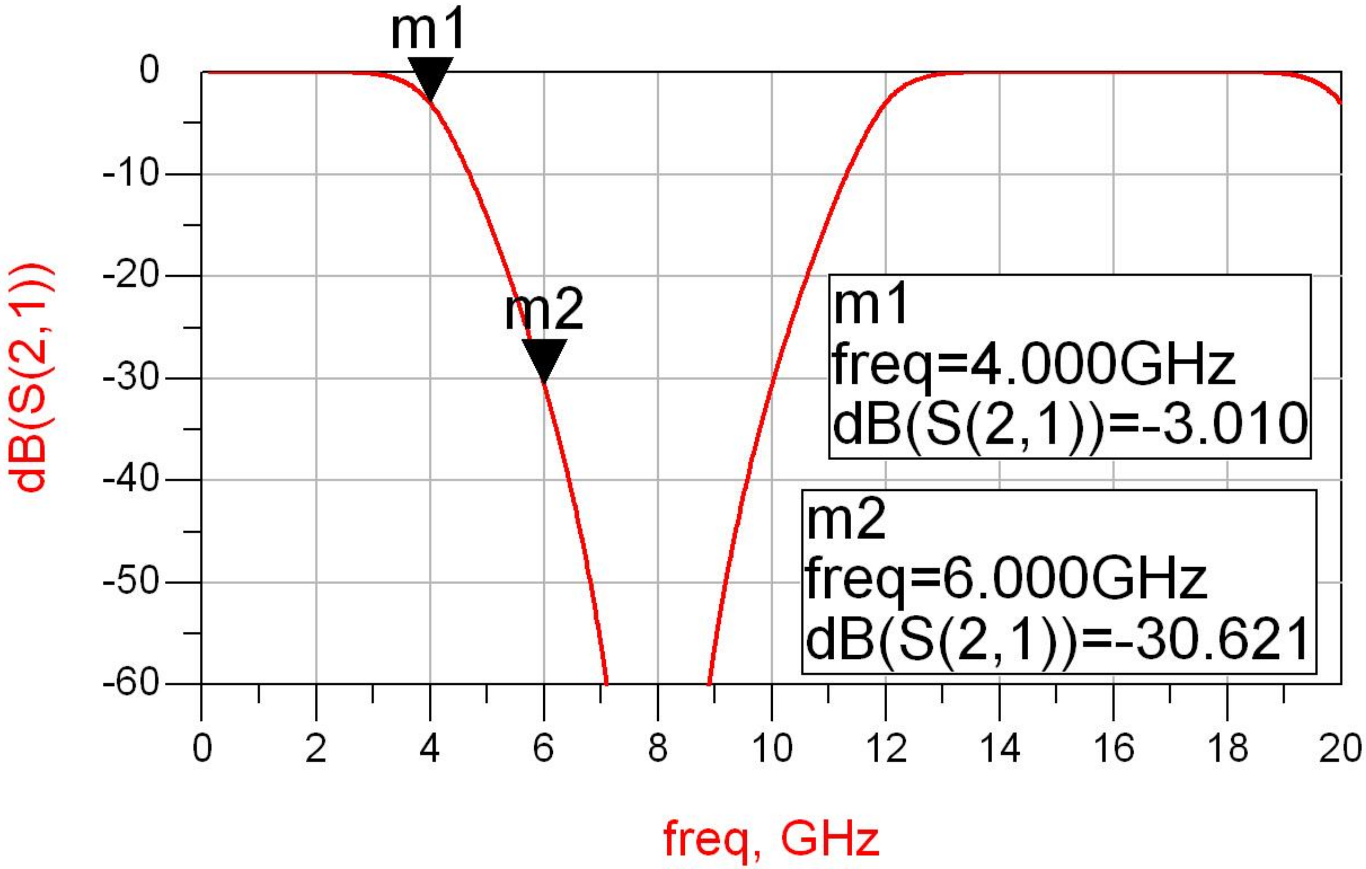


- Impedance scaling (multiply by 50Ω)



Kuroda’s Identities – ADS





Continue

Impedance and Admittance Inverters

- Richards' transformation and Kuroda's identities are useful especially for low-pass filters in technologies where the series stubs would be very difficult/ impossible to implement (microstrip)
- In the case of other filters (example 3rd order BPF):
  - series inductance can be implemented using K1-K2
  - series capacitance cannot be implemented using shunt stubs

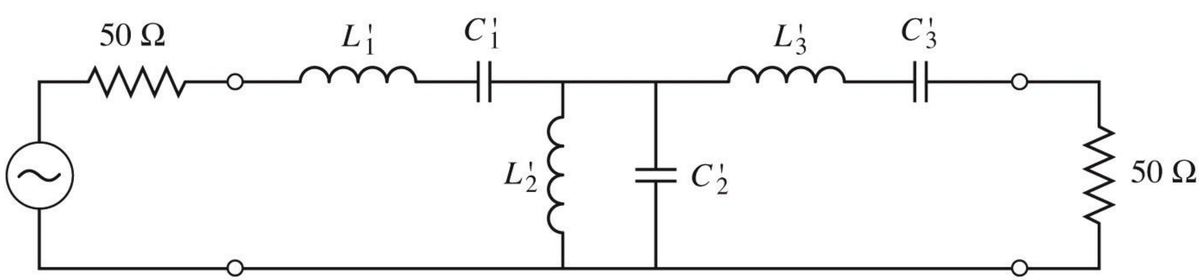


Figure 8.32

Impedance and Admittance Inverters

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

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

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

Impedance inverters

Admittance inverters

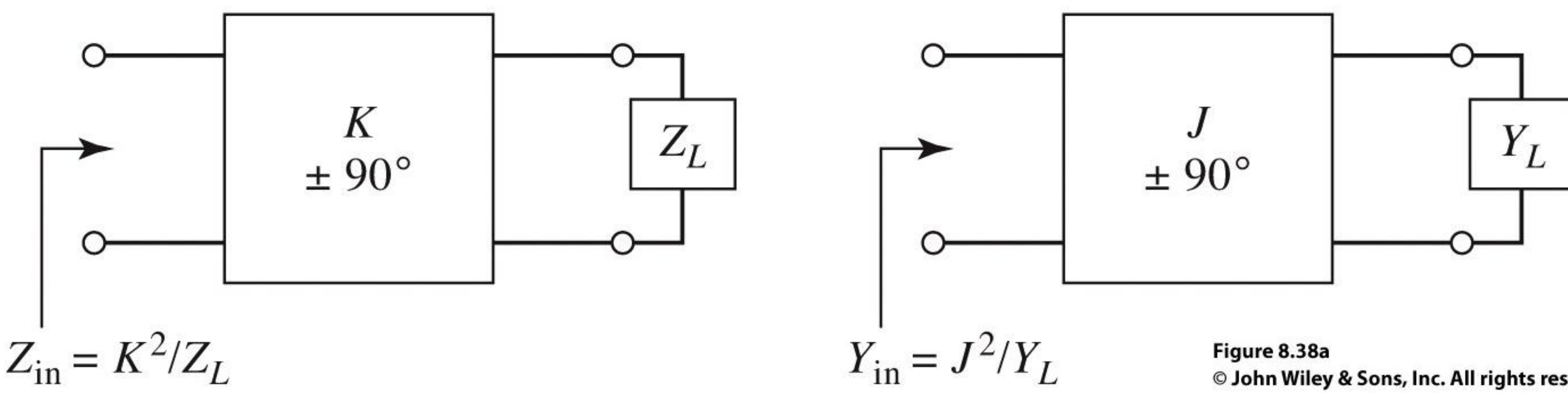
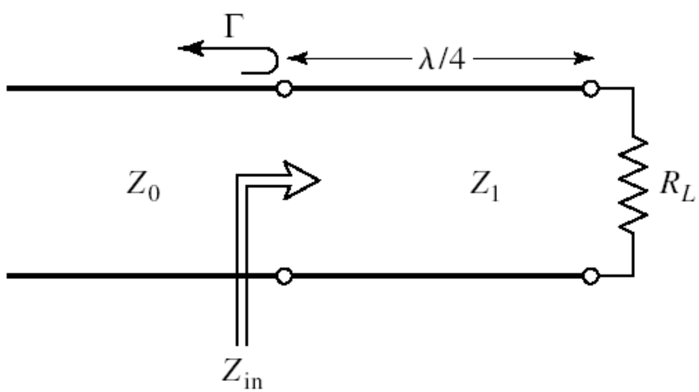


Figure 8.38a

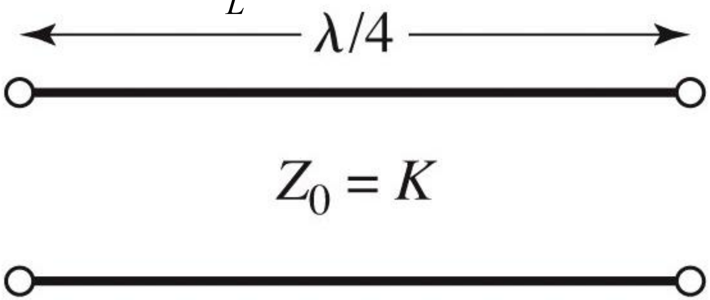
Impedance and Admittance Inverters

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



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

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



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

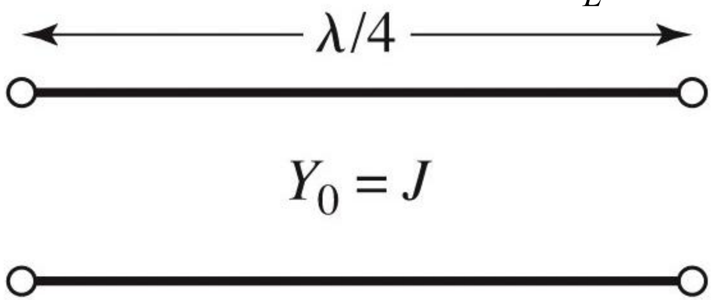
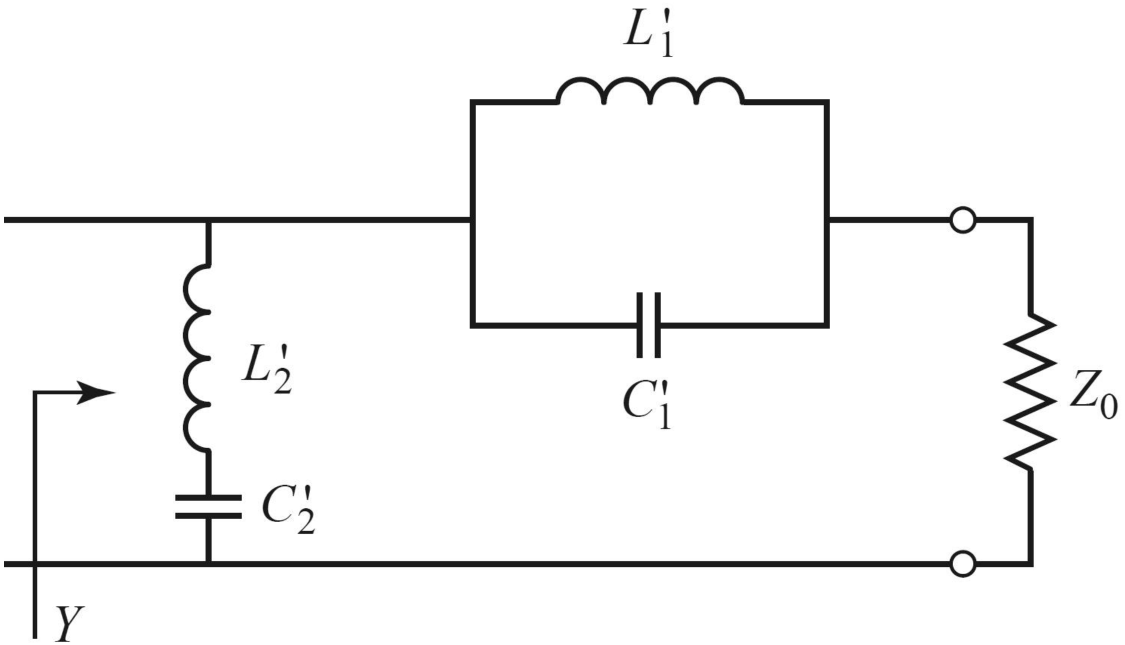


Figure 8.38b

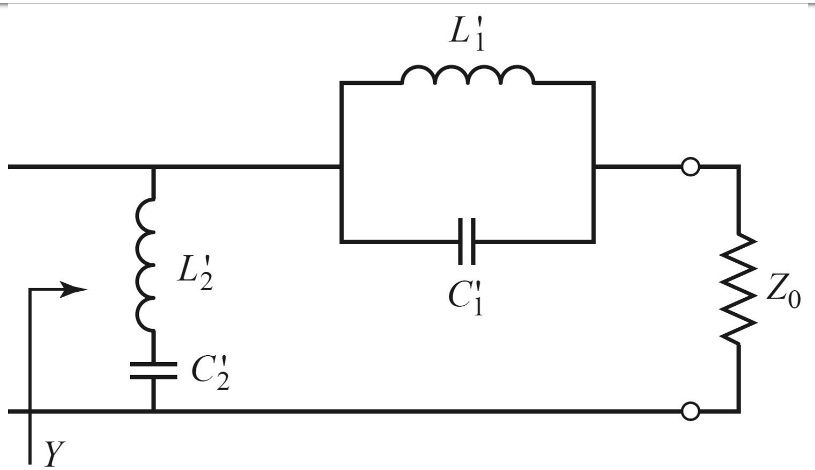
Impedance and Admittance Inverters

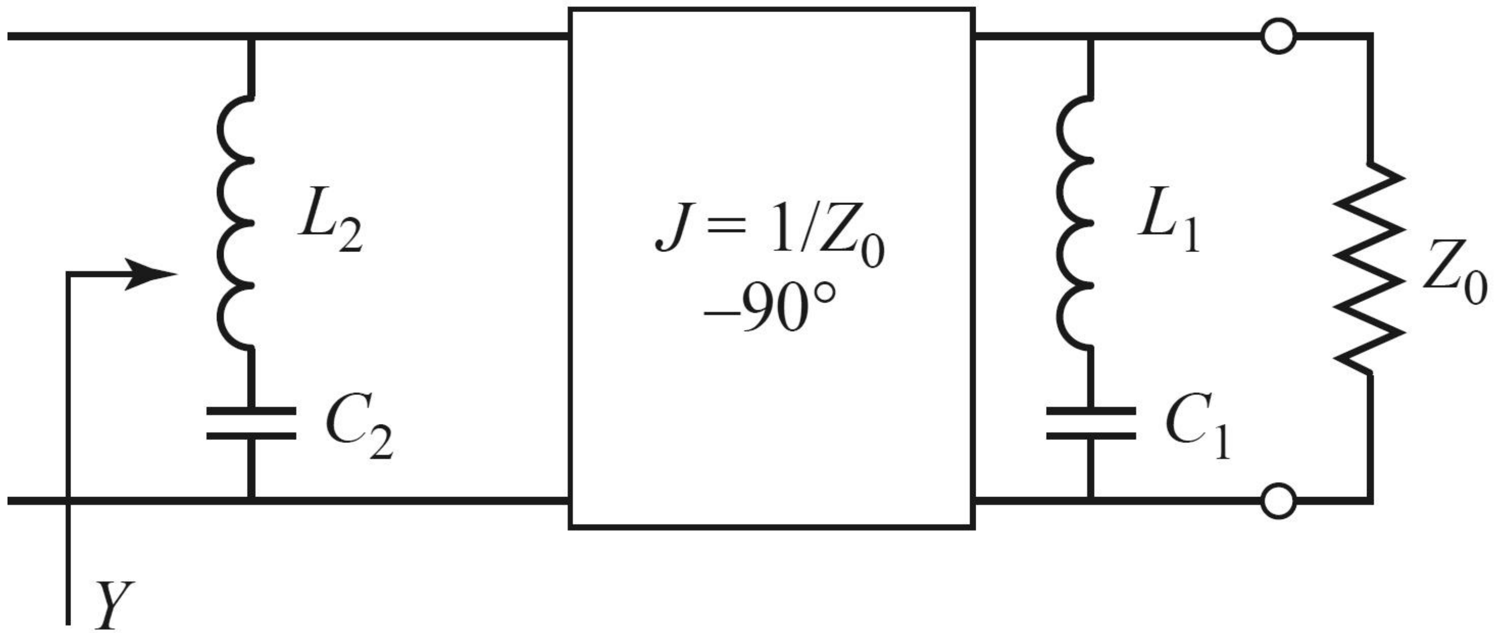
- Impedance/admittance inverters can be used to change the structure of a designed filter to a realizable form
- For example a 2nd order BSF



Impedance and Admittance Inverters

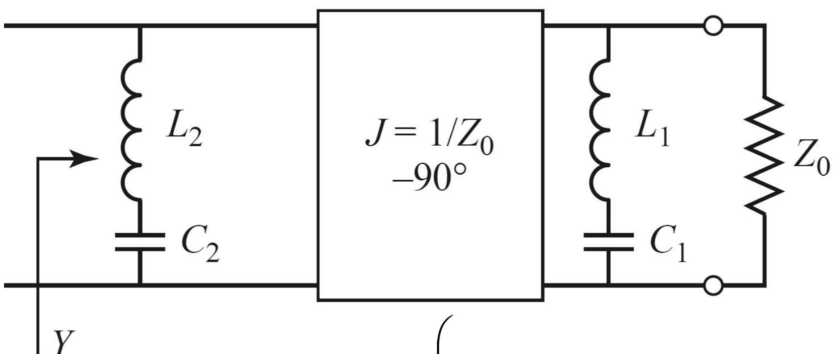
- The series elements can be eliminated/replaced using an admittance inverter

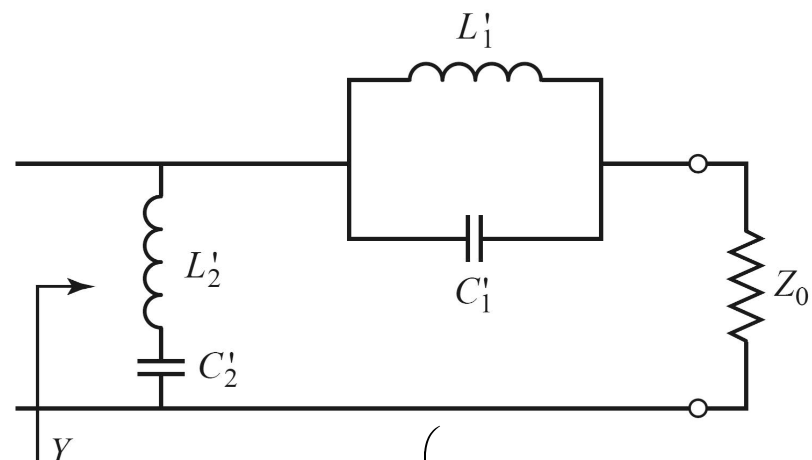




Impedance and Admittance Inverters

- The equivalence of the two schematics (when looking from the left) is proofed by obtaining the same input admittance





$$Y = \frac{1}{j\cdot\omega\cdot L_2 + \dfrac{1}{j\cdot\omega\cdot C_2}} + \frac{1}{Z_0^2}\cdot\left(\frac{1}{j\cdot\omega\cdot L_1 + \dfrac{1}{j\cdot\omega\cdot C_1}} + \frac{1}{Z_0}\right)^{-1}$$

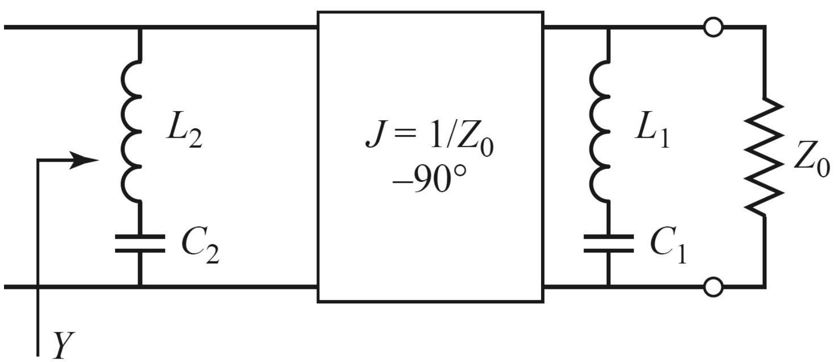
$$Y' = \frac{1}{j\cdot\omega\cdot L'_2 + \dfrac{1}{j\cdot\omega\cdot C'_2}} + \left(\frac{1}{j\cdot\omega\cdot C'_1 + \dfrac{1}{j\cdot\omega\cdot L'_1}} + Z_0\right)^{-1}$$

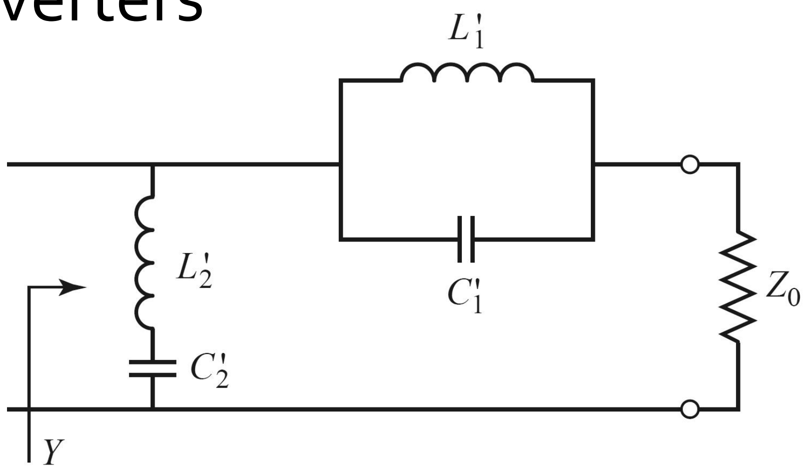
$$L_n\cdot C_n = L'_n\cdot C'_n = \frac{1}{\omega_0^2} \quad\Rightarrow\quad \begin{matrix} \dfrac{1}{Z_0^2}\cdot\sqrt{\dfrac{L_1}{C_1}} = \sqrt{\dfrac{C'_1}{L'_1}} \\ \sqrt{\dfrac{L_2}{C_2}} = \sqrt{\dfrac{L'_2}{C'_2}} \end{matrix} \quad\Rightarrow\quad Y = Y'$$

- A similar result can be obtained for a bandpass filter

Impedance and Admittance Inverters

- The complete equivalence (when looking from both sides) is obtained by enclosing the series LC circuit between two admittance inverters



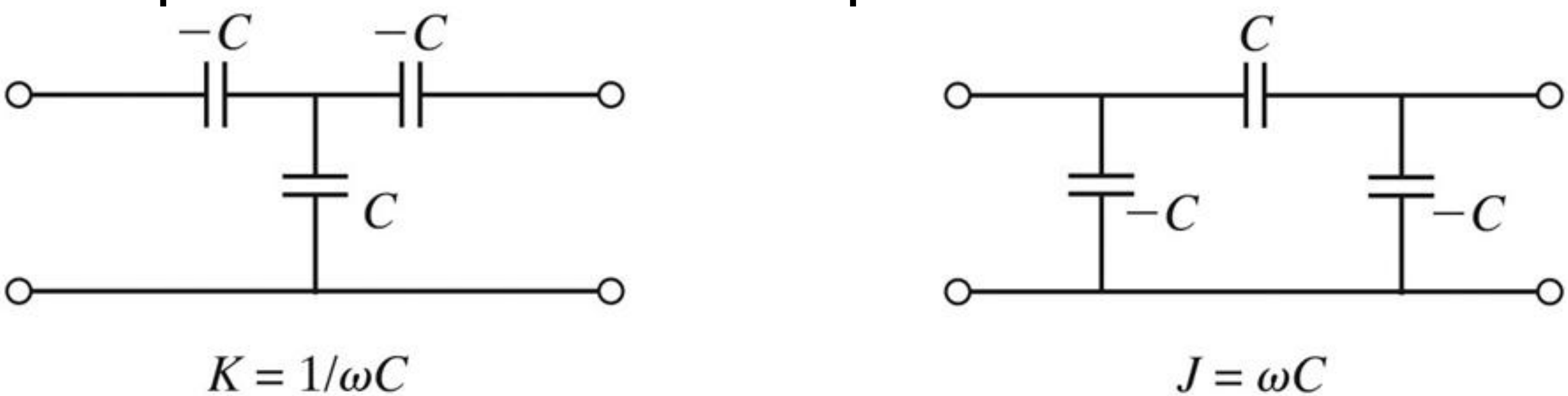


- A **series LC** circuit inserted in series in the circuit can be replaced by a **shunt LC** circuit inserted in parallel enclosed between 2 admittance inverters
- A **shunt LC** circuit inserted in series in the circuit can be replaced by a **series LC** circuit inserted in parallel enclosed between 2 admittance inverters

Practical implementations of impedance/admittance inverters

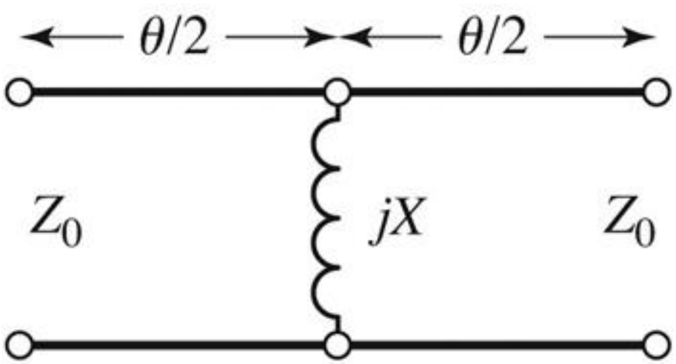
- Most often the quarter-wave transformer is used

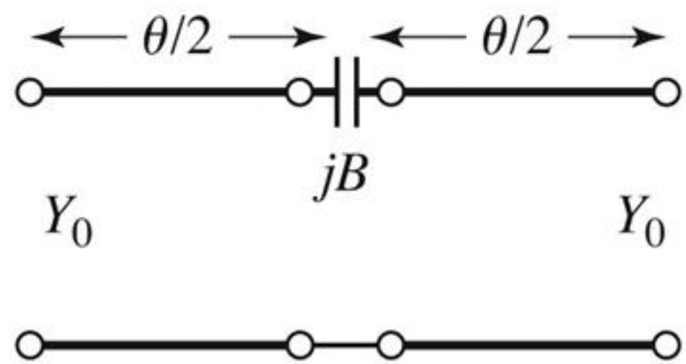
- Implementation with capacitor networks



Practical implementations of impedance/admittance inverters

- Implementation with transmission lines and reactive elements





$$K = Z_0 \cdot \tan\left|\frac{\theta}{2}\right|$$

$$X = \frac{K}{1-\left(\frac{K}{Z_0}\right)^2}$$

$$\theta = -\tan^{-1}\frac{2\cdot X}{Z_0}$$

$$\theta < 0$$

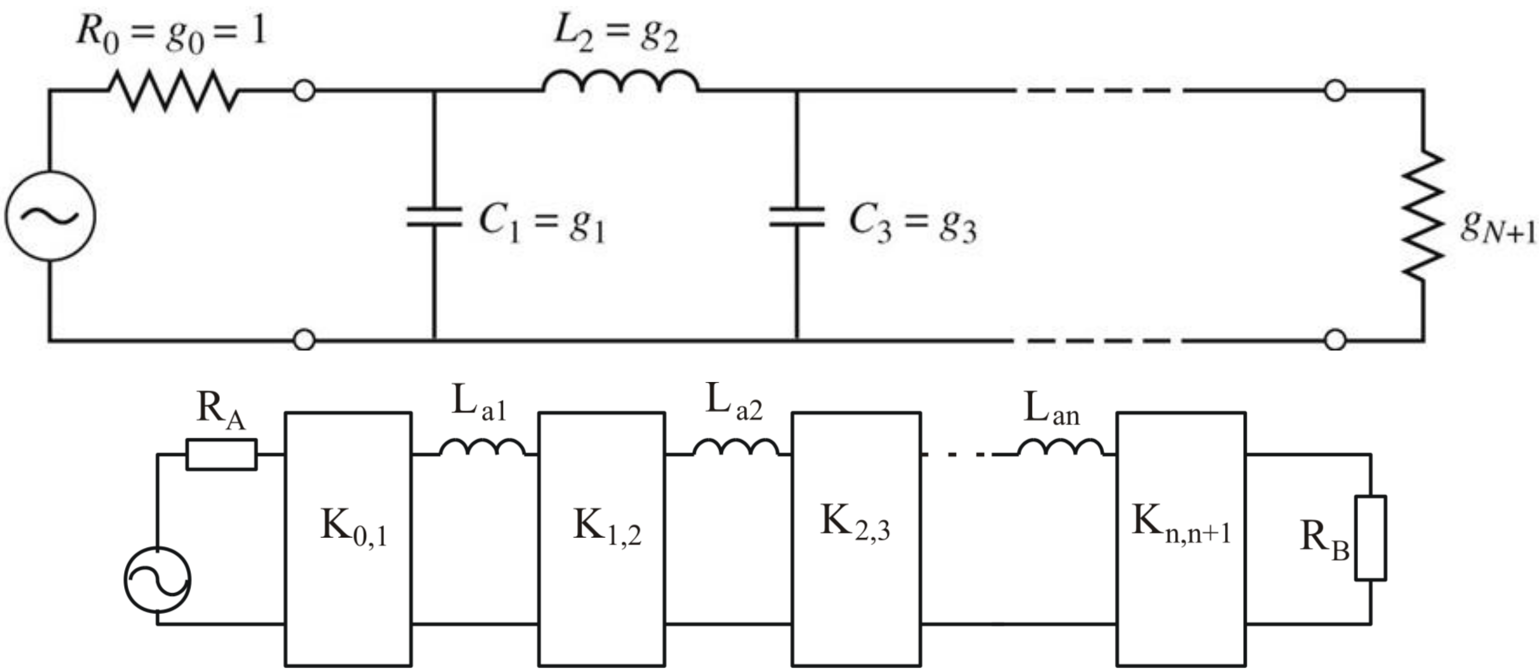
$$J = Y_0 \cdot \tan\left|\frac{\theta}{2}\right|$$

$$B = \frac{J}{1-\left(\frac{J}{Y_0}\right)^2}$$

$$\theta = -\tan^{-1}\frac{2\cdot B}{Y_0}$$

Prototype filters using inverters

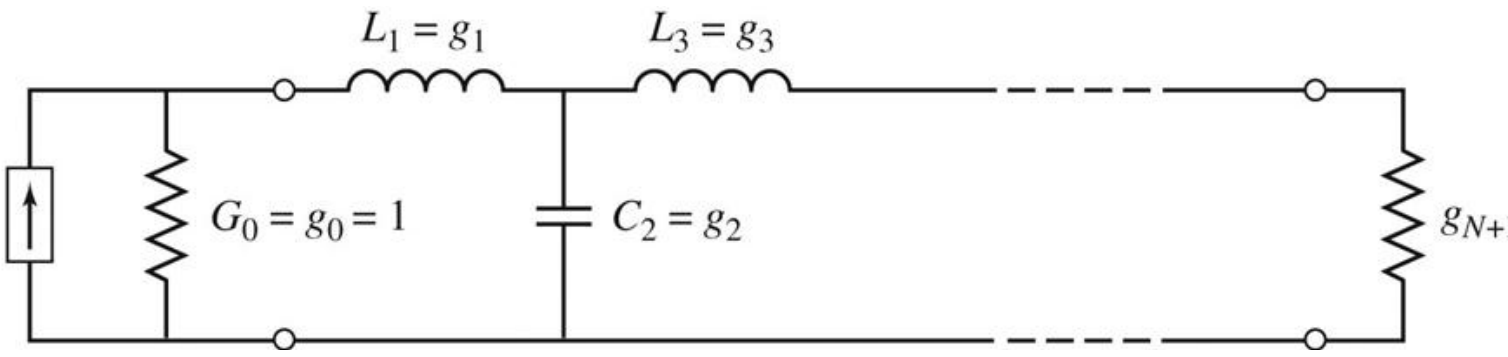
- Using impedance/admittance inverters we can implement prototype filters using a single type of reactive elements
  - Shunt C replaced by series L enclosed between 2 inverters

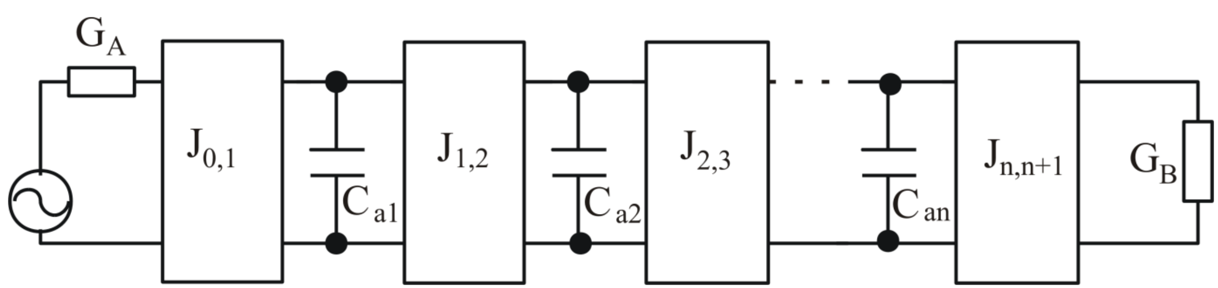


$$K_{0,1} = \sqrt{\frac{R_A \cdot L_{a,1}}{g_0 \cdot g_1}} \qquad K_{k,k+1}\Big|_{k=\overline{1,n-1}} = \sqrt{\frac{L_{a,k} \cdot L_{a,k+1}}{g_k \cdot g_{k+1}}} \qquad K_{n,n+1} = \sqrt{\frac{L_{a,n} \cdot R_B}{g_n \cdot g_{n+1}}}$$

Prototype filters using inverters

- Using impedance/admittance inverters we can implement prototype filters using a single type of reactive elements
  - Series L replaced by shunt C enclosed between 2 inverters





$$J_{0,1} = \sqrt{\frac{G_A \cdot C_{a,1}}{g_0 \cdot g_1}} \qquad J_{k,k+1}\Big|_{k=\overline{1,n-1}} = \sqrt{\frac{C_{a,k} \cdot C_{a,k+1}}{g_k \cdot g_{k+1}}} \qquad J_{n,n+1} = \sqrt{\frac{C_{a,n} \cdot g_B}{g_n \cdot g_{n+1}}}$$

Prototype filters using inverters

- For prototype filters using inverters formulas we have $2 \cdot N+1$ parameters and $N+1$ equations (to ensure the equivalence of the 2 schematics) so N parameters can be chosen freely
  - convenient values for the reactance can be chosen, and the required inverters will be computed from the equivalence equations or,
  - convenient inverters can be chosen, and the required reactance values will be computed from the equivalence equations

BPF and BSF using inverters

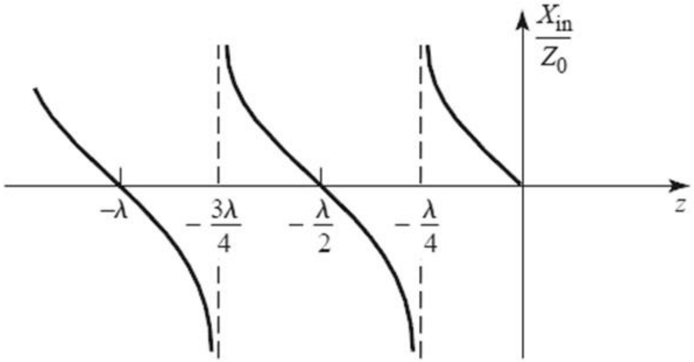
- The same principle can be applied to the BPF and BSF filters, those can be implemented using N+1 inverters and N resonators (**series or shunt LC circuits** with resonant frequency ω_o) connected either in series or in parallel enclosed between 2 inverters
  - BPF are implemented with
    - series LC circuits connected in series between inverters
    - shunt LC circuits connected in parallel between inverters
  - BSF are implemented with
    - shunt LC circuits connected in series between inverters
    - series LC circuits connected in parallel between inverters

Lines as resonators

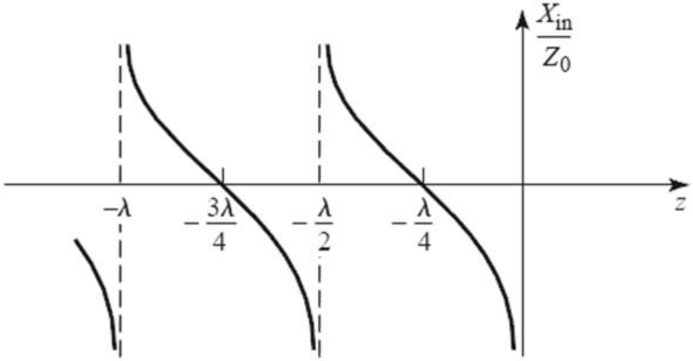
- The impedance of short-circuited or open-circuited line (stub) shows a resonant behavior that can be used to implement required resonators

$$Z_{in} = Z_0 \cdot \frac{Z_L + j \cdot Z_0 \cdot \tan \beta \cdot l}{Z_0 + j \cdot Z_L \cdot \tan \beta \cdot l}$$

$$Z_{in,sc} = j \cdot Z_0 \cdot \tan \beta \cdot l$$

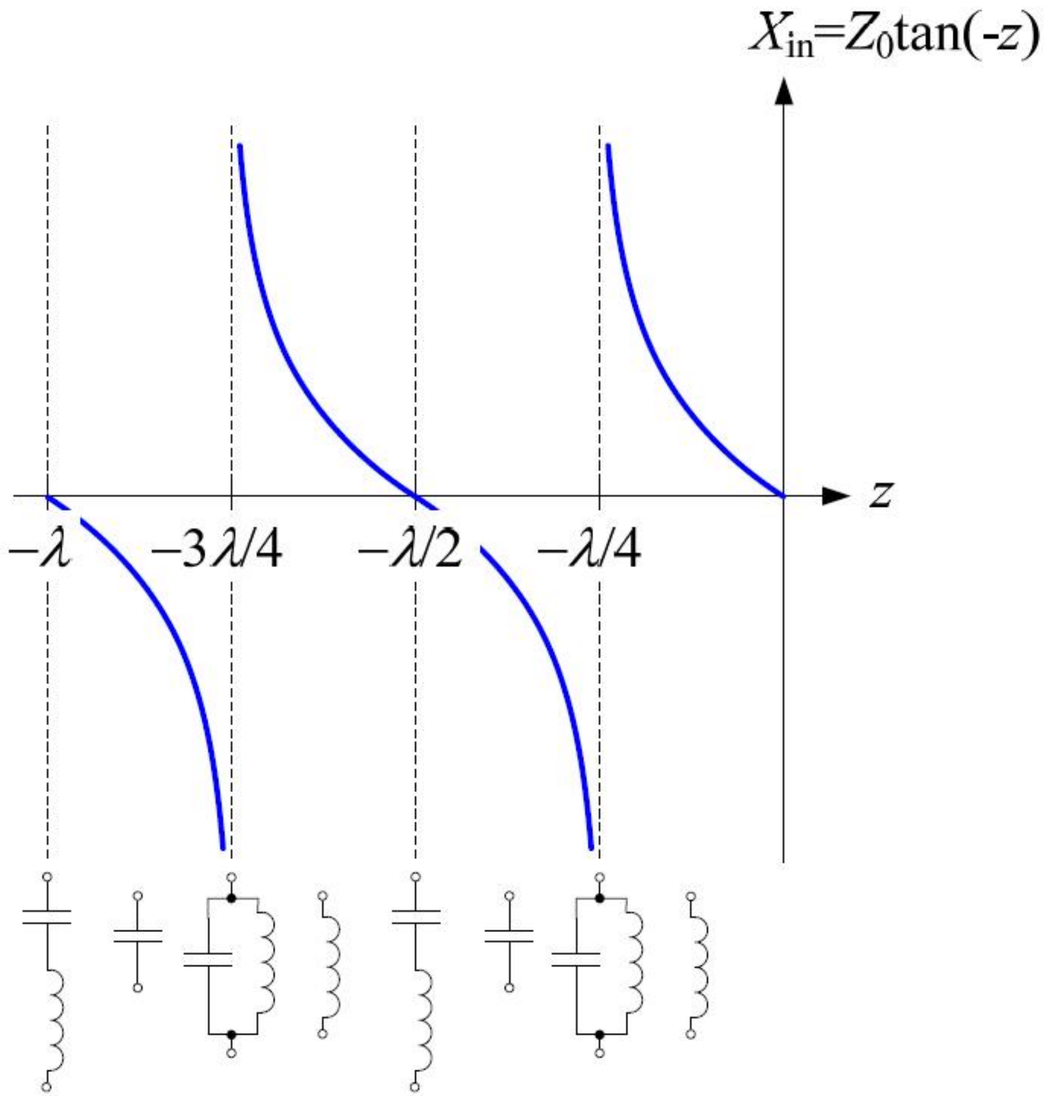


$$Z_{in,oc} = -j \cdot Z_0 \cdot \cot \beta \cdot l$$



Lines as resonators

- Short-circuited line
- For the frequency at which l = λ/4 (ω_o) the line behaves as an shunt LC resonator circuit
  - the line shows capacitive behavior for lower frequencies (l>λ/4)
  - the line shows inductive behavior for higher frequencies (l<λ/4)
- Similar discussion for the open circuited line (equivalent to a series LC resonator around the frequency at which l=λ/4)



BPF/BSP design formulas

- When the admittance inverters are implemented with quarter-wave transformers with Zo characteristic impedance
  - BPF – short-circuited shunt stubs with l = λ/4

$$Z_{0n} \approx \frac{\pi \cdot Z_0 \cdot \Delta}{4 \cdot g_n}$$

  - BSF – open-circuited shunt stubs with l = λ/4

$$Z_{0n} \approx \frac{4 \cdot Z_0}{\pi \cdot g_n \cdot \Delta}$$

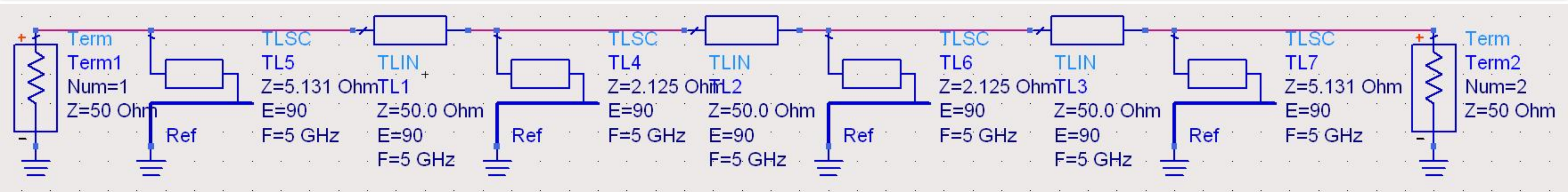
Example

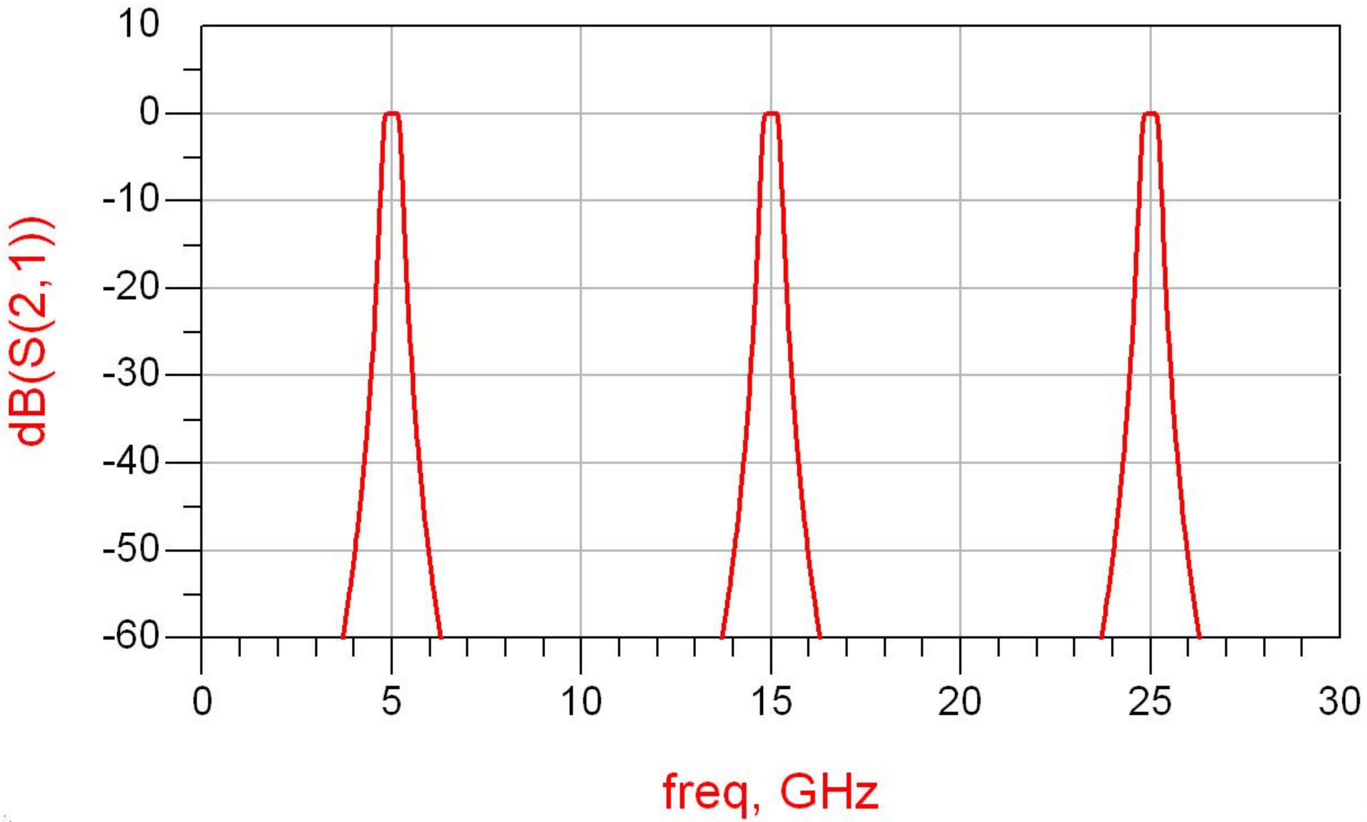
- Similar to a project assignment
- Follows the amplifier designed as in L8
- 4^{th} order bandpass filter, fo = 5GHz, fractional bandwidth of the passband 10 %
- maximally flat table or formulas for g_n:

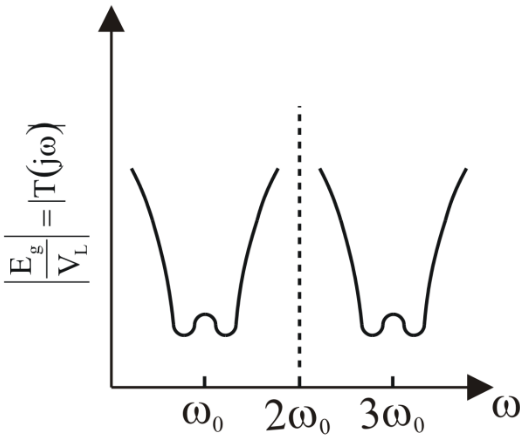
n	g_n	$Z_{on}(\Omega)$
1	0.7654	5.131
2	1.8478	2.125
3	1.8478	2.125
4	0.7654	5.131

$$Z_{0n} \approx \frac{\pi \cdot Z_0 \cdot \Delta}{4 \cdot g_n}$$

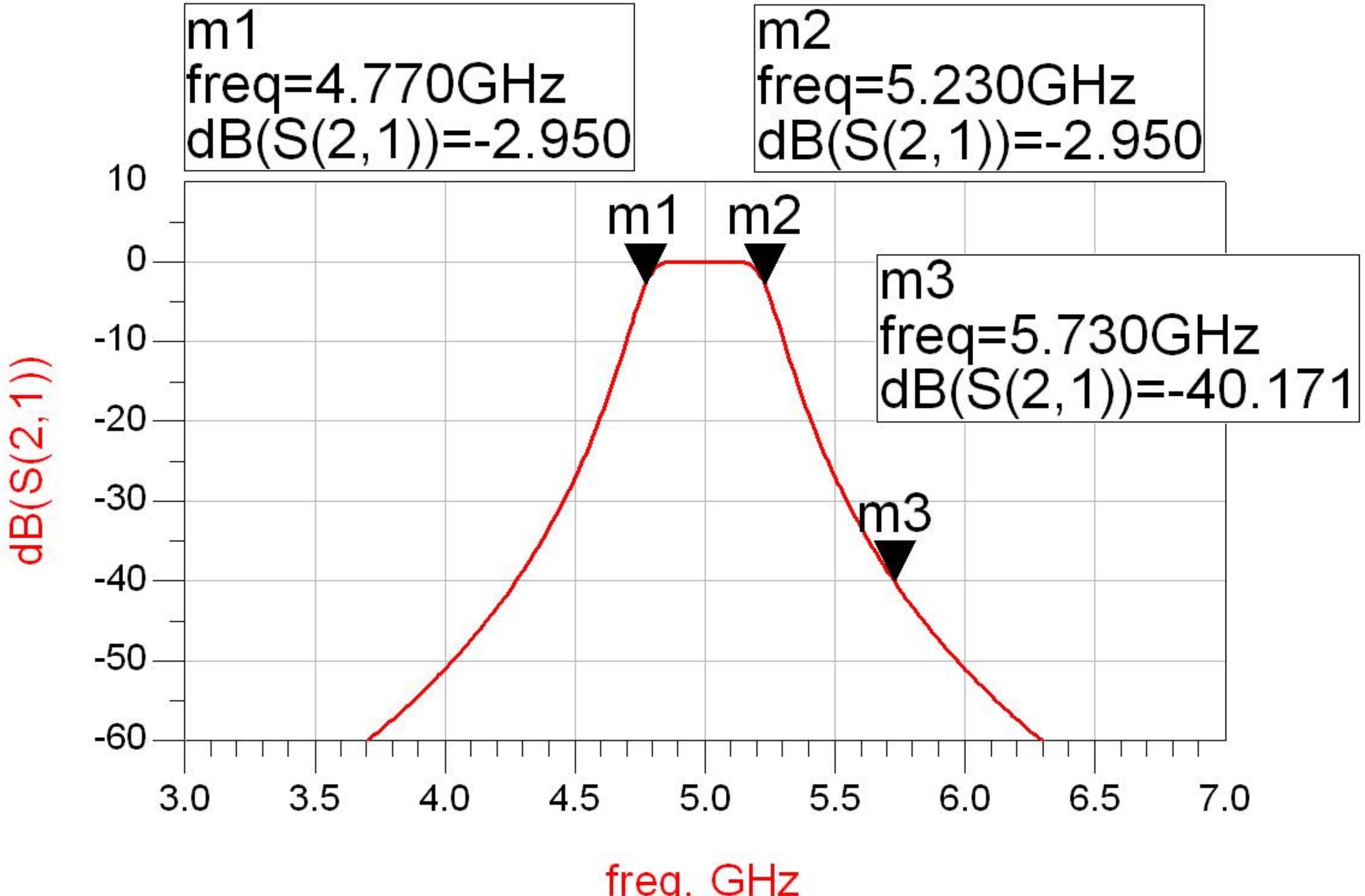
ADS – BPF


Term
Term1
Num=1
Z=50 Ohm
Ref
TLSC
TL5
Z=5.131 Ohm
E=90
F=5 GHz
TLIN
TL1
Z=50.0 Ohm
E=90
F=5 GHz
Ref
TLSC
TL4
Z=2.125 Ohm
E=90
F=5 GHz
TLIN
TL2
Z=50.0 Ohm
E=90
F=5 GHz
Ref
TLSC
TL6
Z=2.125 Ohm
E=90
F=5 GHz
TLIN
TL3
Z=50.0 Ohm
E=90
F=5 GHz
Ref
TLSC
TL7
Z=5.131 Ohm
E=90
F=5 GHz
Term
Term2
Num=2
Z=50 Ohm


dB(S(2,1))
10
0
-10
-20
-30
-40
-50
-60
0
5
10
15
20
25
30
freq, GHz


$\left|\frac{E_g}{V_L}\right| = \left|T(j\omega)\right|$
ω_0
$2\omega_0$
$3\omega_0$
ω

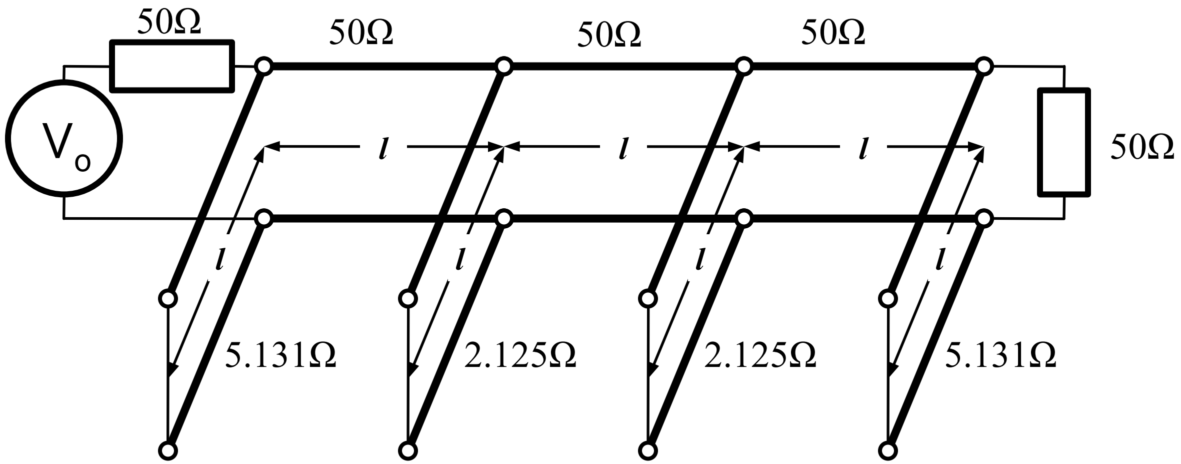
ADS – BPF



m1
freq=4.770GHz
dB(S(2,1))=-2.950
m2
freq=5.230GHz
dB(S(2,1))=-2.950
m3
freq=5.730GHz
dB(S(2,1))=-40.171
10
0
-10
-20
-30
-40
-50
-60
dB(S(2,1))
3.0
3.5
4.0
4.5
5.0
5.5
6.0
6.5
7.0
freq, GHz
m1
m2
m3

Example

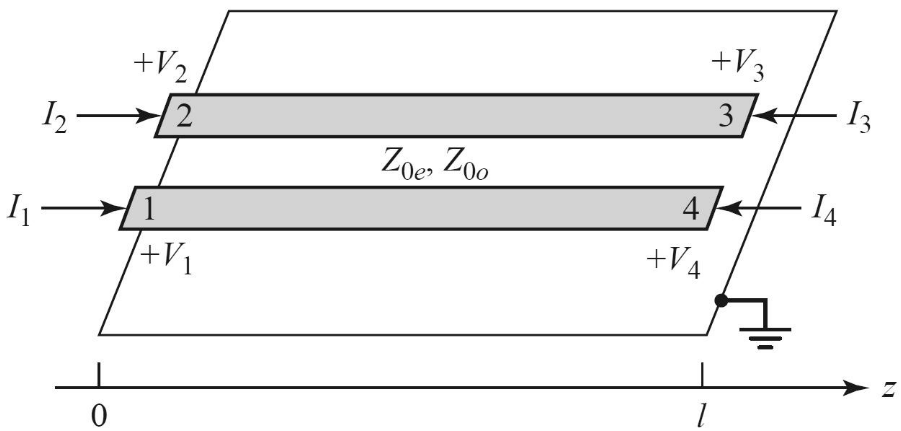
$$l = \frac{\lambda}{4} \quad \Rightarrow \quad \beta \cdot l = \frac{\pi}{2}$$

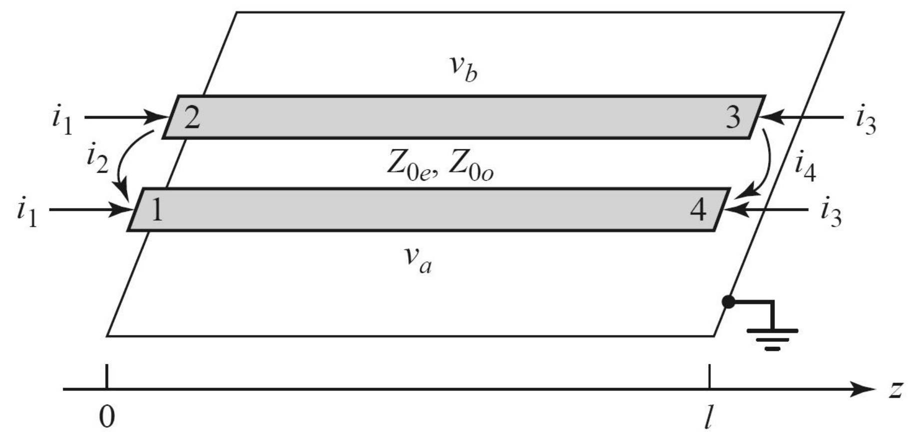


- Disadvantages of the filters using impedance inverters and lines as resonators:
  - short-circuited stubs (via-hole) for BPF
  - often the **characteristic impedances** for the stubs have **values** difficult to implement (2.125Ω)

Coupled Line Filters

- A parallel coupled line section model is obtained by even/odd mode analysis
- Even and odd modes are characterized by the characteristic even/odd mode impedances whose required values will impose the lines' geometry (width / distance between lines, depending on the line technology we use)





Coupled Lines

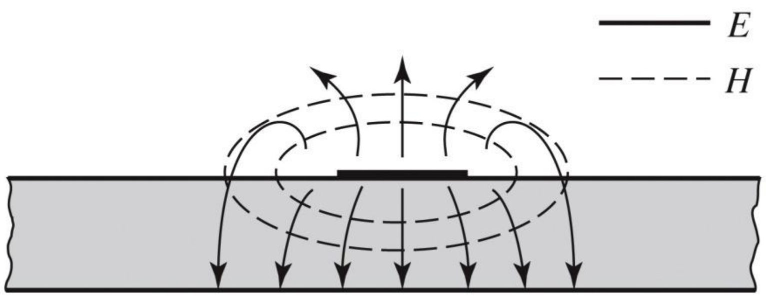
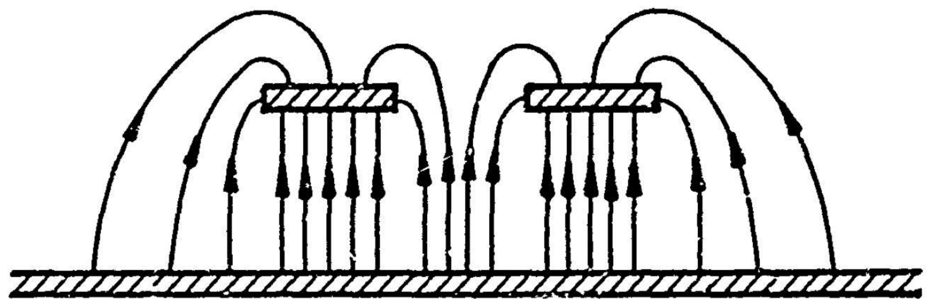
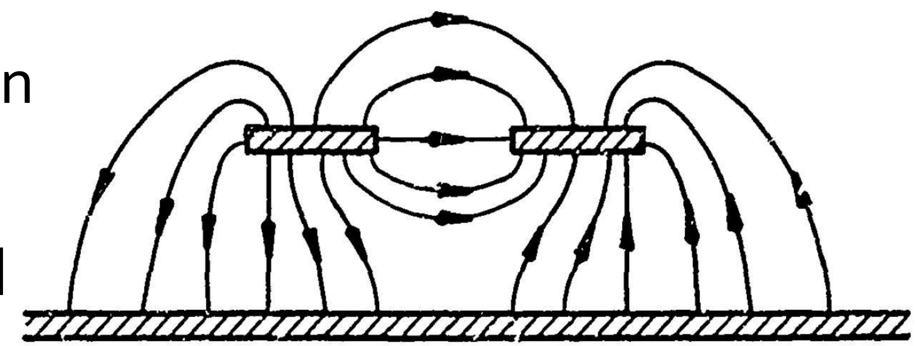


Figure 3.25b



b) EVEN MODE ELECTRIC FIELD PATTERN (SCHEMATIC)

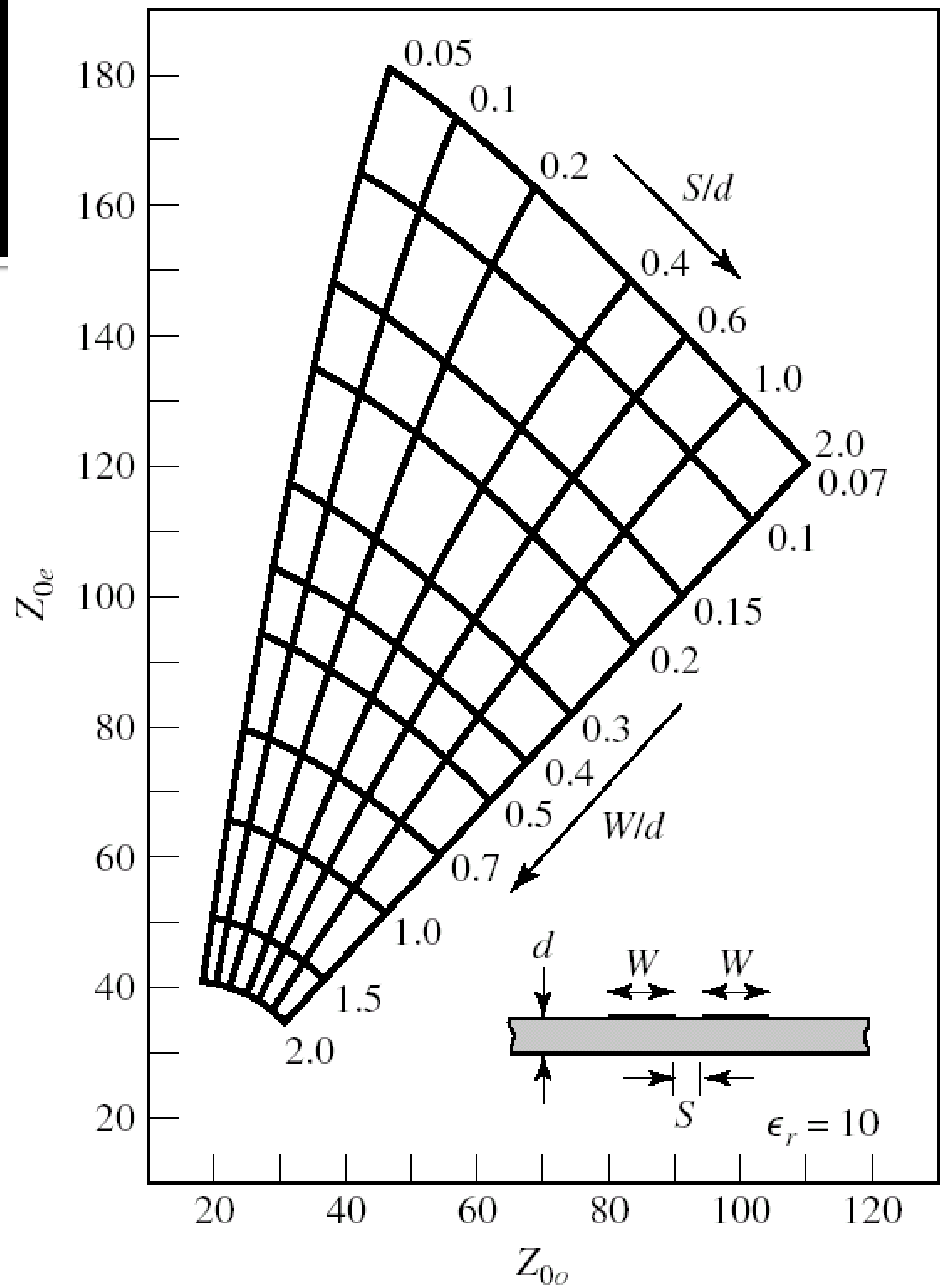
- Even mode - characterizes the common mode signal on the two lines
- Odd mode - characterizes the differential mode signal between the two lines
- Each of the two modes is characterized by **different** characteristic impedances



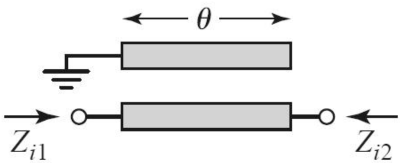
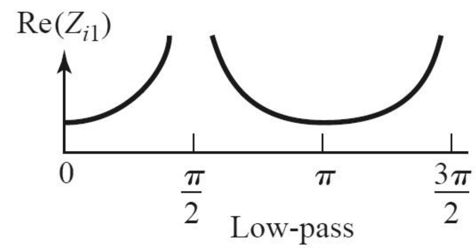
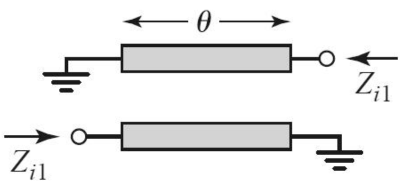
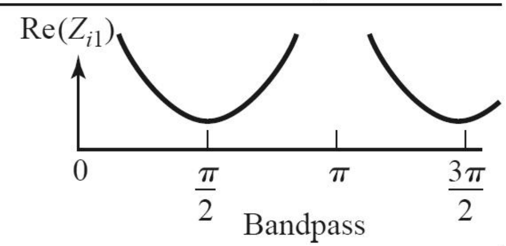
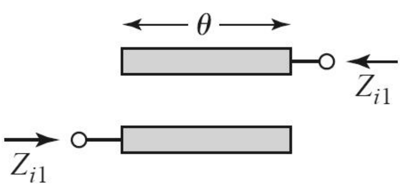
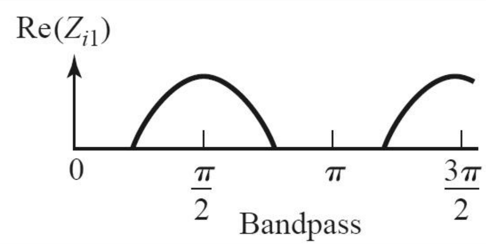
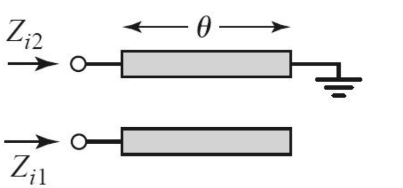
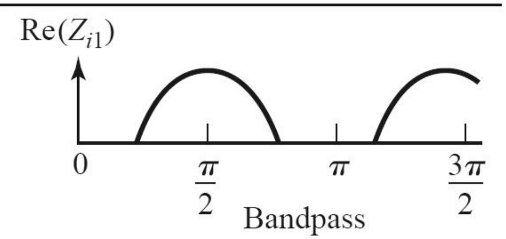
c) ODD MODE ELECTRIC FIELD PATTERN (SCHEMATIC)

Even- and odd-mode characteristic impedance

- Even- and odd-mode characteristic impedance design data for coupled microstrip lines on a substrate with ε_r = 10.



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)

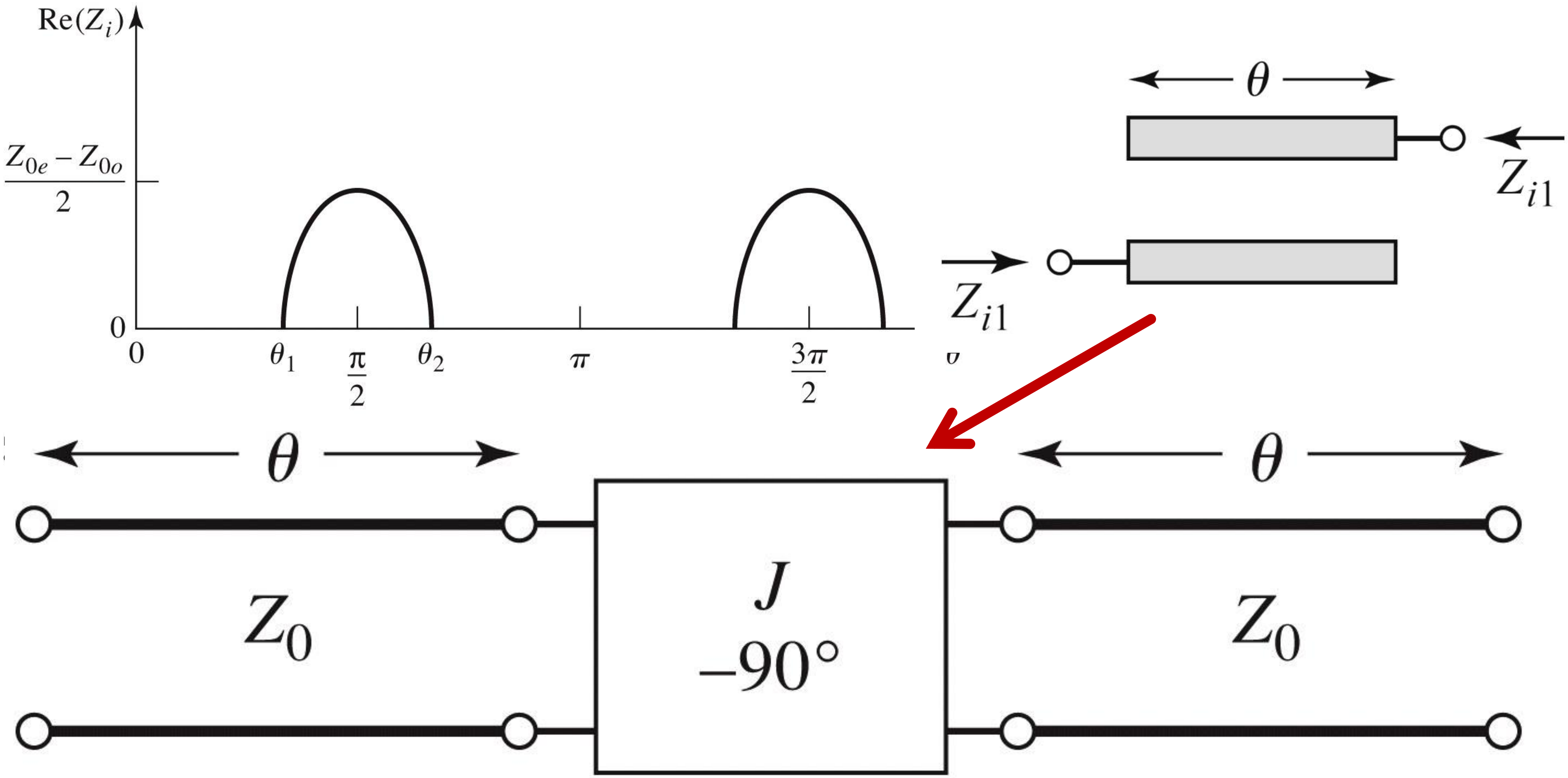
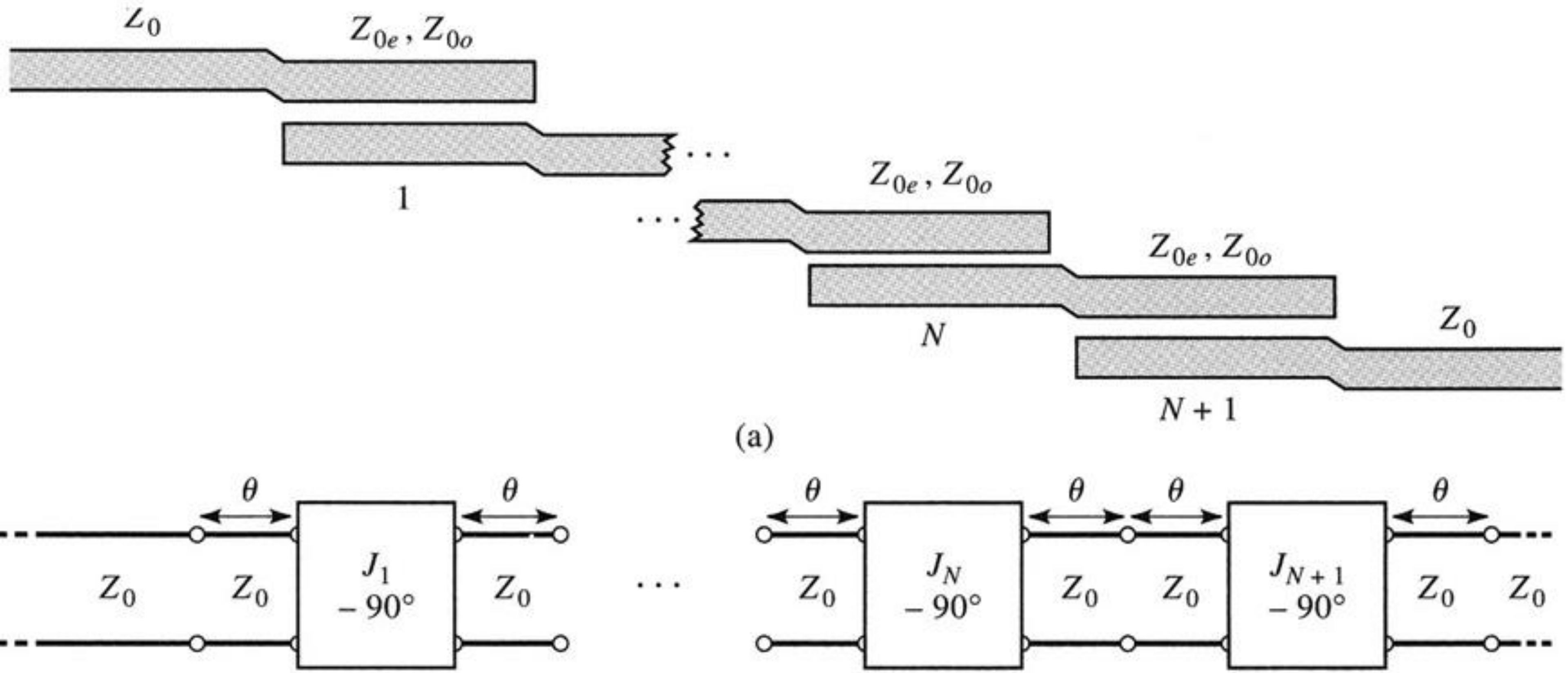


Figure 8.44

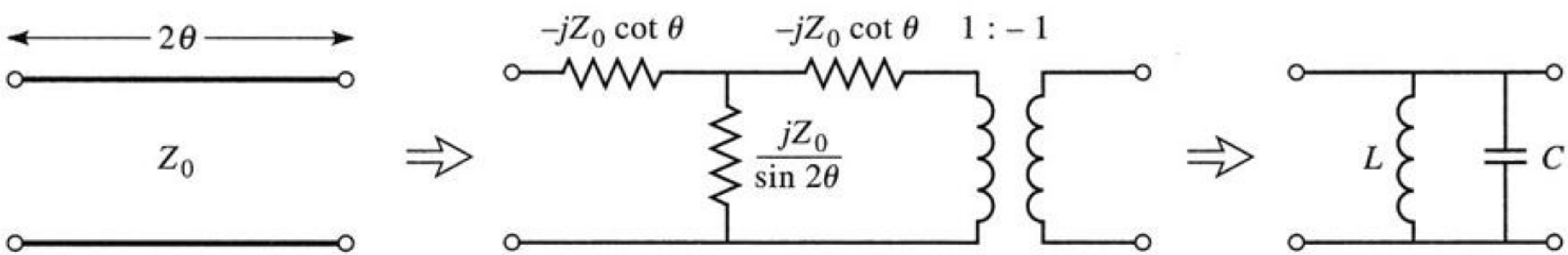
Coupled Line Filters

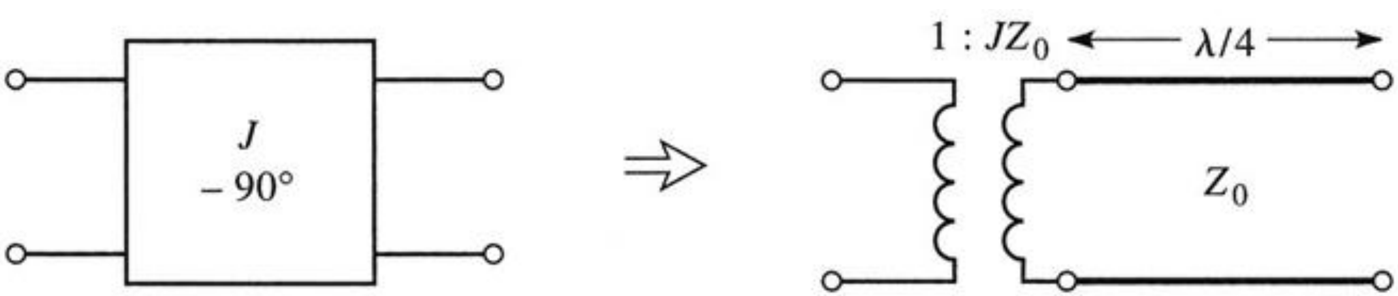
- We get a Nth order filter with N+1 parallel coupled line section



Coupled Line Filters

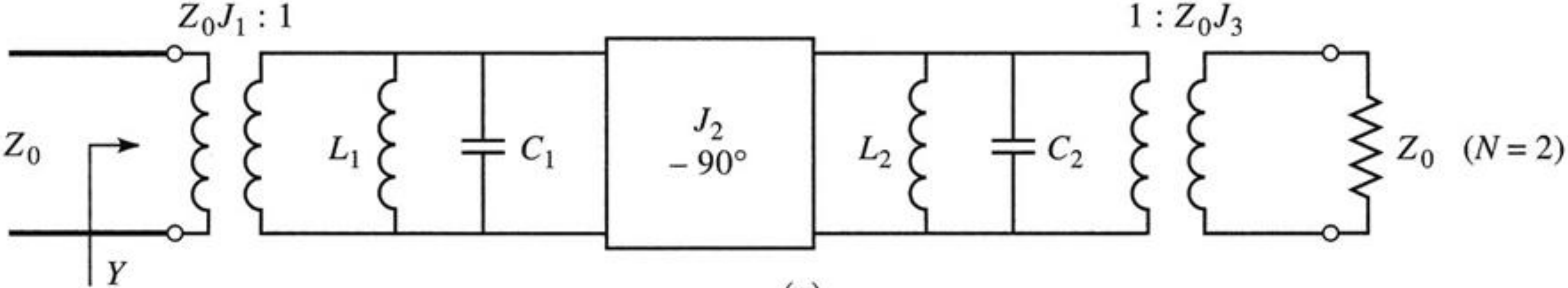
- Equivalent circuits for
  - transmission lines of length 2θ
  - admittance inverters



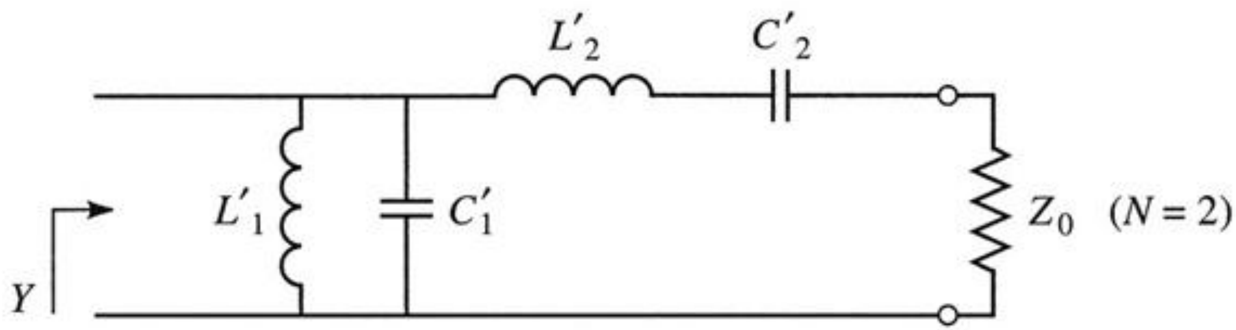


Coupled Line Filters N=2

- We get a 2^{nd} order BPF behavior with 3 coupled lines sections



(e)



(f)

Figure 8.45def

Coupled Line Filters design formulas

- Compute the inverters from prototype parameters

$$Z_0 \cdot J_1 = \sqrt{\frac{\pi \cdot \Delta}{2 \cdot g_1}} \qquad Z_0 \cdot J_n = \frac{\pi \cdot \Delta}{2 \cdot \sqrt{g_{n-1} \cdot g_n}}, n = \overline{2, N} \qquad Z_0 \cdot J_{N+1} = \sqrt{\frac{\pi \cdot \Delta}{2 \cdot g_N \cdot g_{N+1}}}$$

- Compute coupled line parameters Zoe/Zoo (all of length *l*=λ/4)

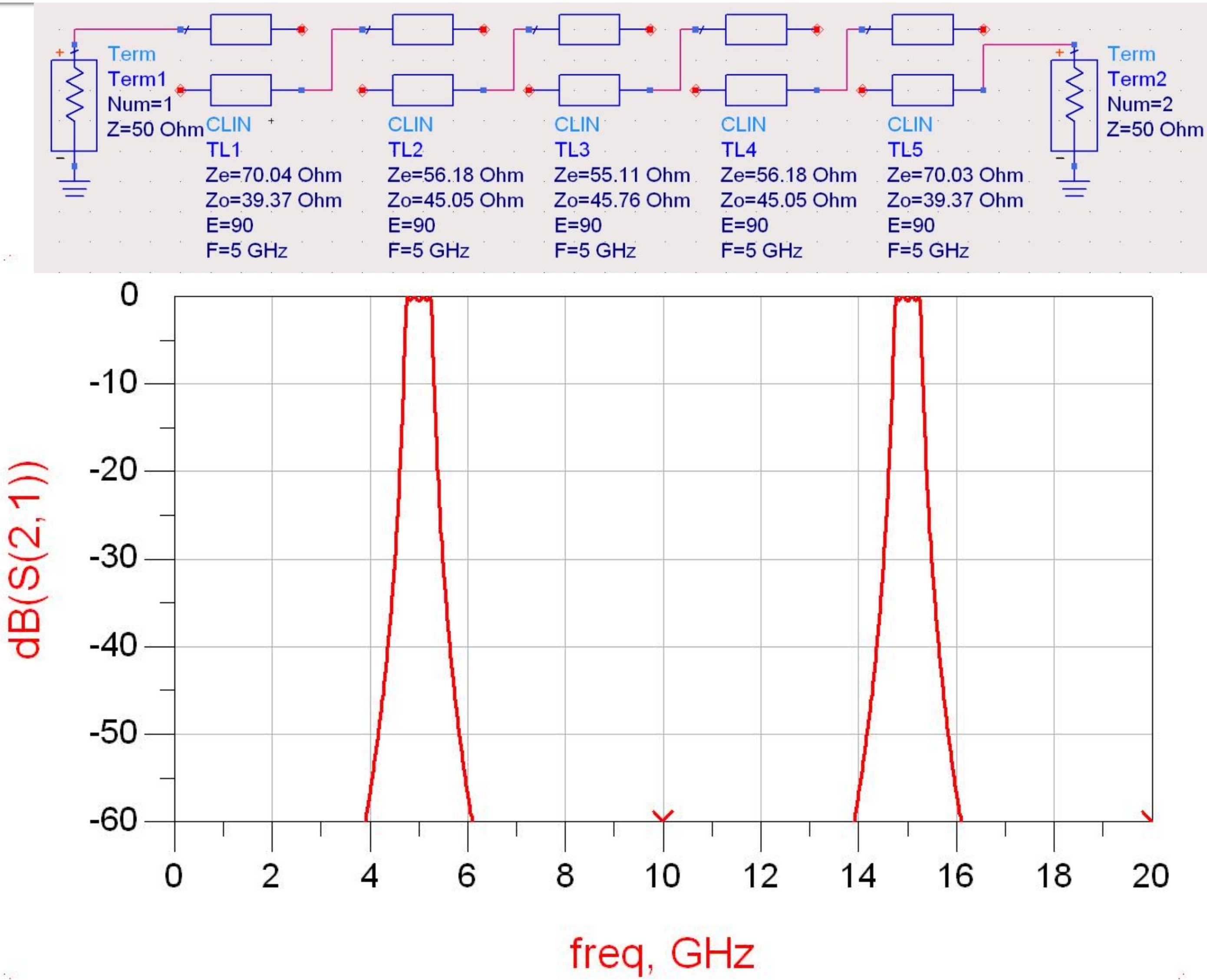
$$Z_{0e,n} = Z_0 \cdot \left[1 + J_n \cdot Z_0 + (J_n \cdot Z_0)^2\right]$$
$$Z_{0o,n} = Z_0 \cdot \left[1 - J_n \cdot Z_0 + (J_n \cdot Z_0)^2\right]$$
$$n = \overline{1, N+1}$$

Example

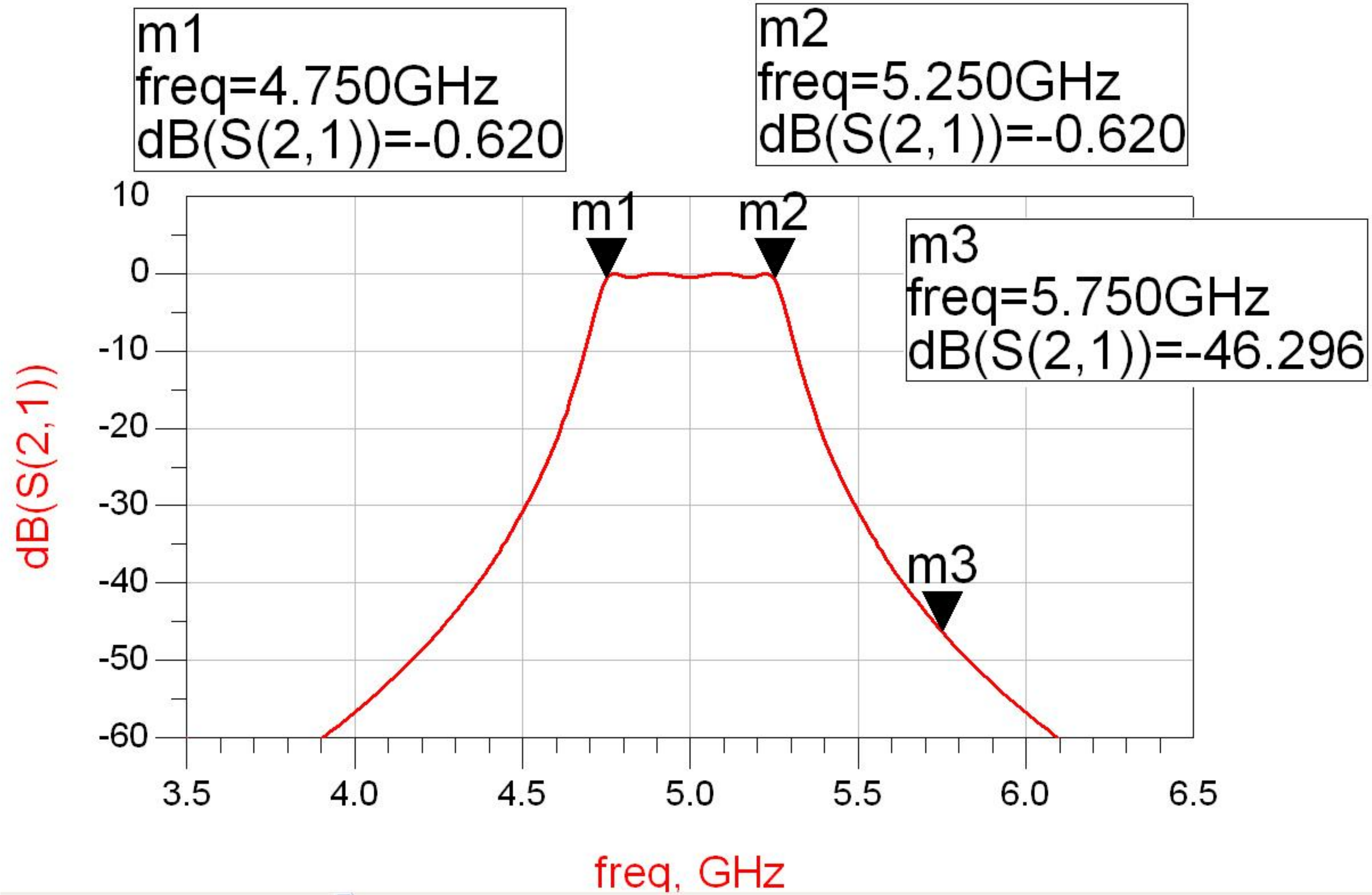
- Similar to a project assignment
- Follows the amplifier designed as in L10
- 4th order bandpass filter, fo = 5GHz, fractional bandwidth of the passband 10 %
- 0.5dB equal-ripple table for g_n followed by filter design formulas

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

ADS – coupled line BPF



ADS – coupled line BPF



Examples

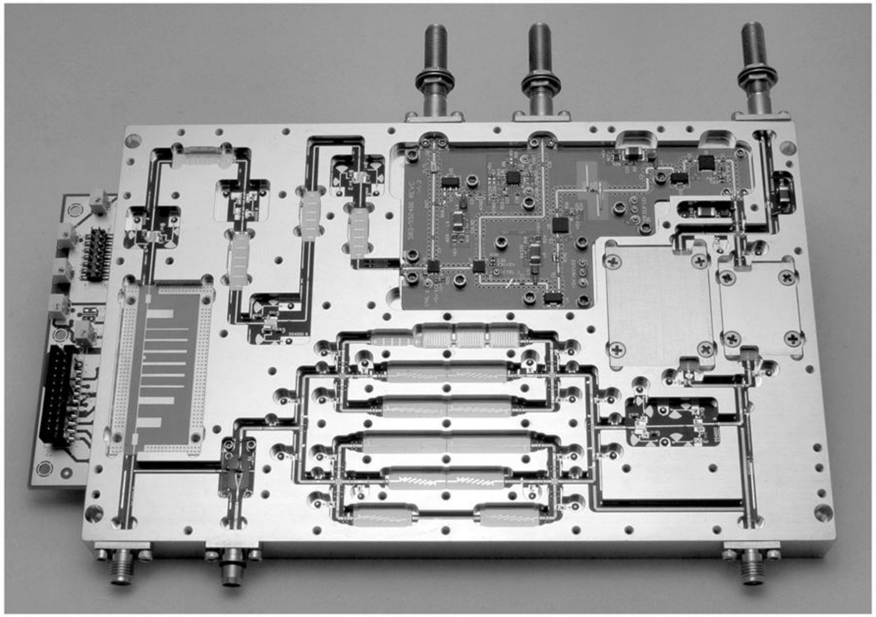
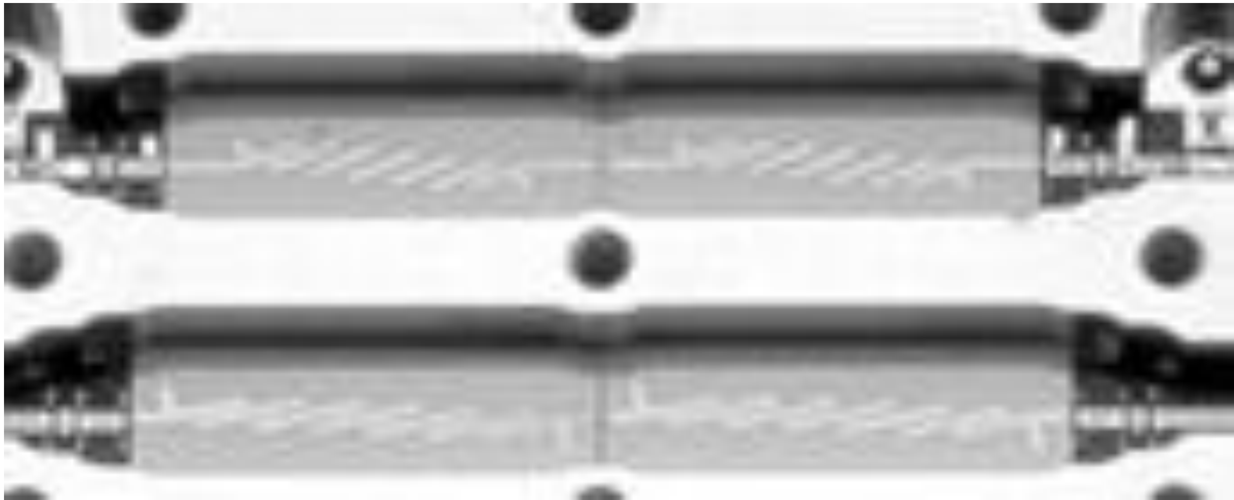
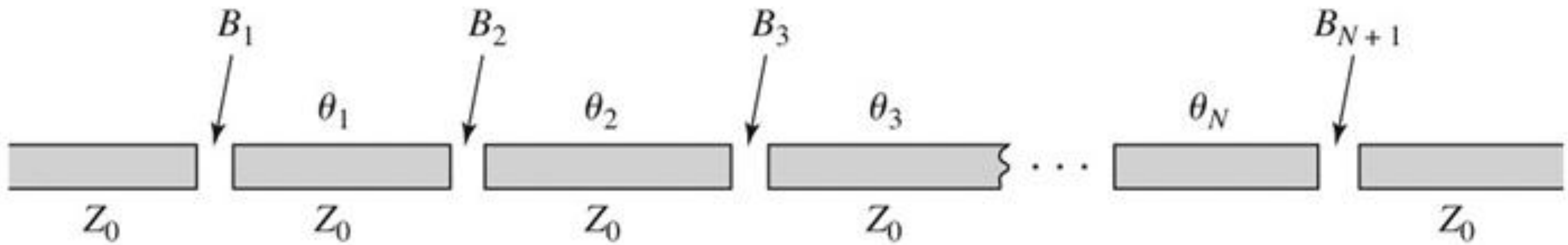


Figure 8.55
Courtesy of LNX Corporation, Salem, N.H.

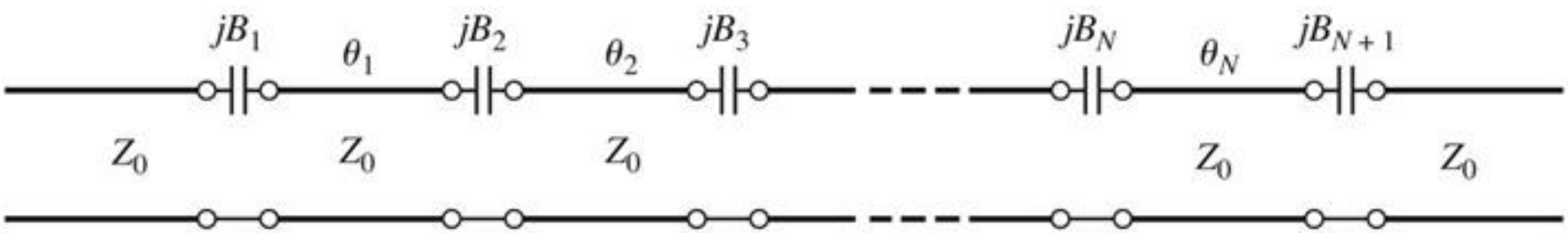


Bandpass Filters Using Capacitively Coupled Series Resonators

- The gaps between the resonators (~λ/2) generate a capacitive coupling between two resonators and can be approximated as series capacitors



(a)



(b)

Bandpass Filters Using Capacitively Coupled Series Resonators

- From the real physical length of the resonators, some part is used implement a admittance inverter (the remainder ϕ=π, *l*=λ/2, resonator)

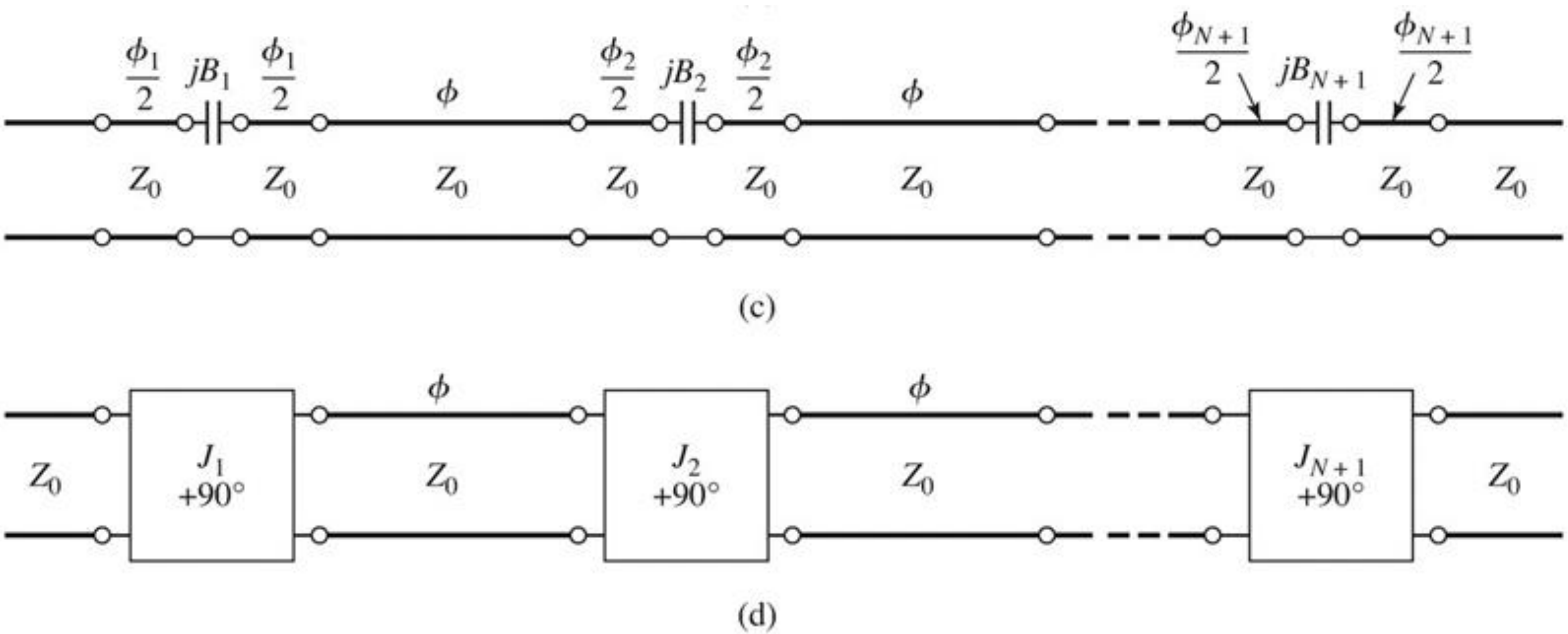


Figure 8.50

Bandpass Filters Using Capacitively Coupled Series Resonators design

- Compute the inverters (similar to coupled lines)

$$Z_0 \cdot J_1 = \sqrt{\frac{\pi \cdot \Delta}{2 \cdot g_1}} \qquad Z_0 \cdot J_n = \frac{\pi \cdot \Delta}{2 \cdot \sqrt{g_{n-1} \cdot g_n}}, n = \overline{2, N} \qquad Z_0 \cdot J_{N+1} = \sqrt{\frac{\pi \cdot \Delta}{2 \cdot g_N \cdot g_{N+1}}}$$

- Compute capacitive susceptances

$$B_n = \frac{J_n}{1 - (Z_0 \cdot J_n)^2}, n = \overline{1, N+1}$$

- Compute the line lengths that must be "borrowed" to implement the inverters

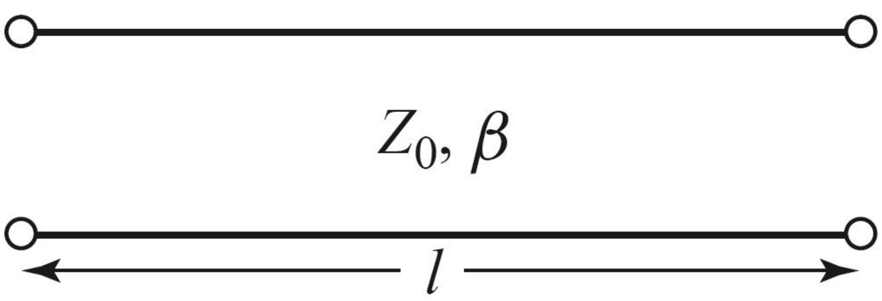
$$\phi_n = -\tan^{-1}(2 \cdot Z_0 \cdot B_n), n = \overline{1, N+1} \qquad \phi_n < 0, n = \overline{1, N+1}$$

- Compute the actual length of the lines (λ/2 + borr.)

$$\theta_i = \pi + \frac{1}{2} \cdot (\phi_i + \phi_{i+1}) = \pi - \frac{1}{2} \cdot \left[\tan^{-1}(2 \cdot Z_0 \cdot B_i) + \tan^{-1}(2 \cdot Z_0 \cdot B_{i+1})\right], i = \overline{1, N}$$

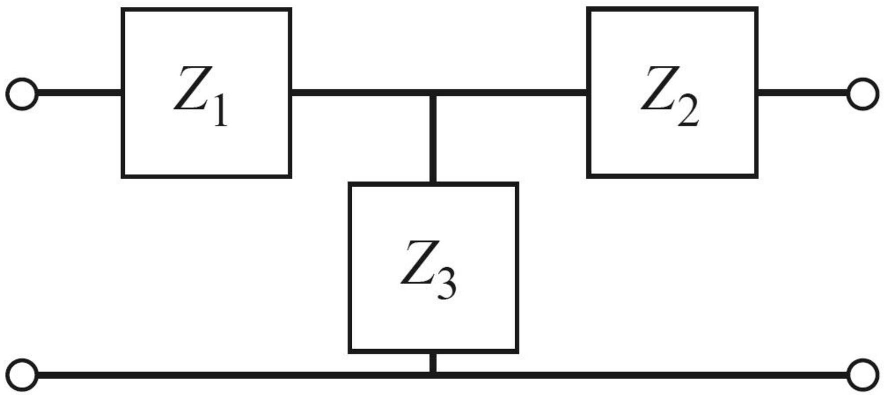
Equivalent circuits for short sections of transmission lines

- ABCD matrix (L5)
- short line, model with lumped elements is valid



$$A = \cos \beta \cdot l \qquad B = j \cdot Z_0 \cdot \sin \beta \cdot l$$

$$C = j \cdot Y_0 \cdot \sin \beta \cdot l \qquad D = \cos \beta \cdot l$$



$$A = 1 + \frac{Z_1}{Z_3} \qquad B = Z_1 + Z_2 + \frac{Z_1 \cdot Z_2}{Z_3}$$

$$C = \frac{1}{Z_3} \qquad D = 1 + \frac{Z_2}{Z_3}$$

Equivalent circuits for short sections of transmission lines

- The shunt element is capacitive

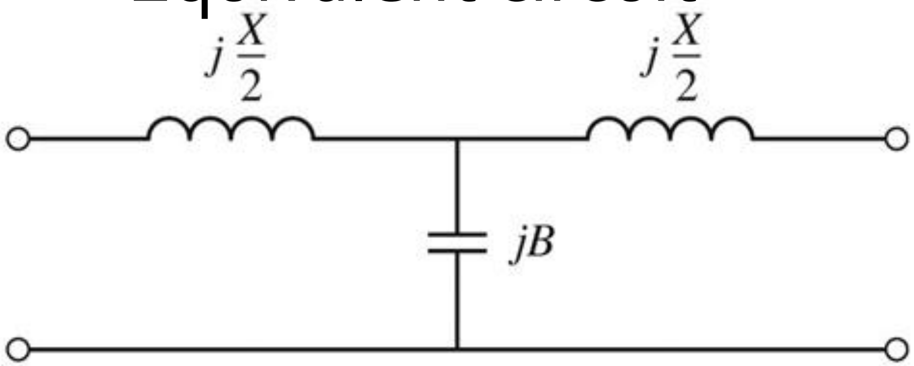
$$Z_3 = \frac{1}{j \cdot Y_0 \cdot \sin \beta \cdot l}$$

- Series elements are equal, and inductive

$$\cos \beta \cdot l = 1 + \frac{Z_1}{Z_3} = 1 + \frac{Z_2}{Z_3}$$

$$Z_1 = Z_2 = Z_3 \cdot \left(\cos \beta \cdot l - 1\right) = -j \cdot Z_0 \cdot \frac{\cos \beta \cdot l - 1}{\sin \beta \cdot l} = j \cdot Z_0 \cdot \tan \frac{\beta \cdot l}{2}$$

- Equivalent circuit

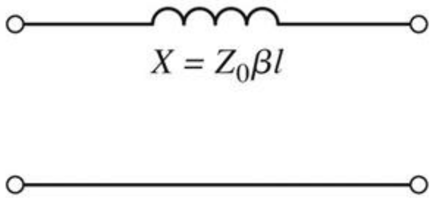


$$\frac{X}{2} = Z_0 \cdot \tan \frac{\beta \cdot l}{2}$$

$$B = \frac{1}{Z_0} \cdot \sin \beta \cdot l$$

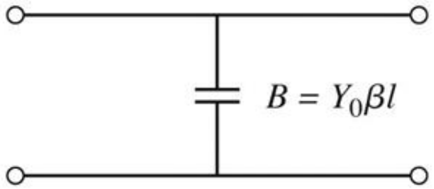
Equivalent circuits for short sections of transmission lines

- depending on the characteristic impedance:
  - high Zo >>



$$X \cong Z_0 \cdot \beta \cdot l \qquad \beta \cdot l < \frac{\pi}{4} \qquad Z_0 = Z_h$$

  - low Zo <<



$$B \cong Y_0 \cdot \beta \cdot l \qquad \beta \cdot l < \frac{\pi}{4} \qquad Z_0 = Z_l$$

Stepped-impedance low-pass filter

- Series L, shunt C, we realize low-pass filters
- We use
  - lines with high characteristic impedance to implement an series inductor

$$\beta \cdot l = \frac{L \cdot R_0}{Z_h}$$

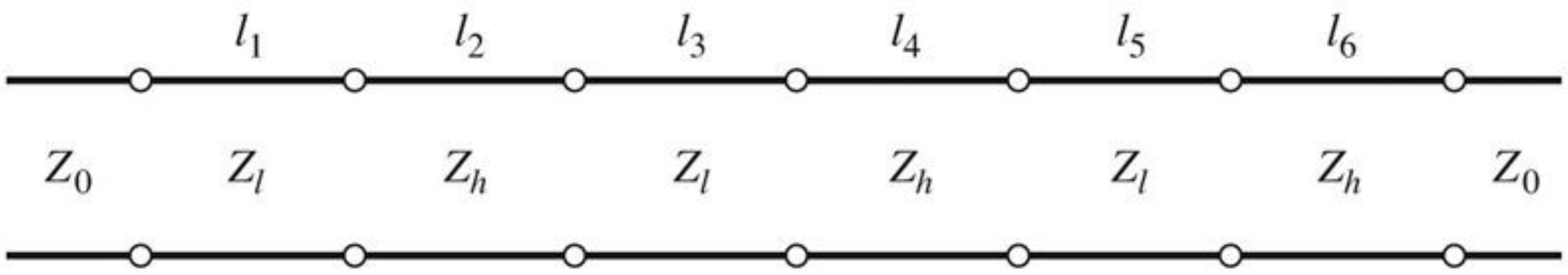
  - lines with low characteristic impedance to implement a shunt capacitor

$$\beta \cdot l = \frac{C \cdot Z_l}{R_0}$$

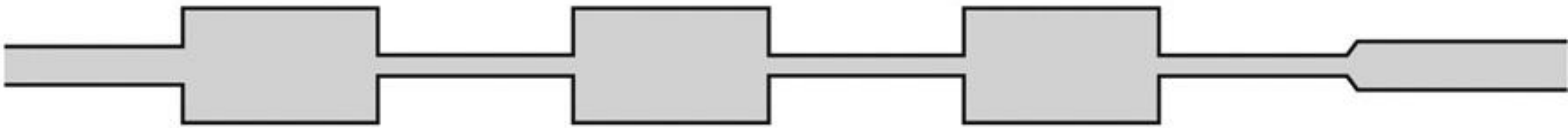
- usually the highest and lowest characteristic impedance that can be practically fabricated

Stepped-impedance LPF

- Not all the lines will result with the same length so the filter response is not periodic in frequency



(b)



(c)

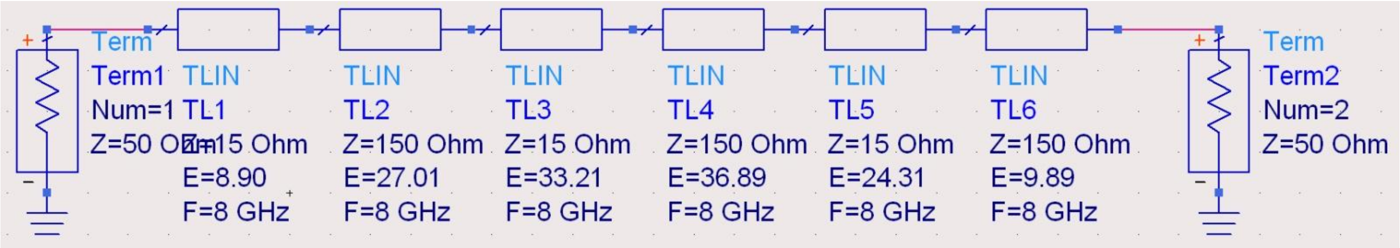
Figure 8.40

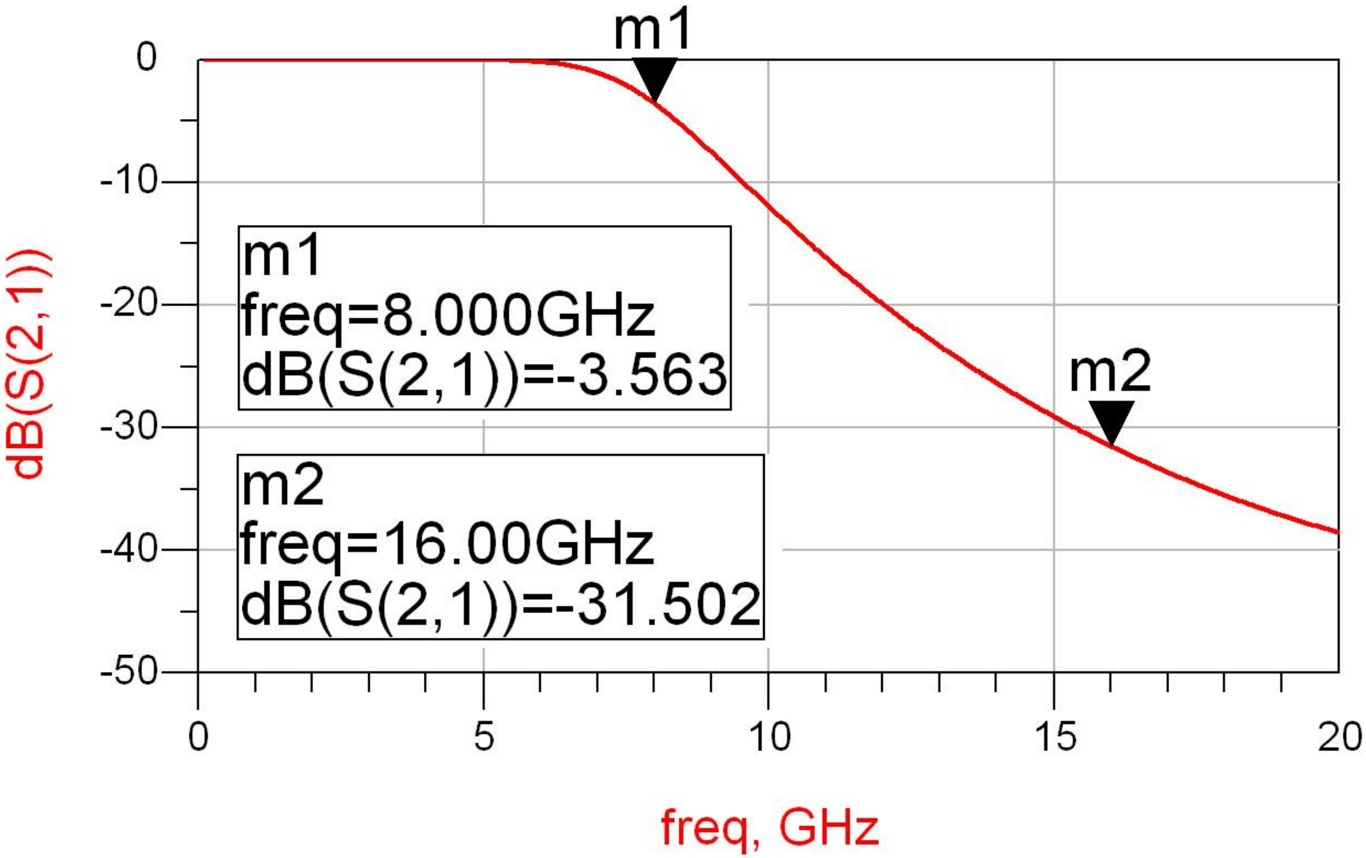
Example

- LPF with 8GHz cutoff frequency, 6th order. Maximum realizable impedance is 150Ω and lowest 15Ω.

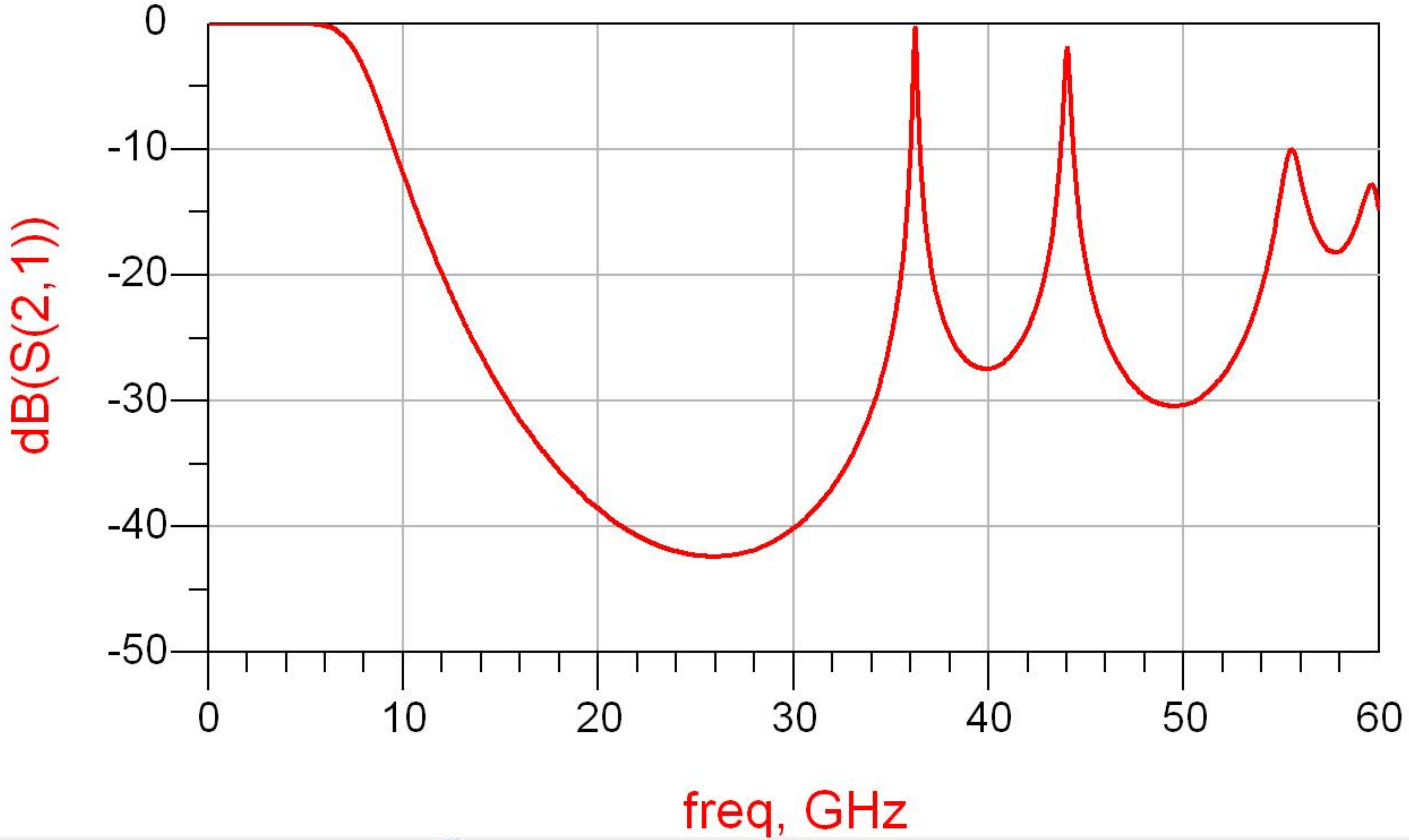
n	g_n	L/C_n	Z	θ_n[rad]	θ_n[°]
1	0.5176	0.206pF	15	0.155	8.90
2	1.4142	1.407nH	150	0.471	27.01
3	1.9318	0.769pF	15	0.580	33.21
4	1.9318	1.922nH	150	0.644	36.89
5	1.4142	0.563pF	15	0.424	24.31
6	0.5176	0.515nH	150	0.173	9.89

ADS – Stepped-impedance LPF

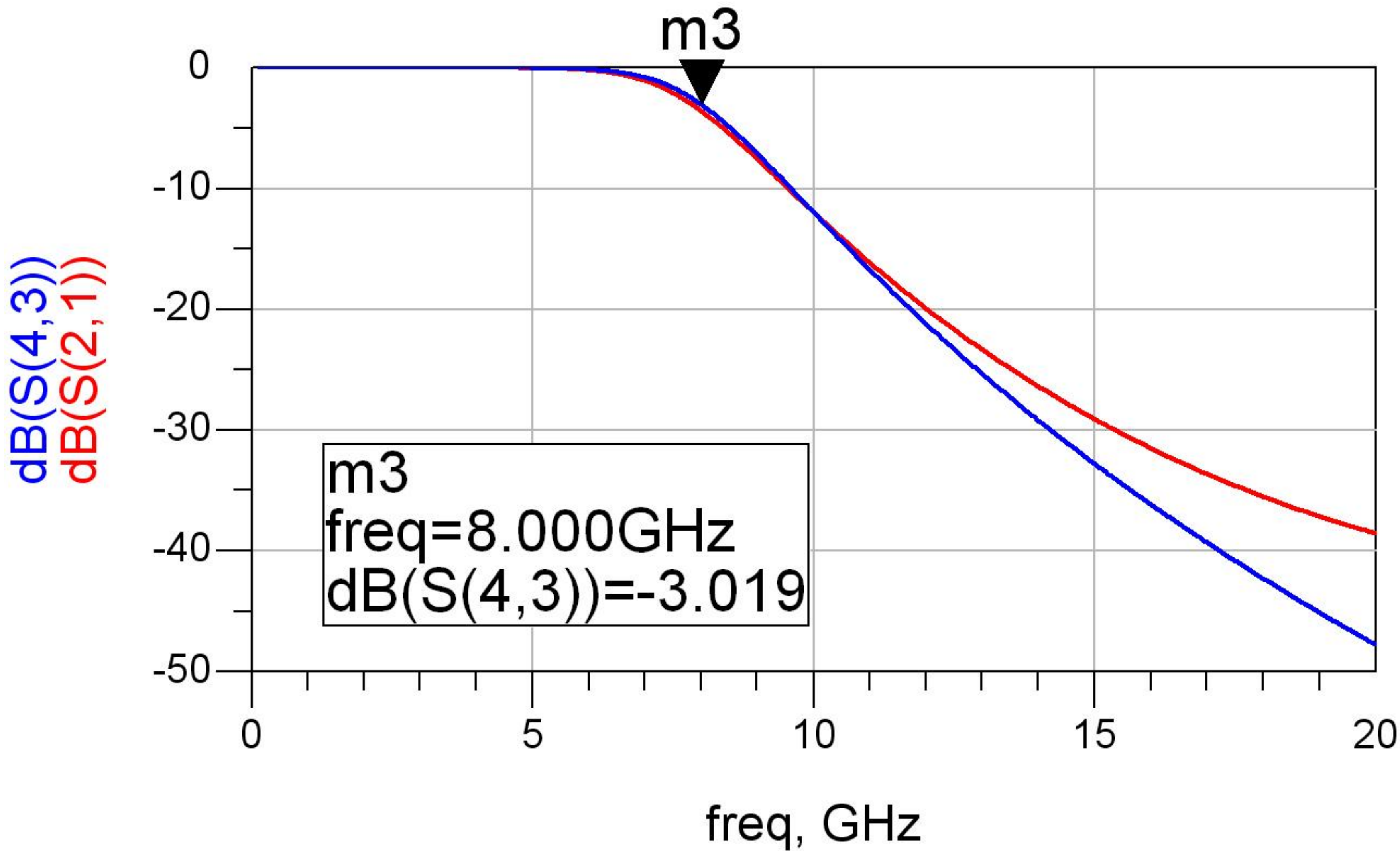




ADS – Stepped-impedance LPF



ADS – Stepped-impedance LPF – compared with lumped elements



Examples

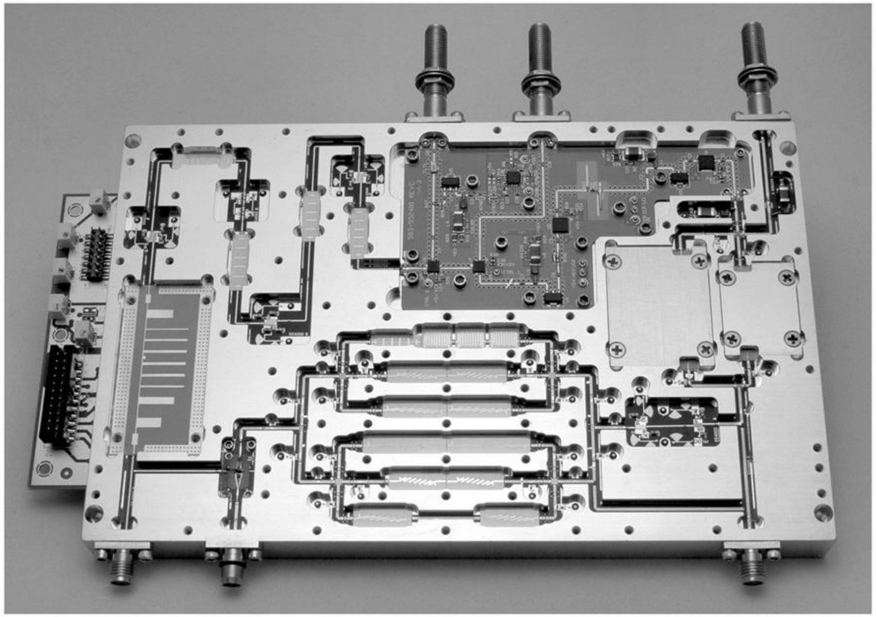
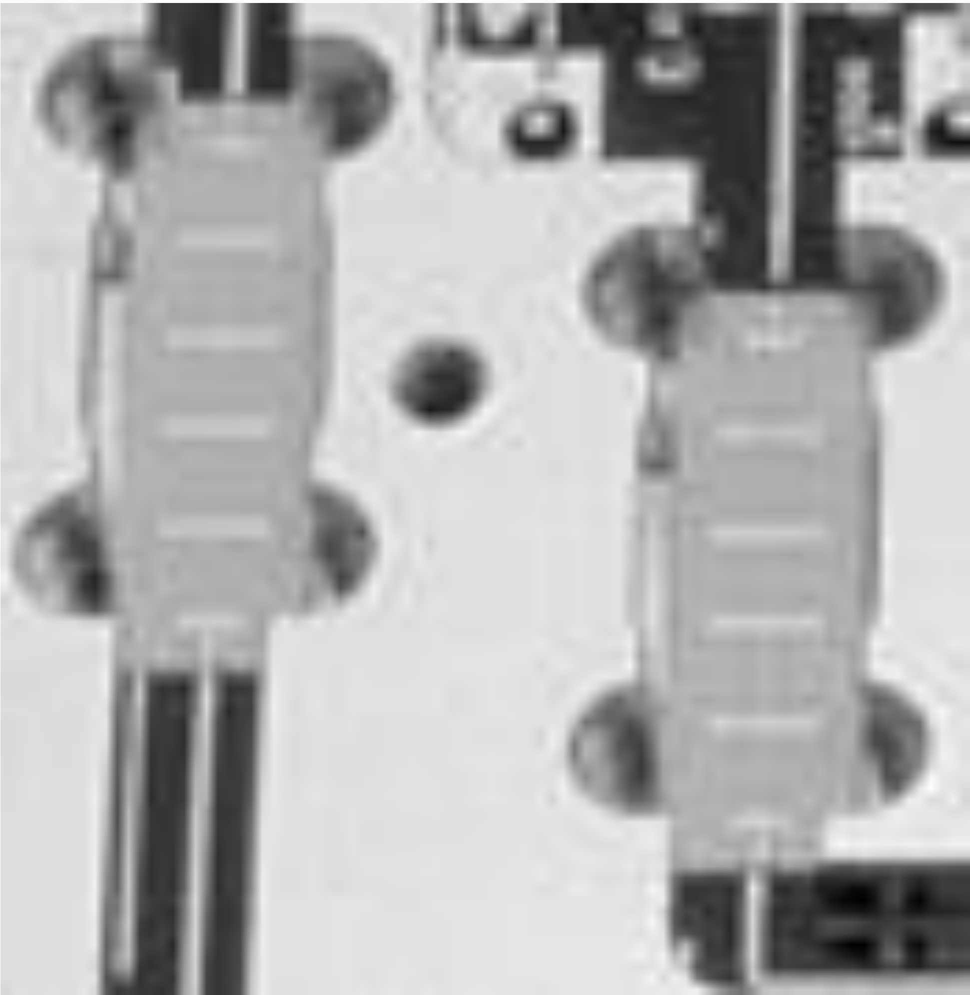


Figure 8.55
Courtesy of LNX Corporation, Salem, N.H.



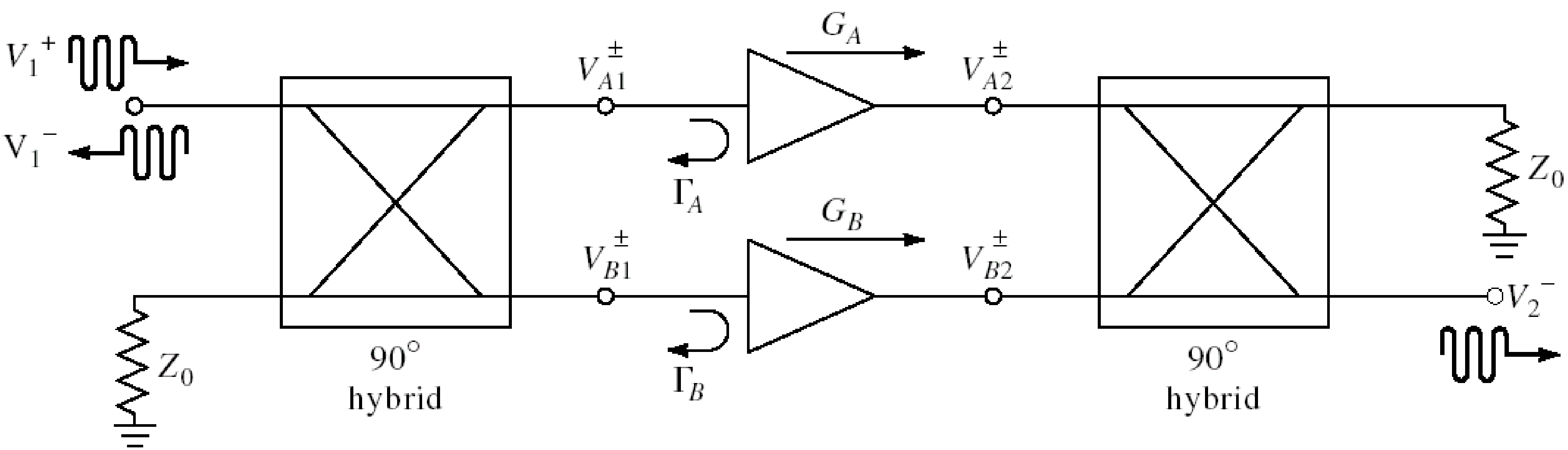
Microwave Amplifiers

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



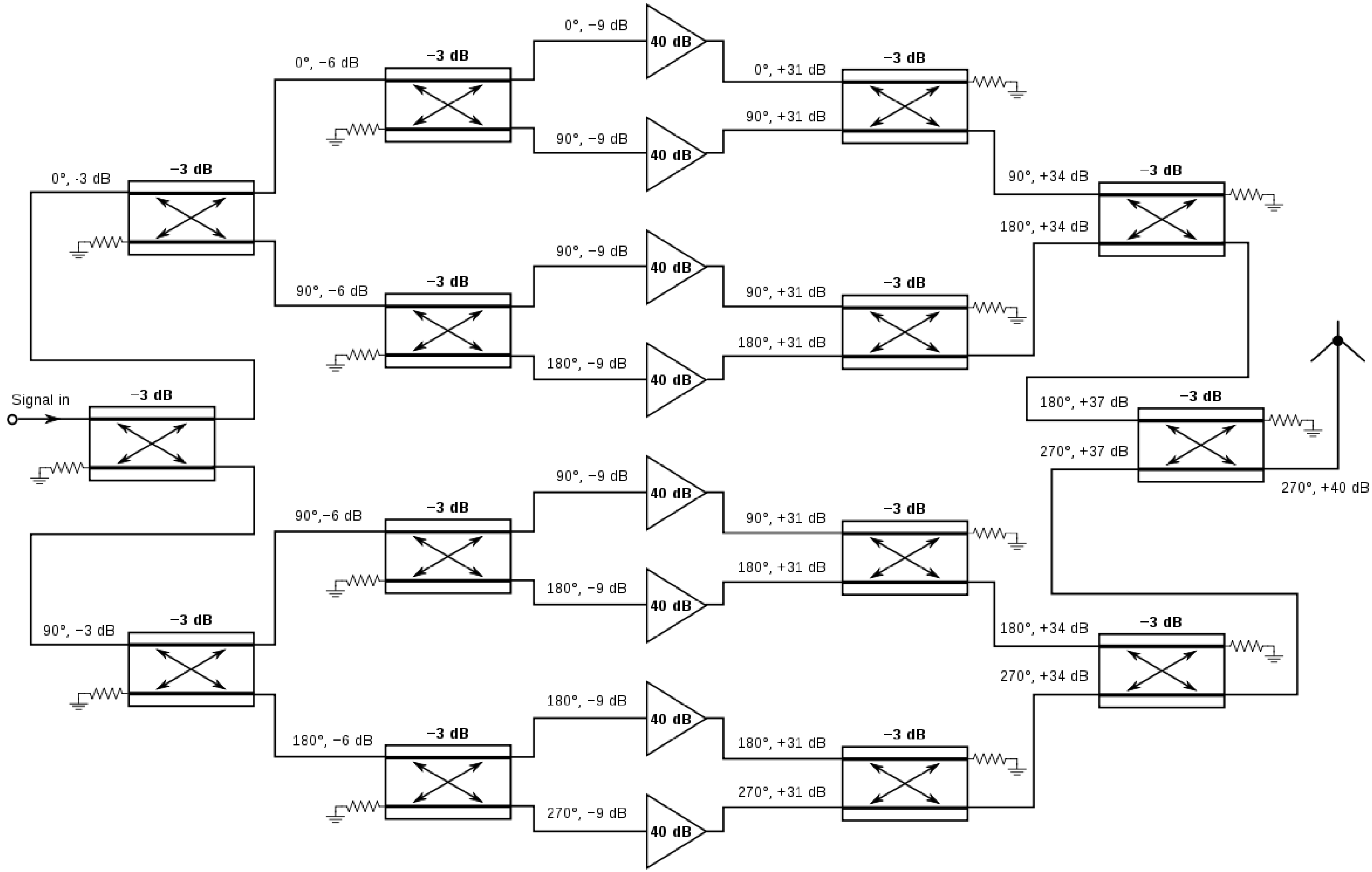
- two identical amplifiers with two hybrid couplers 3 dB / 90° to cancel input and output reflections

$$S_{21} = \frac{-j}{2}\cdot\left(G_A + G_B\right) \qquad S_{21}\big|_{A=B} = -j\cdot G$$

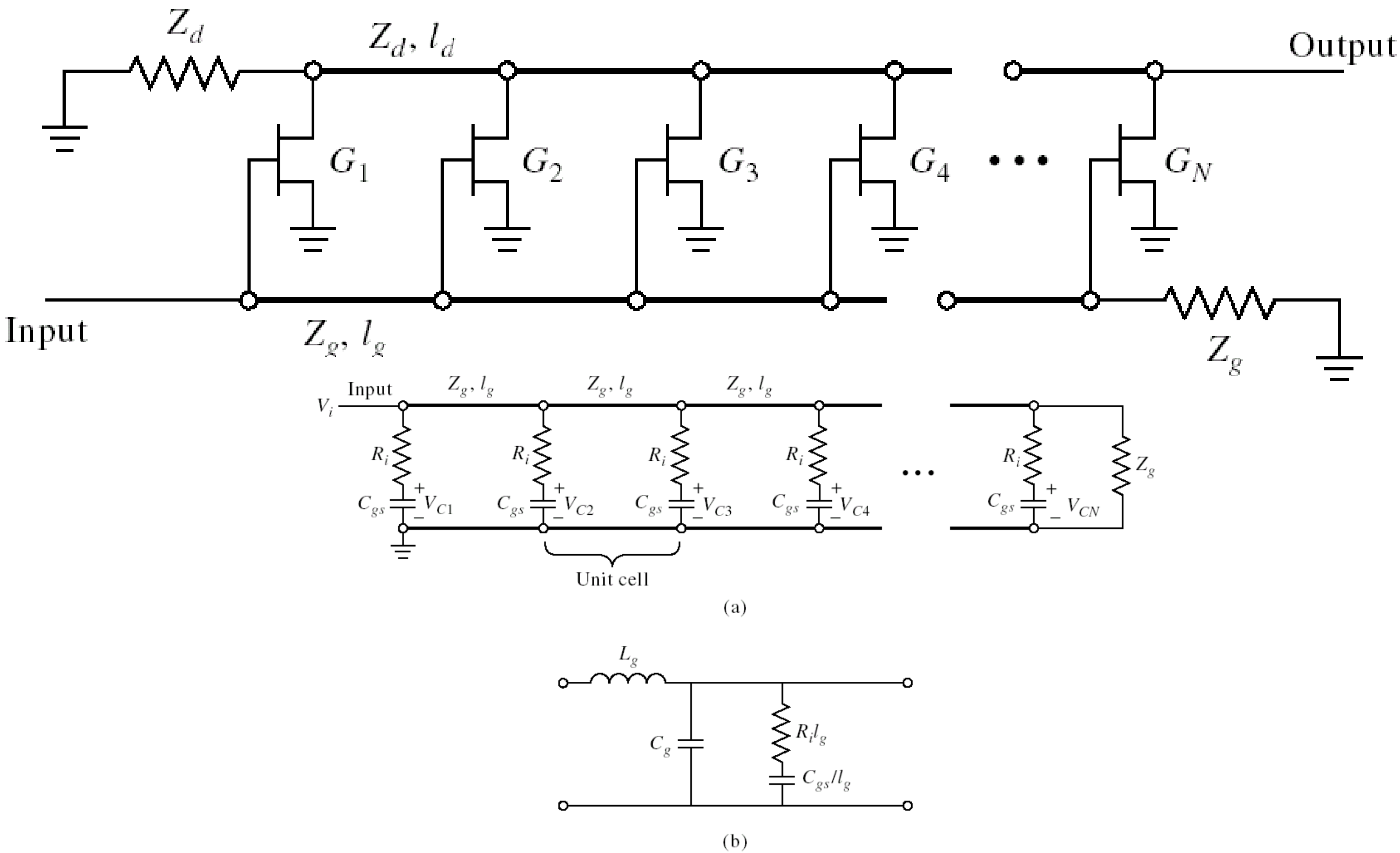
$$F = \frac{1}{2}\cdot\left(F_A + F_B\right)$$

$$S_{11} = \frac{1}{2}\cdot\left(\Gamma_A - \Gamma_B\right) \qquad S_{11}\big|_{A=B} = 0$$

Balanced amplifiers



Distributed amplifiers



Distributed amplifiers

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

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

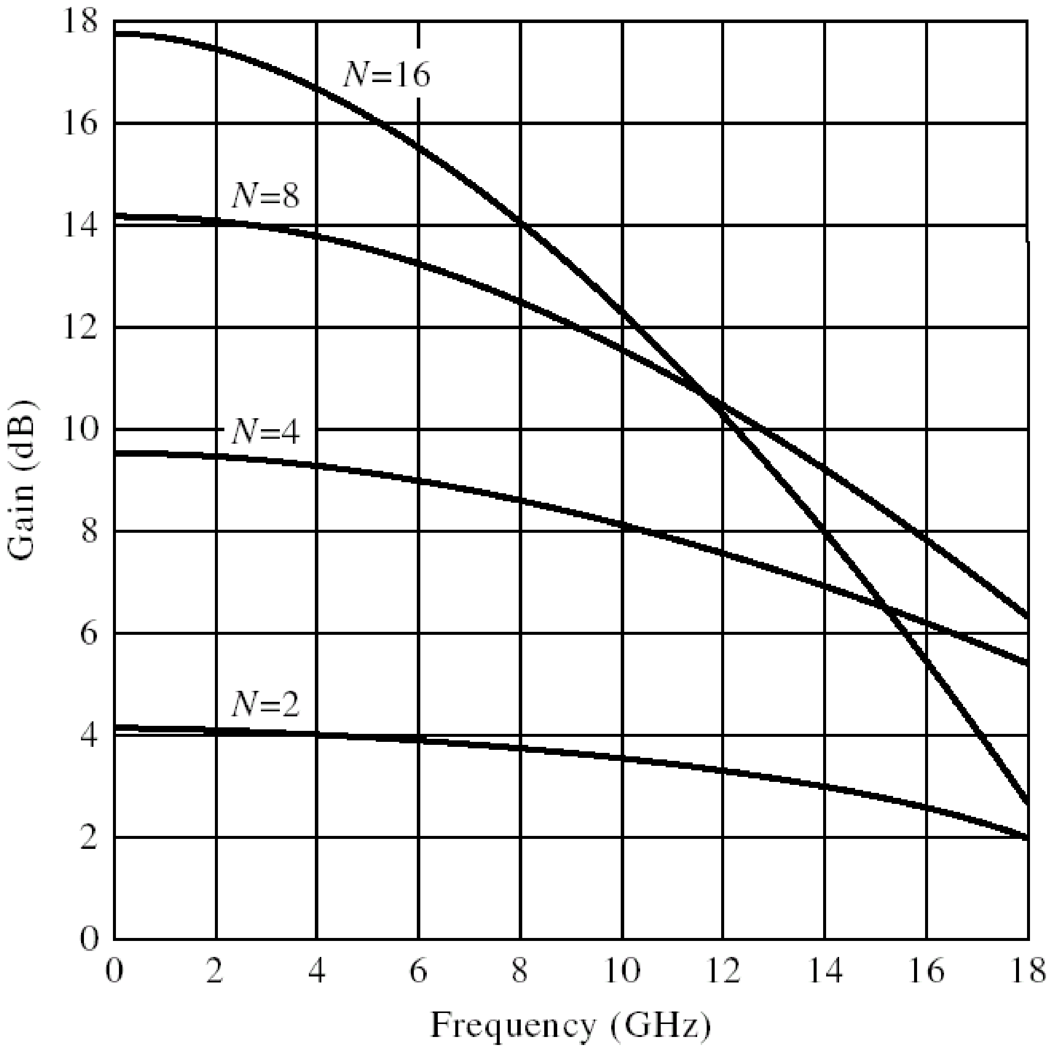
- Power gain

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

- Lossless power gain

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

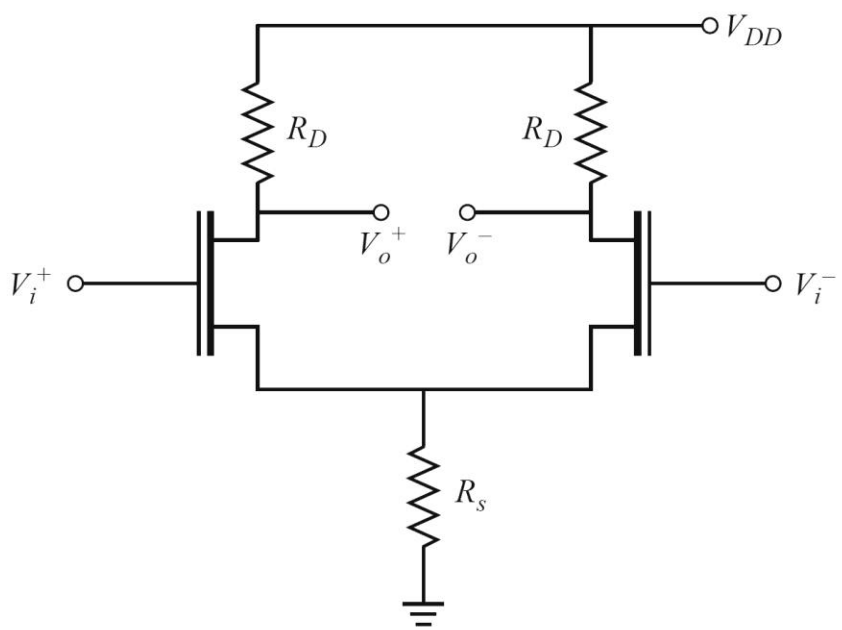
Distributed amplifiers



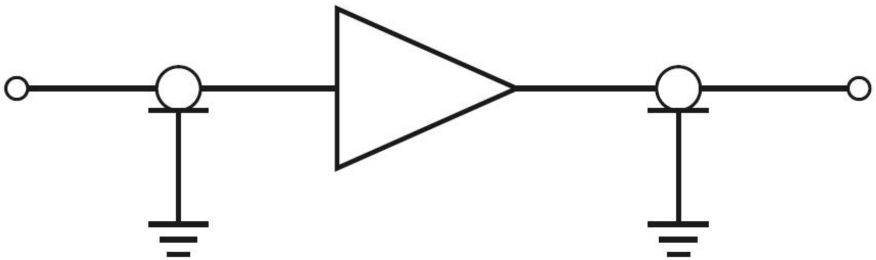
$$N_{opt} = \frac{\ln(\alpha_g \cdot l_g) - \ln(\alpha_d \cdot l_d)}{\alpha_g \cdot l_g - \alpha_d \cdot l_d}$$

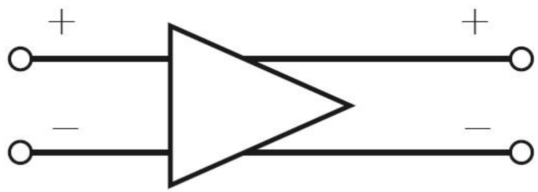
Differential amplifiers

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

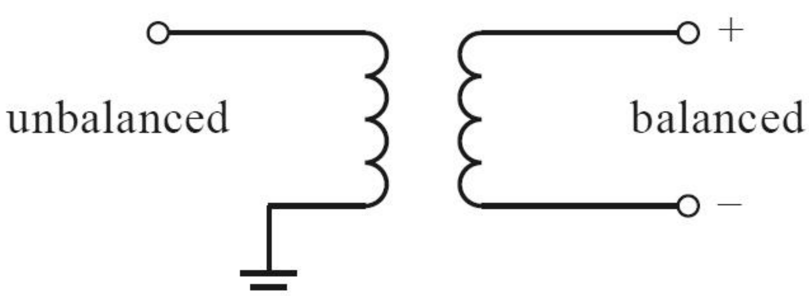


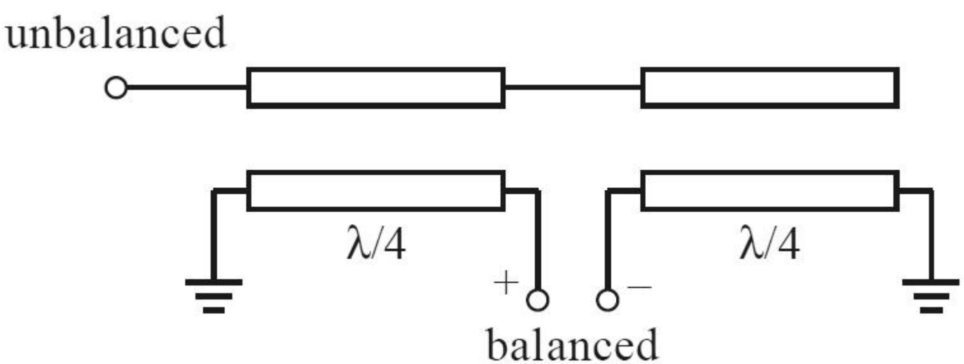
Differential amplifiers





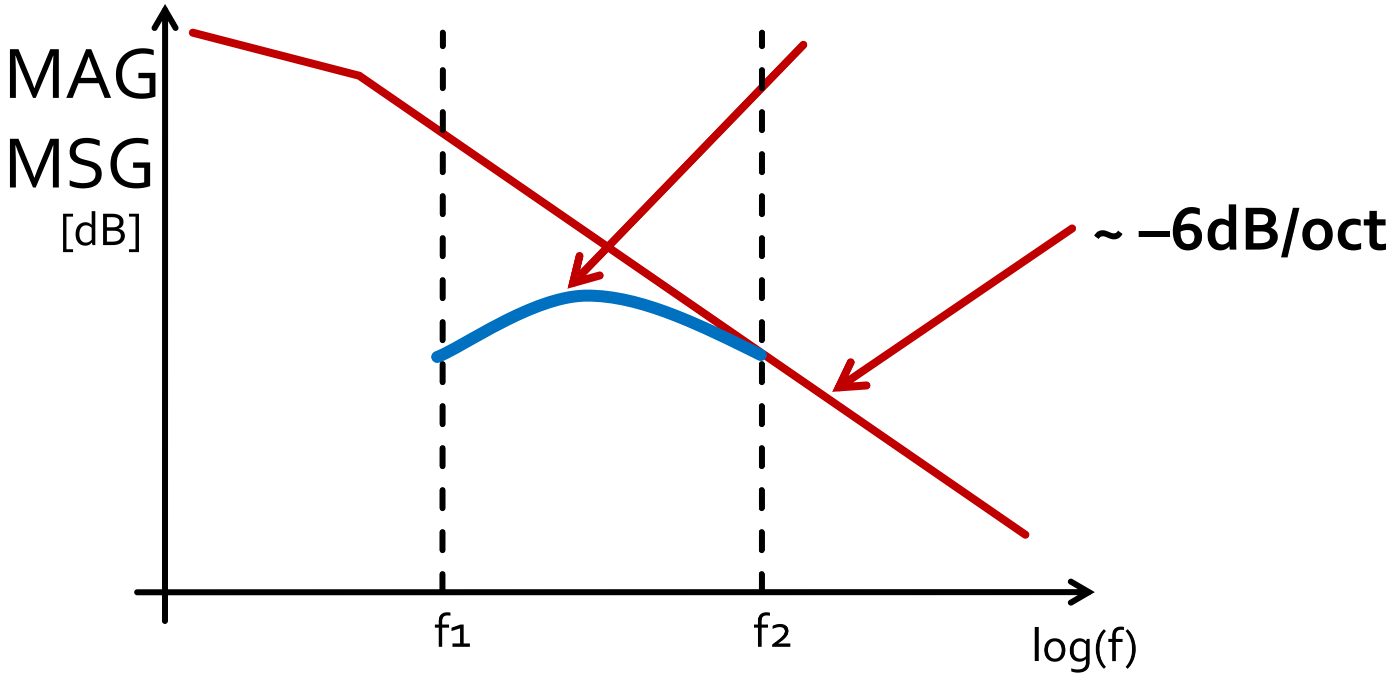
- We use circuits to transition from an unbalanced signal to a balanced signal (or vice versa)
  - hybrid couplers 3dB / 180°
  - "balun" (balanced - unbalanced)





Compensated matching networks

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

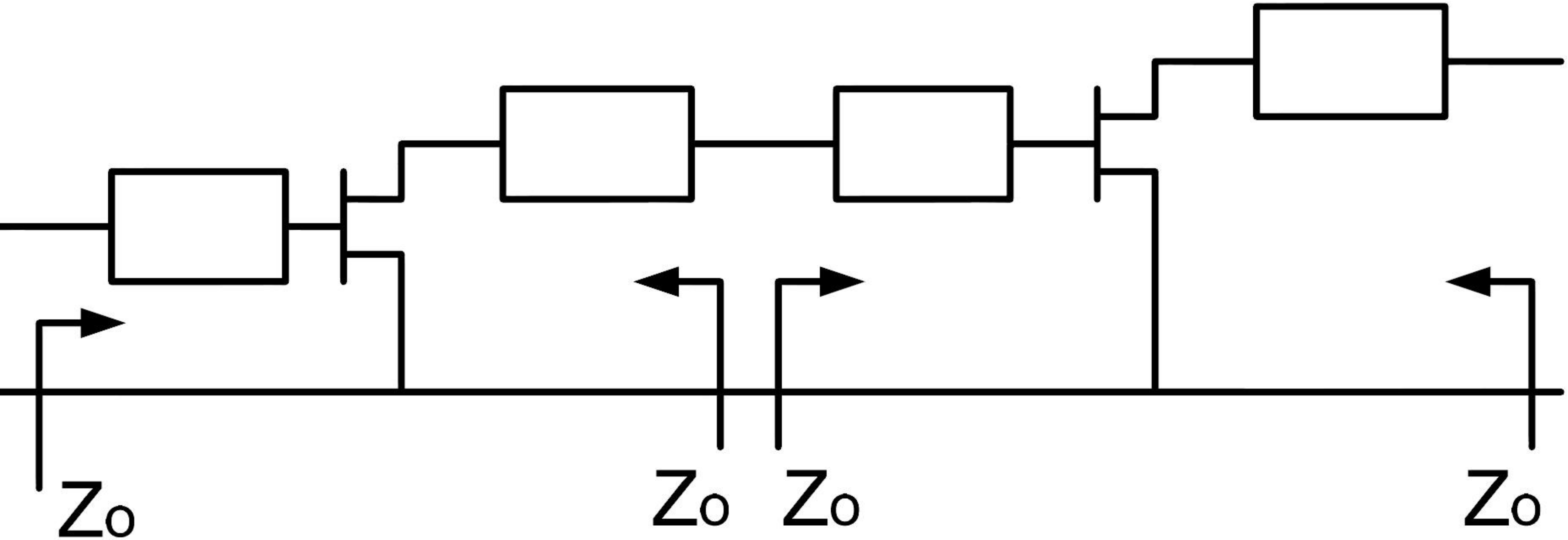


Microwave Amplifiers

Multistage Amplifier Design

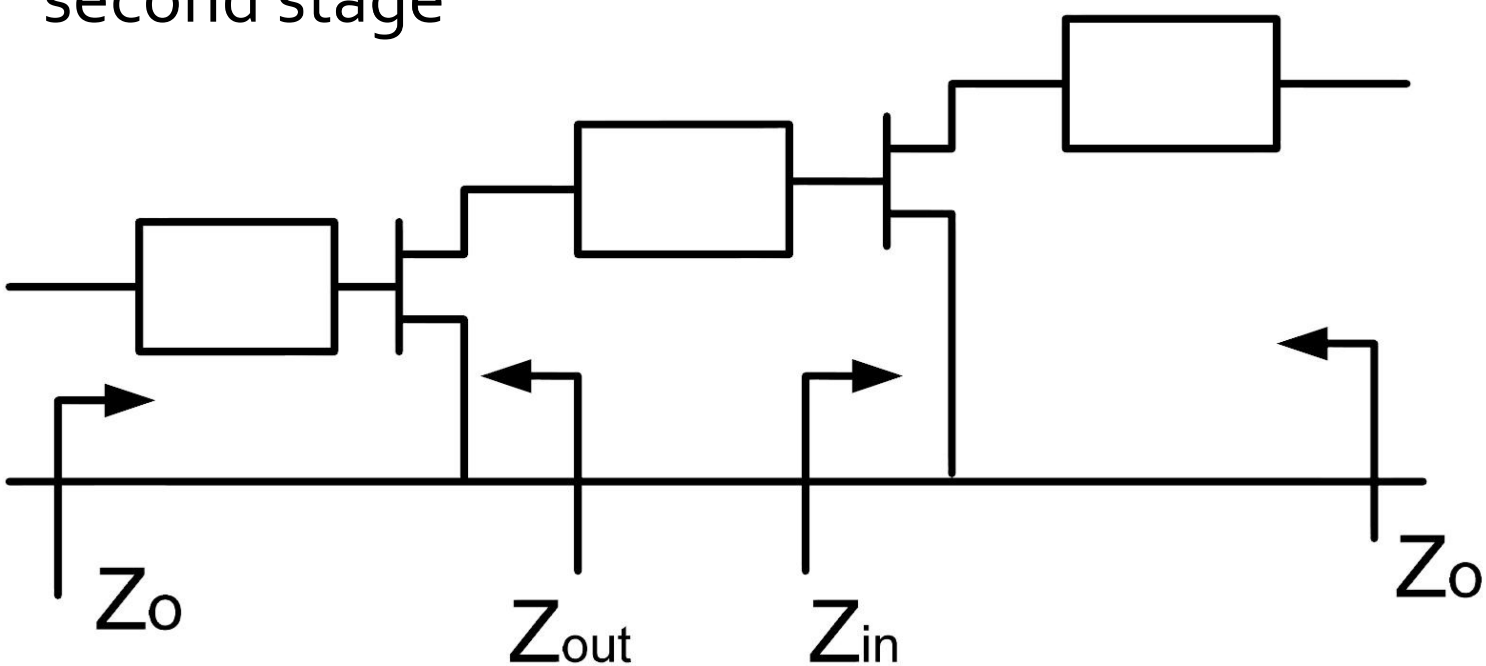
Multistage amplifiers

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



Multistage amplifiers

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

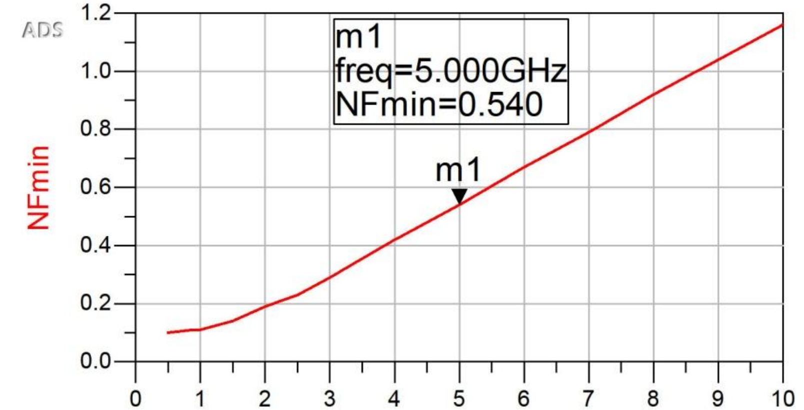


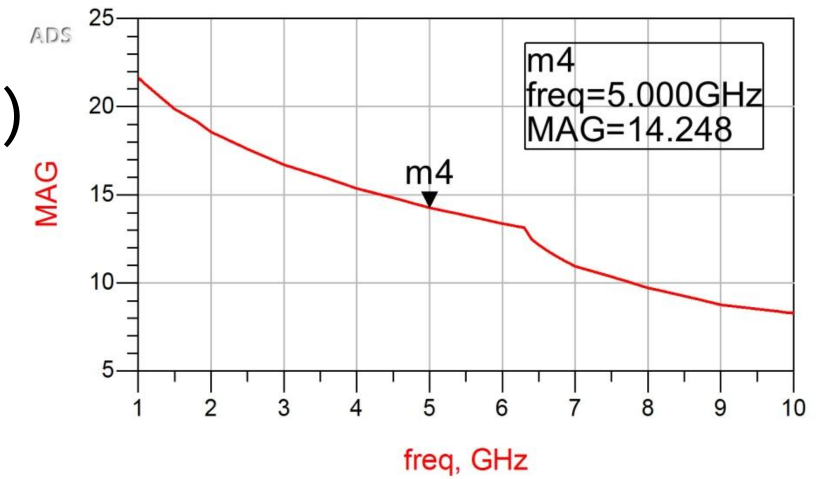
Example multistage LNA

- 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)

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

- 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**

AppCAD - [NoiseCalc]

File Calculate Application Examples Options Help

NoiseCalc Set Number of Stages = 2 Calculate [F4]

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

Enter System Parameters:

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

System Analysis:

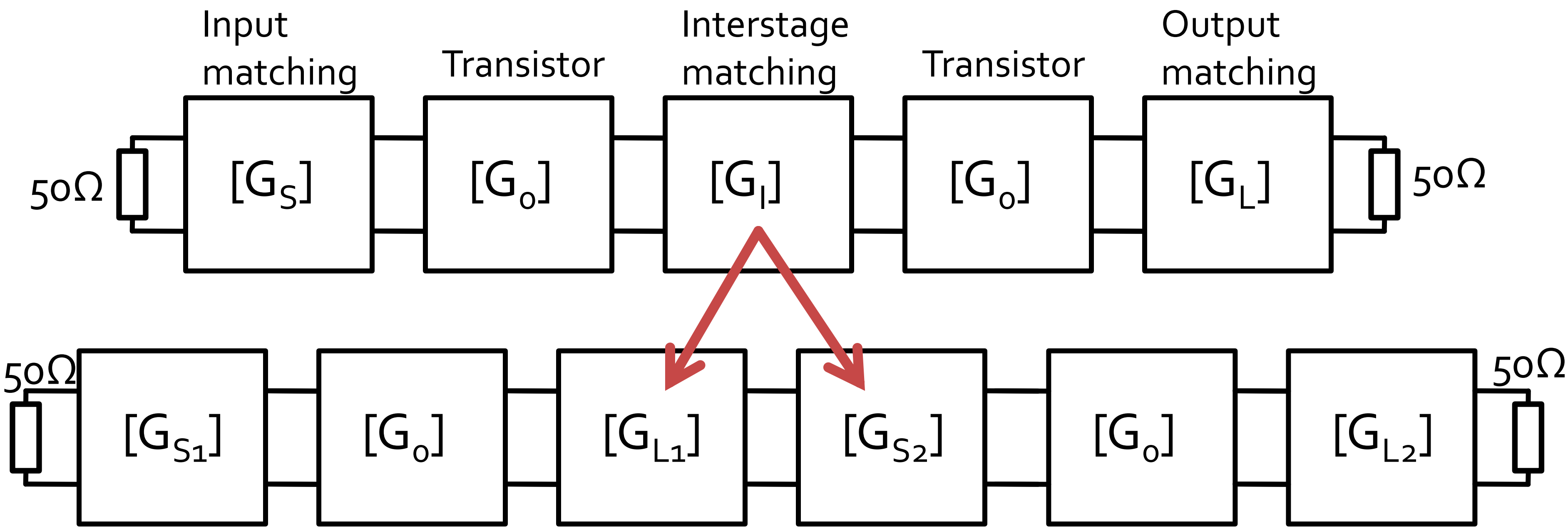
Gain =	22.00	dB
Noise Figure =	0.85	dB
Noise Temp =	62.94	°K
SNR =	63.13	dB
MDS =	-113.13	dBm
Sensitivity =	-103.13	dBm
Noise Floor =	-173.13	dBm/Hz

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

Multistage amplifier design

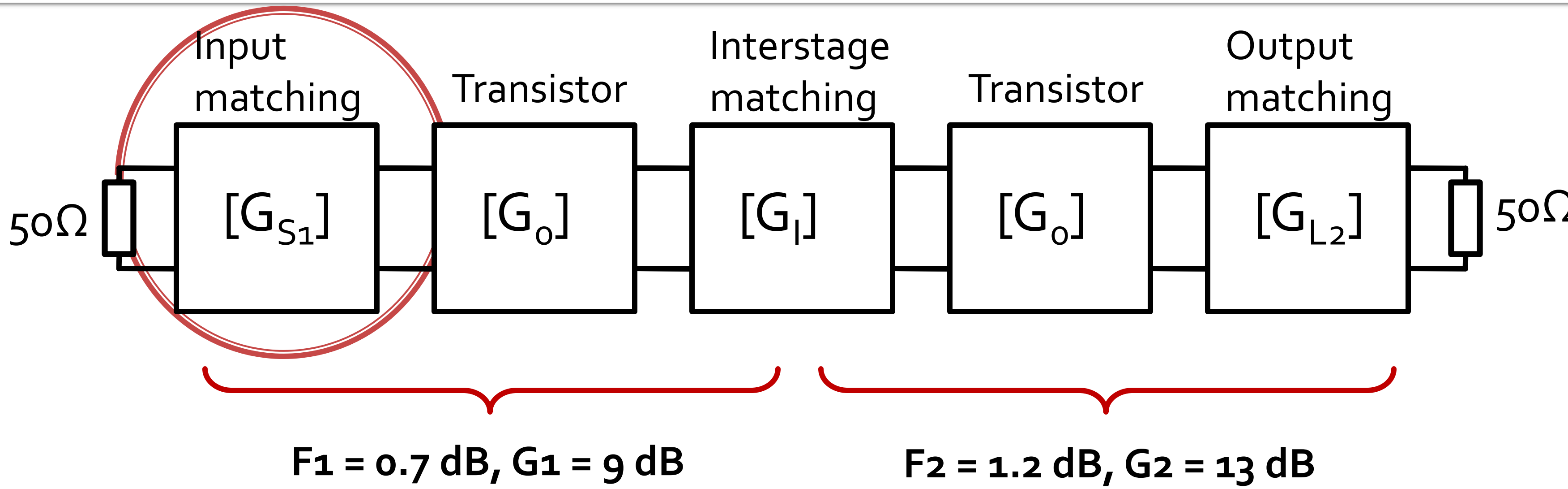
- Separation of the design parameters on the 2 amplification stages (Estimated!)
  - input stage: $F_1 = 0.7$ dB, $G_1 = 9$ dB
  - output stage: $F_2 = 1.2$ dB, $G_2 = 13$ dB
  - total: $F = 0.85$ dB, $G = 22$ dB
- Meets design specifications (with design margin)
- We can reuse some of the results in 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



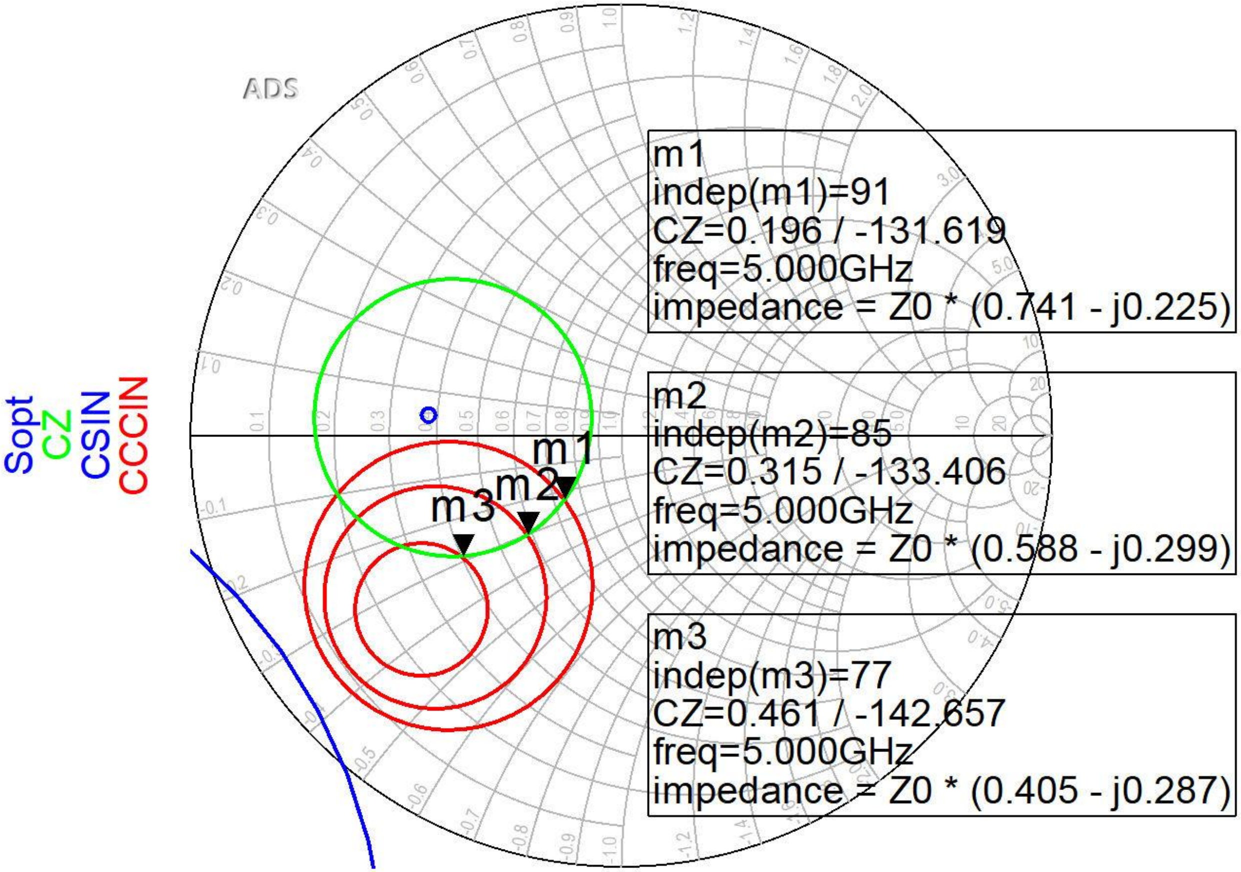
- Gain computation
  - Interstage matching can supplement the gain for both amplifier stages
  - The design for input and output matching must be achieved on a single transistor schematic (recommended: easier)

Input matching stage 1 (S1)



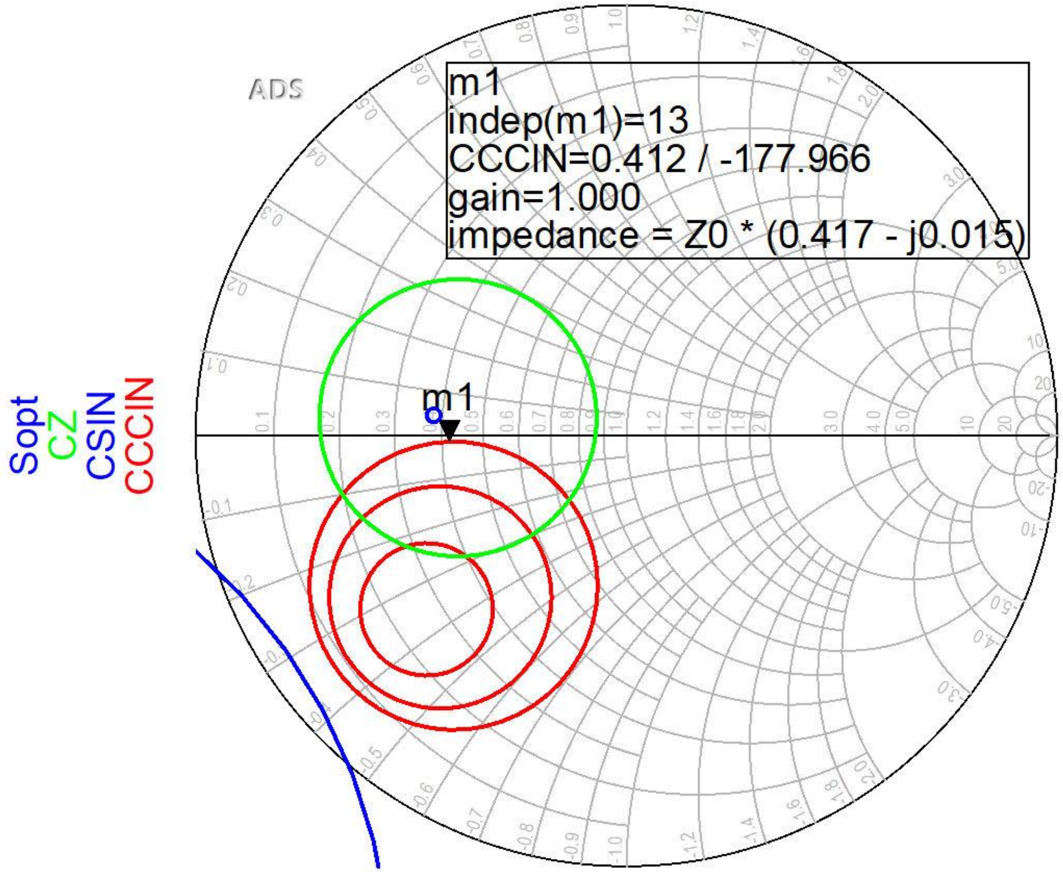
- We favor optimization for **noise** (low/minimum)
- Also considered
  - Power gain (can be lower, but not too much)
  - Bandwidth (through Q, quality factor)
  - Stability

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**

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

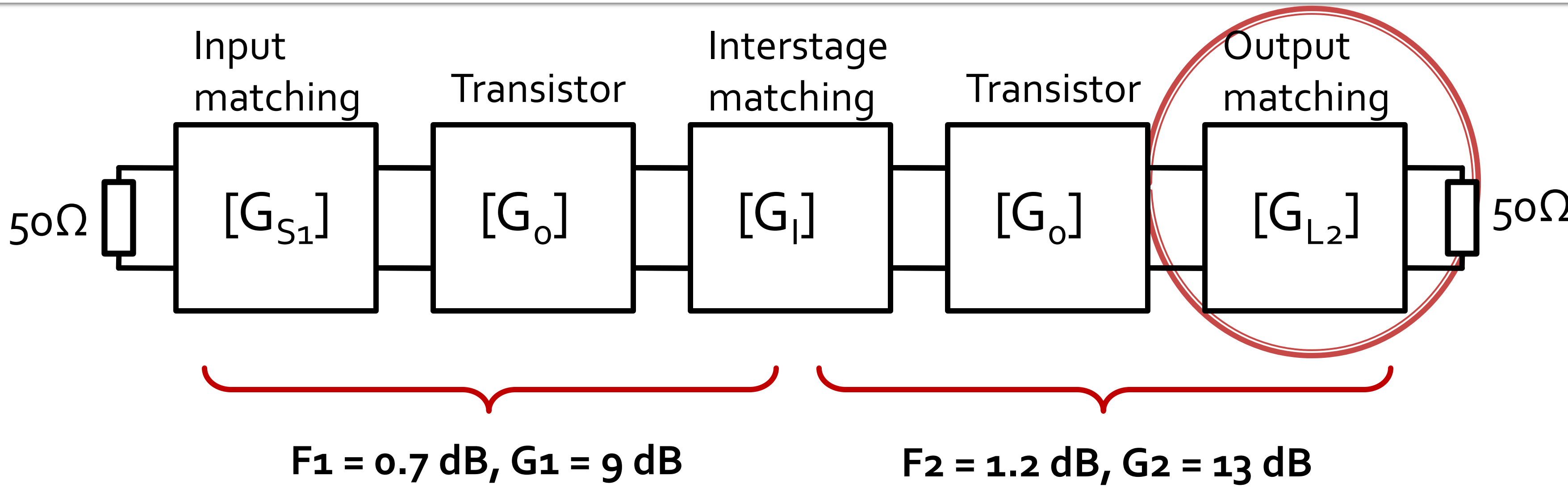
$$\cos(\varphi + 2\theta) = -|\Gamma_S| \qquad \mathrm{Im}[y_S(\theta)] = \frac{\mp 2 \cdot |\Gamma_S|}{\sqrt{1-|\Gamma_S|^2}}$$

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

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

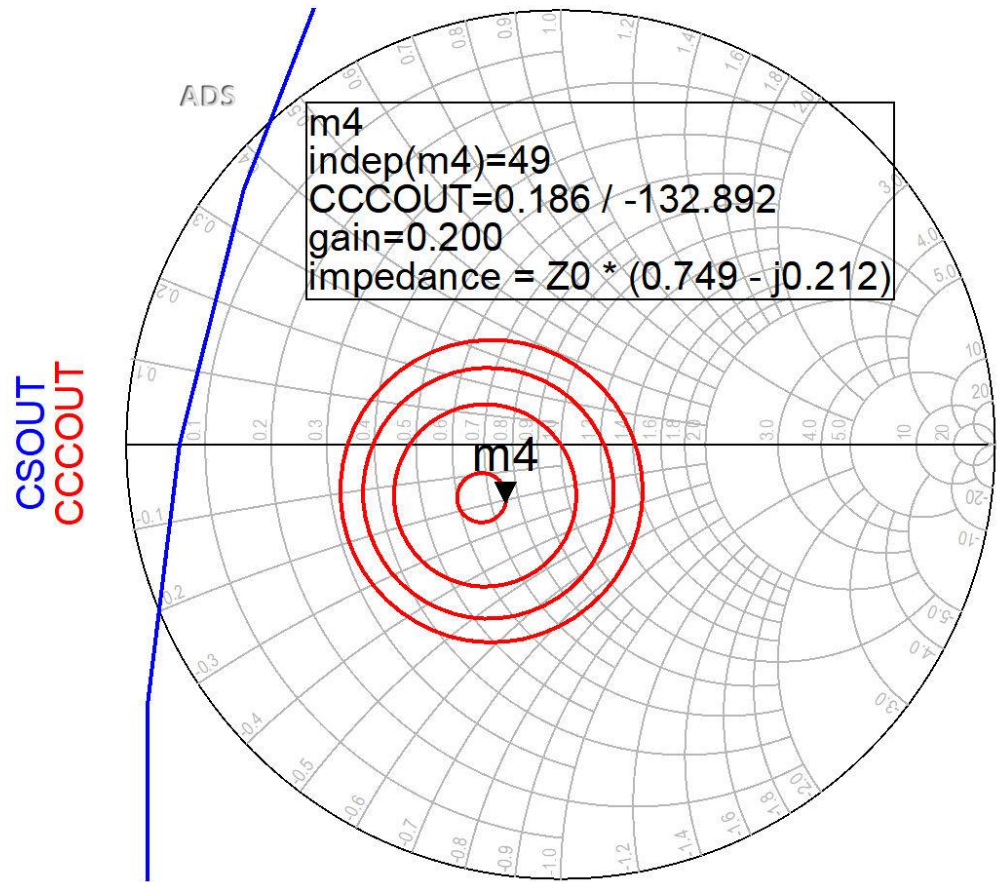
$$(\varphi + 2\theta) = \begin{cases} +114.33^\circ \\ -114.33^\circ \end{cases} \quad \theta = \begin{cases} 146.2^\circ \\ 31.8^\circ \end{cases} \quad \mathrm{Im}[y_S(\theta)] = \begin{cases} -0.904 \\ +0.904 \end{cases} \quad \theta_{sp} = \begin{cases} 137.9^\circ \\ 42.1^\circ \end{cases}$$

Output matching stage 2 (L2)



- 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_{L2}**: Position m4 in complex plane, **0.2dB**

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

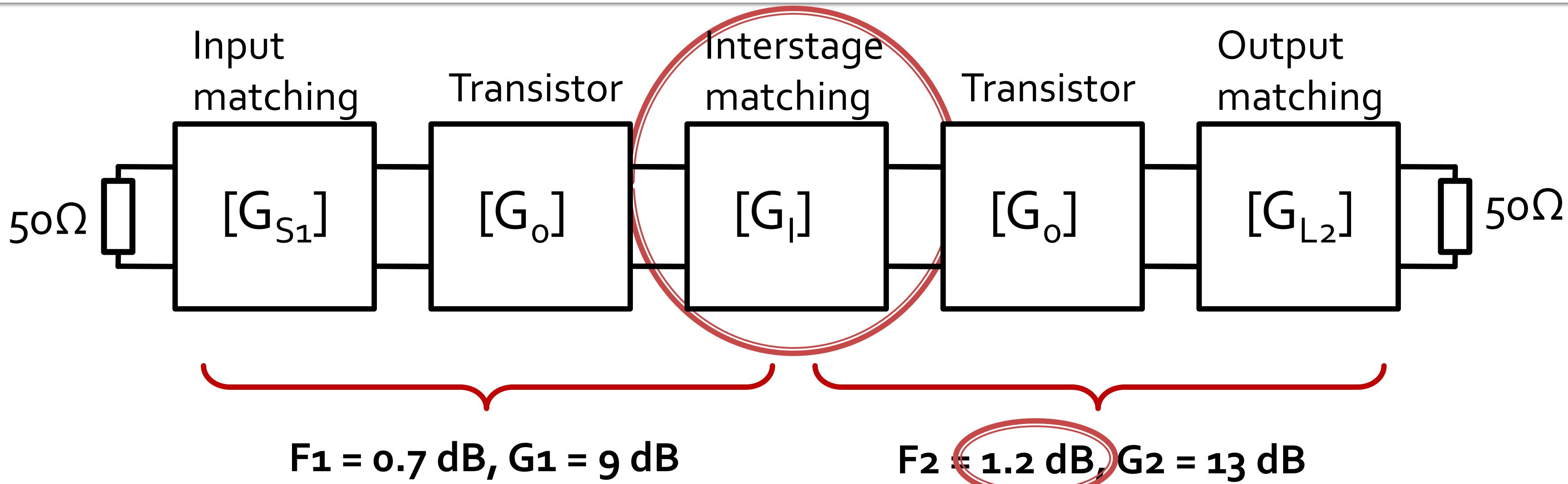
$$\cos(\varphi + 2\theta) = -|\Gamma_L| \qquad \mathrm{Im}[y_L(\theta)] = \frac{-2\cdot|\Gamma_L|}{\sqrt{1-|\Gamma_L|^2}} = -0.379$$

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

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

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

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).

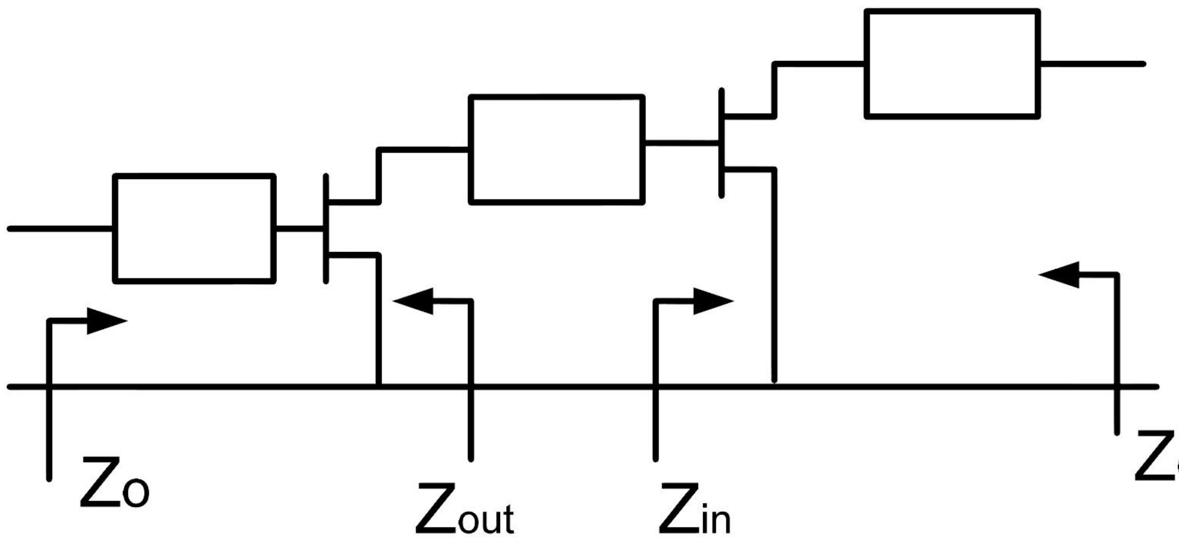
Multistage amplifier

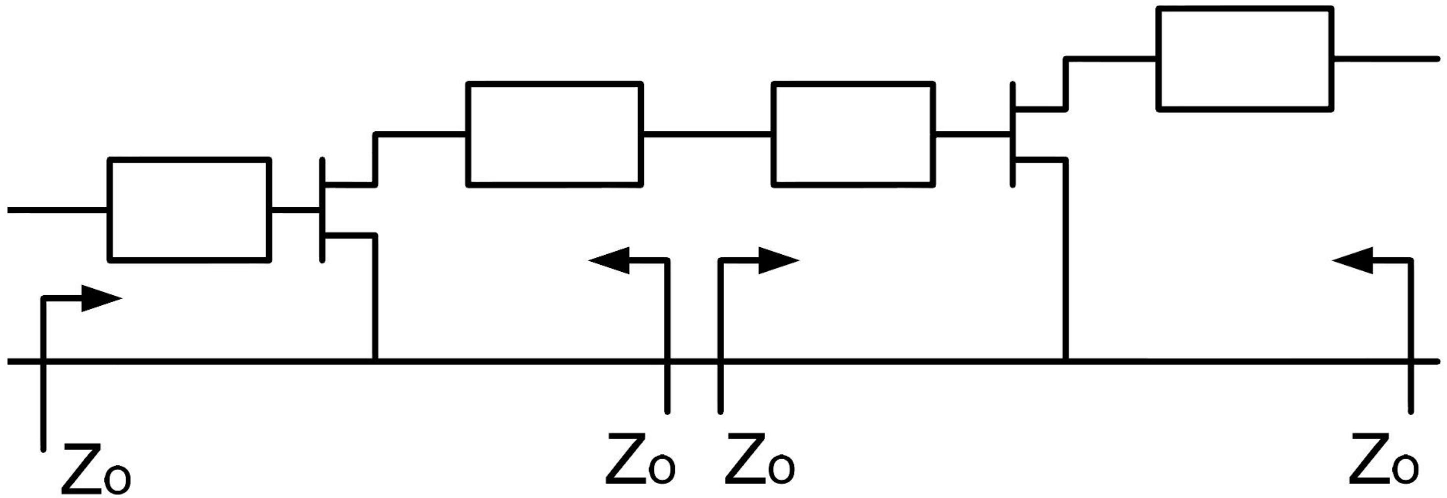
- Power gain

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

- 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

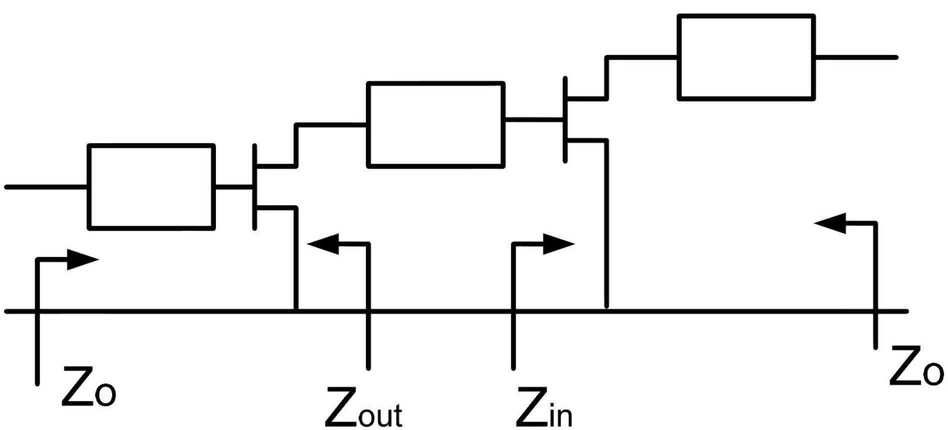
Interstage matching 1/2

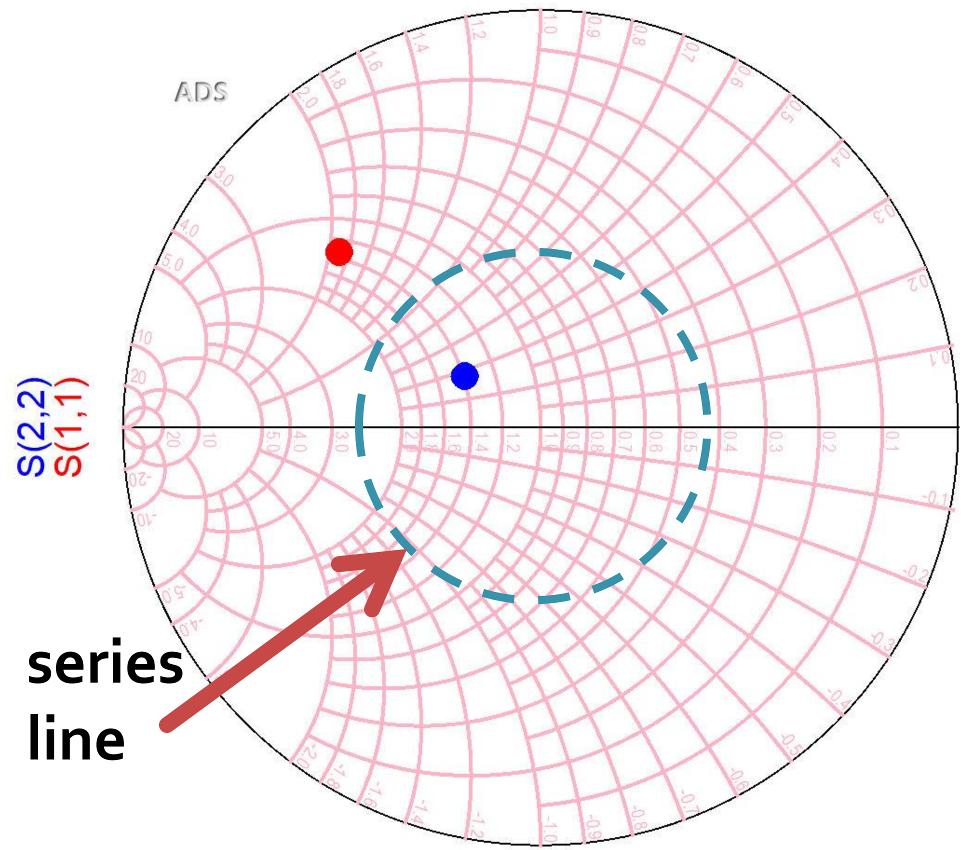




Interstage matching 1

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



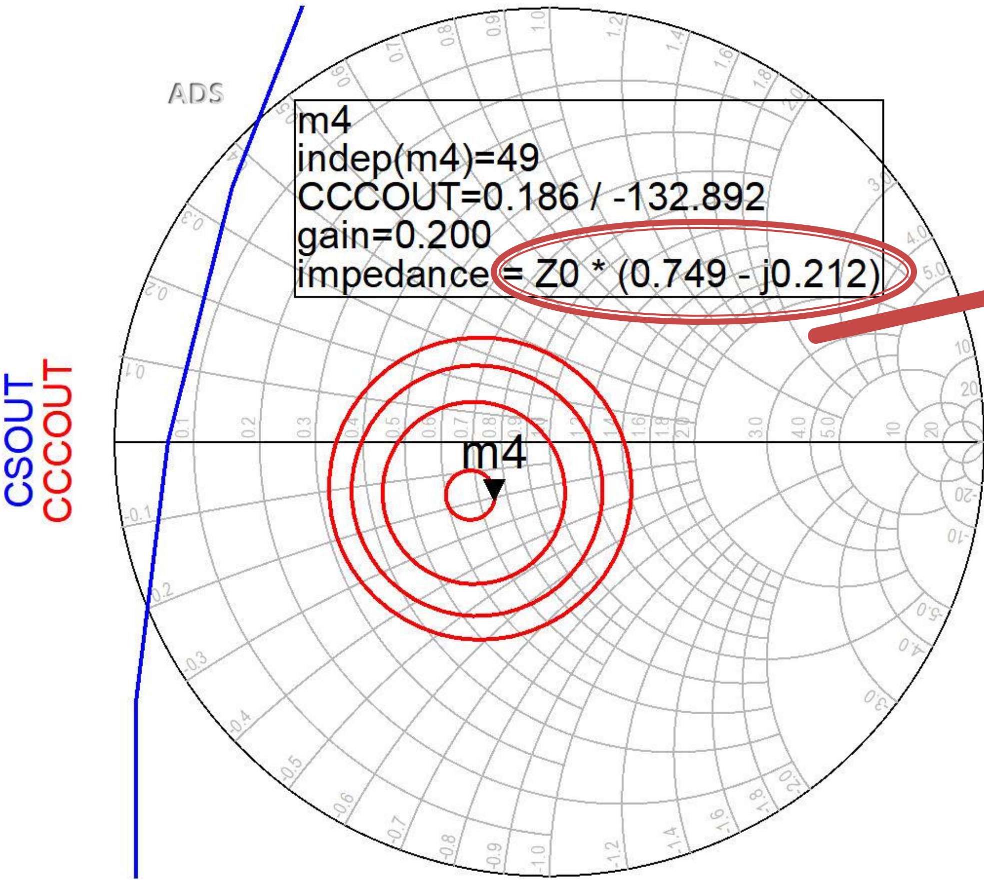


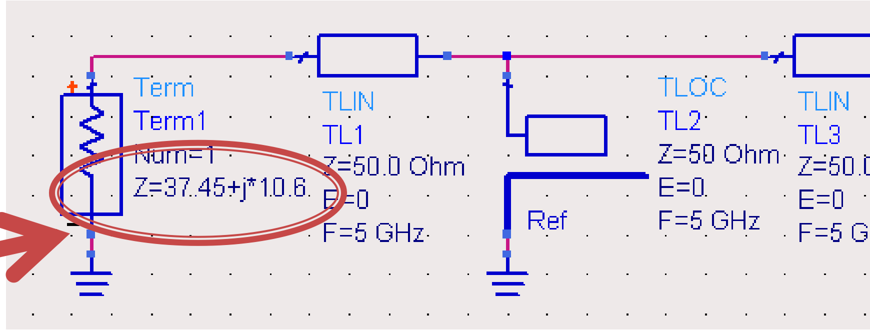
Interstage matching 1

- Can be designed in two ways:
  - starting from the output of the first stage (reflection coefficient S_{22}^*) towards the circles (drawn for the second stage):
    - stability
    - gain
    - noise
  - starting from the input of the second stage (reflection coefficient S_{11}^*) 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





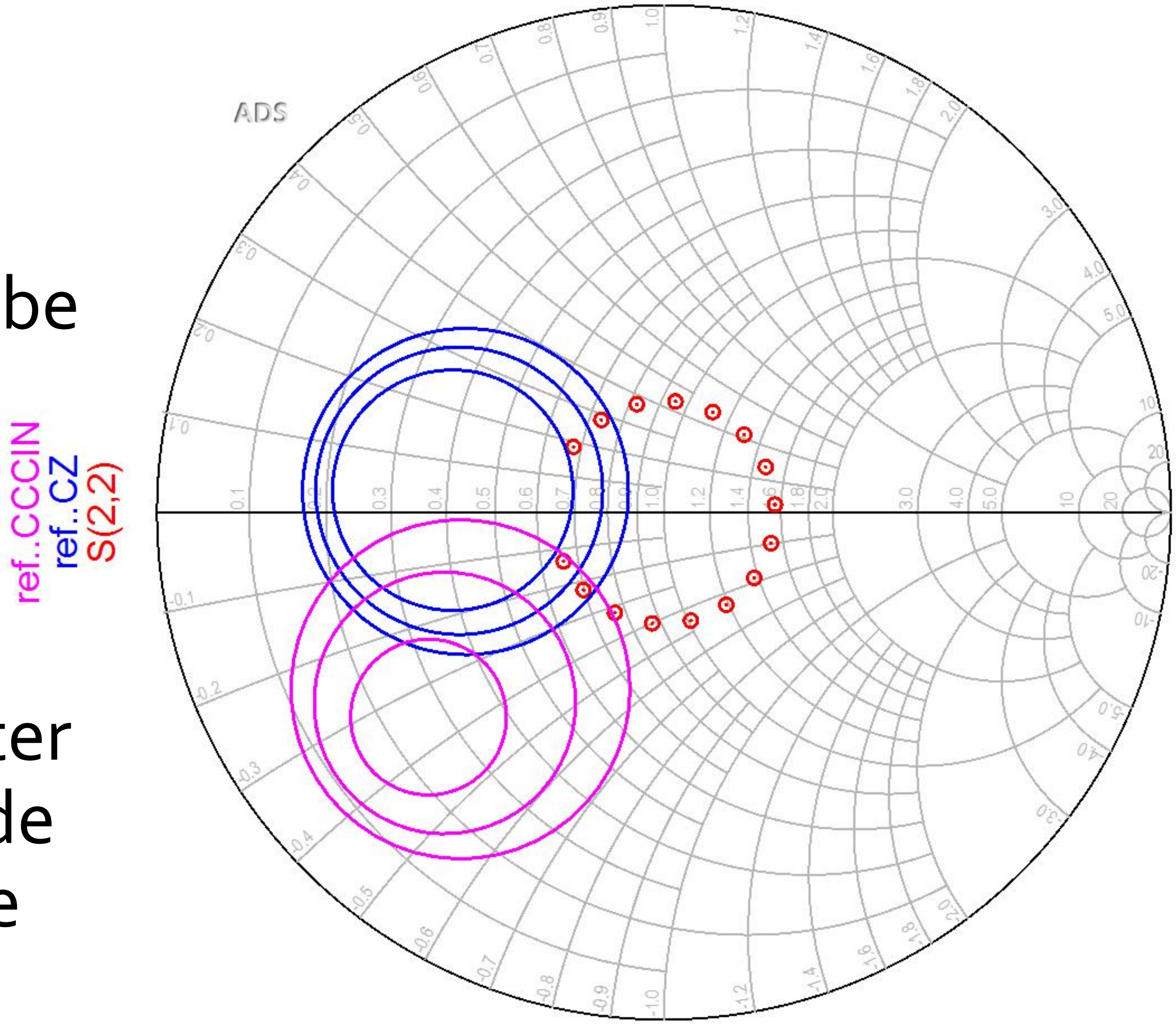
$$Z = 50\Omega \cdot (0.749 - j \cdot 0.212)$$

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

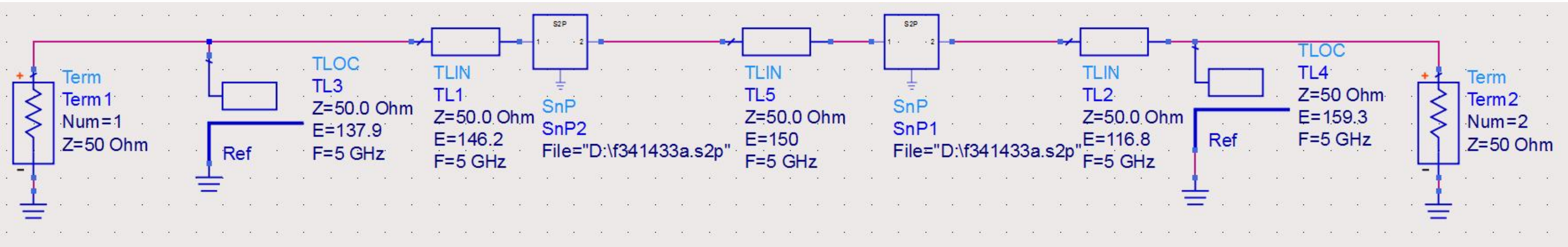
$$Z^* = 37.45\Omega + j \cdot 10.6\Omega$$

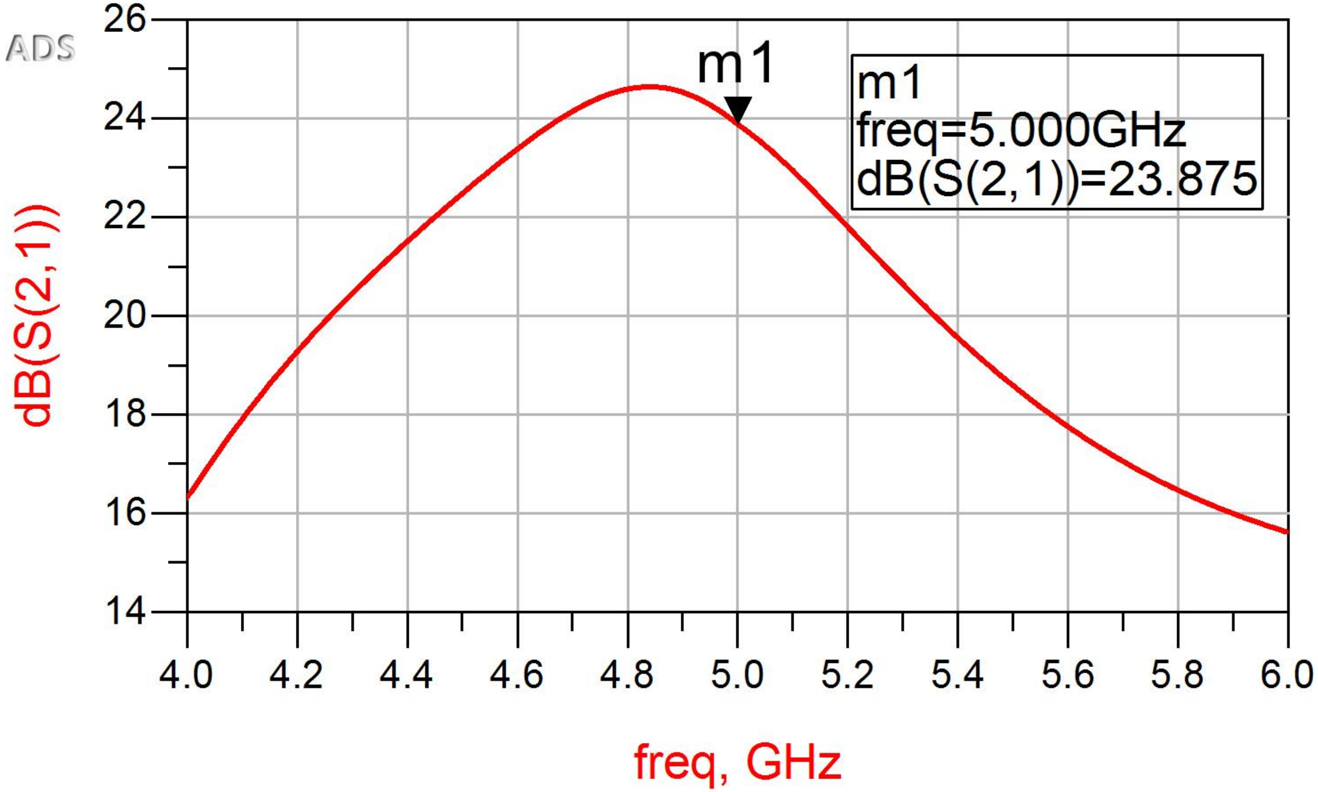
Interstage matching 1

- 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

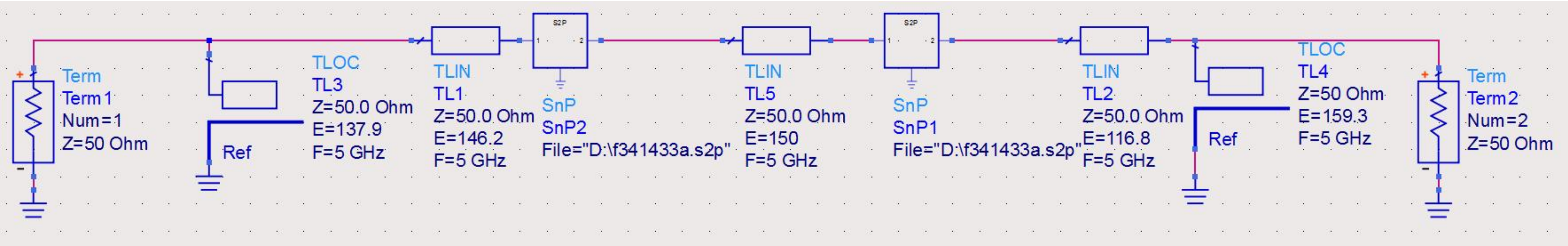


ADS

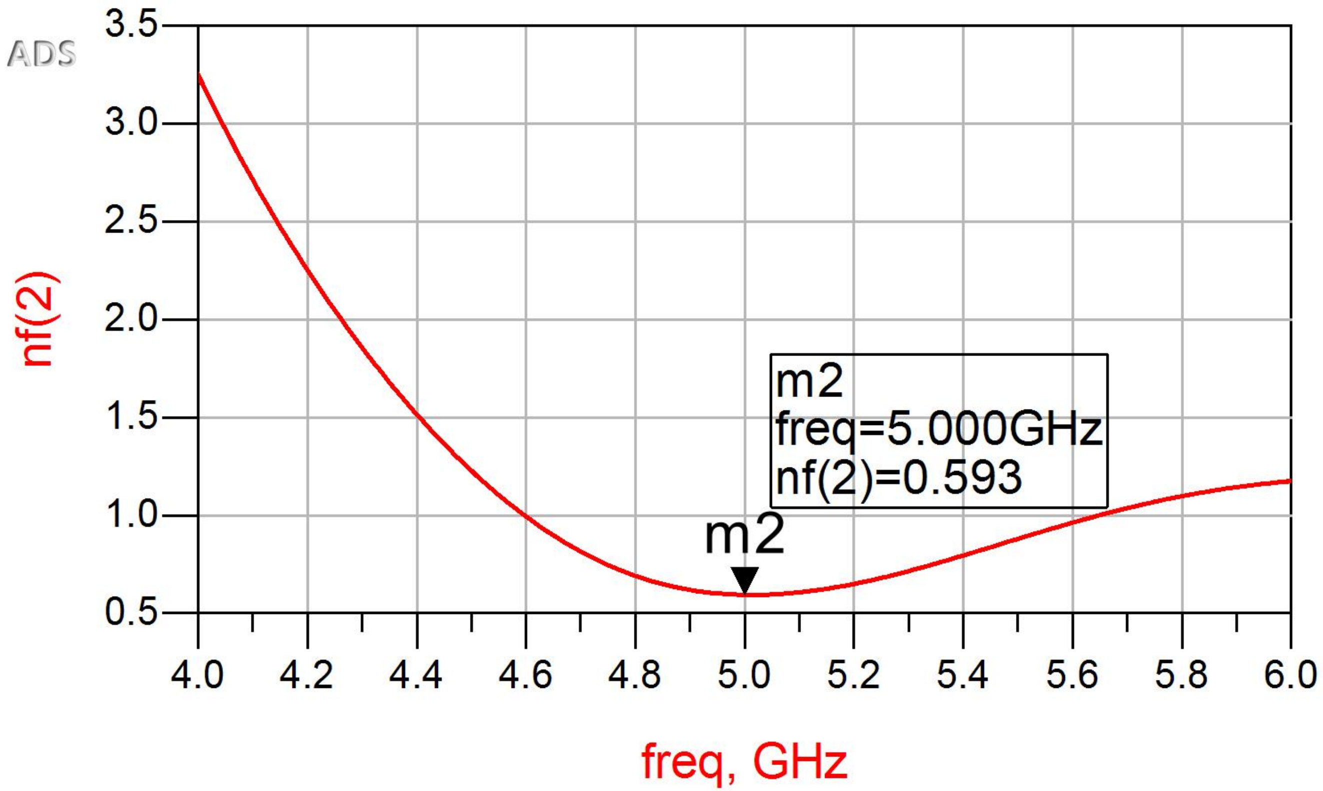

Term
Term1
Num=1
Z=50 Ohm
Ref
TLOC
TL3
Z=50.0 Ohm
E=137.9
F=5 GHz
TLIN
TL1
Z=50.0 Ohm
E=146.2
F=5 GHz
SnP
SnP2
File="D:\f341433a.s2p"
TLIN
TL5
Z=50.0 Ohm
E=150
F=5 GHz
SnP
SnP1
File="D:\f341433a.s2p"
TLIN
TL2
Z=50.0 Ohm
E=116.8
F=5 GHz
Ref
TLOC
TL4
Z=50 Ohm
E=159.3
F=5 GHz
Term
Term2
Num=2
Z=50 Ohm


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



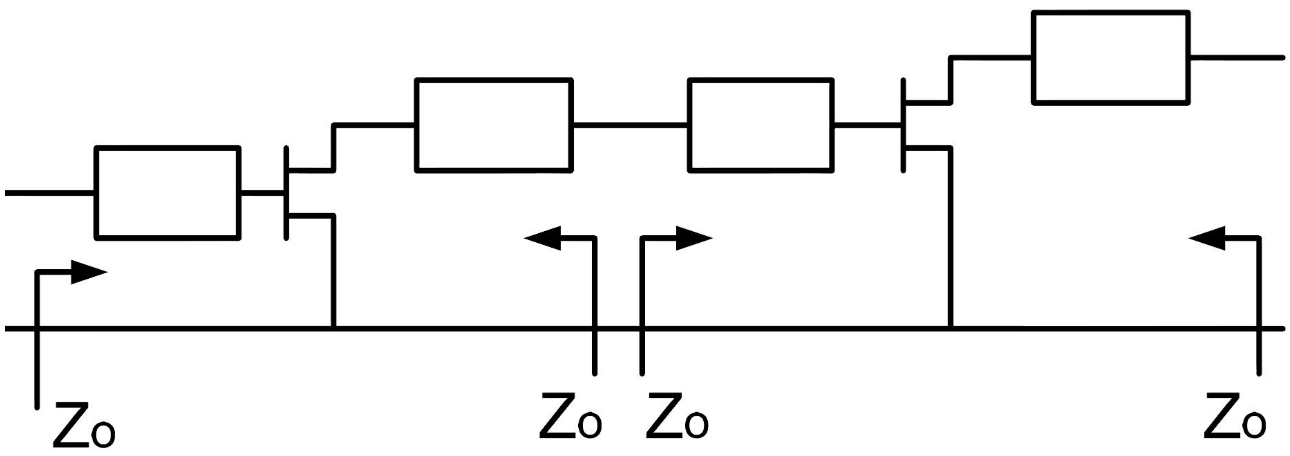
Term
Term1
Num=1
Z=50 Ohm
Ref
TLOC
TL3
Z=50.0 Ohm
E=137.9
F=5 GHz
TLIN
TL1
Z=50.0 Ohm
E=146.2
F=5 GHz
SnP
SnP2
File="D:\f341433a.s2p"
TLIN
TL5
Z=50.0 Ohm
E=150
F=5 GHz
SnP
SnP1
File="D:\f341433a.s2p"
TLIN
TL2
Z=50.0 Ohm
E=116.8
F=5 GHz
Ref
TLOC
TL4
Z=50 Ohm
E=159.3
F=5 GHz
Term
Term2
Num=2
Z=50 Ohm



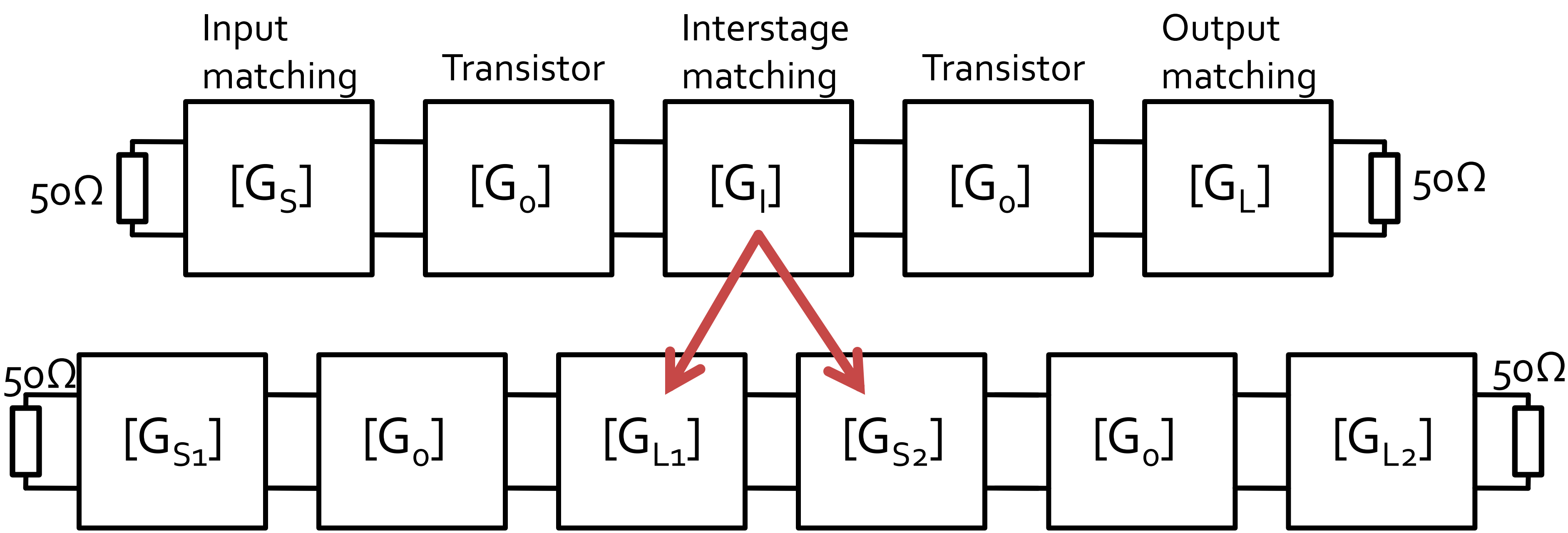
ADS
nf(2)
3.5
3.0
2.5
2.0
1.5
1.0
0.5
m2
freq=5.000GHz
nf(2)=0.593
m2
4.0
4.2
4.4
4.6
4.8
5.0
5.2
5.4
5.6
5.8
6.0
freq, GHz

Interstage matching 2

- Using multiple transmission lines for matching each stage to a intermediate Γ=o (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

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

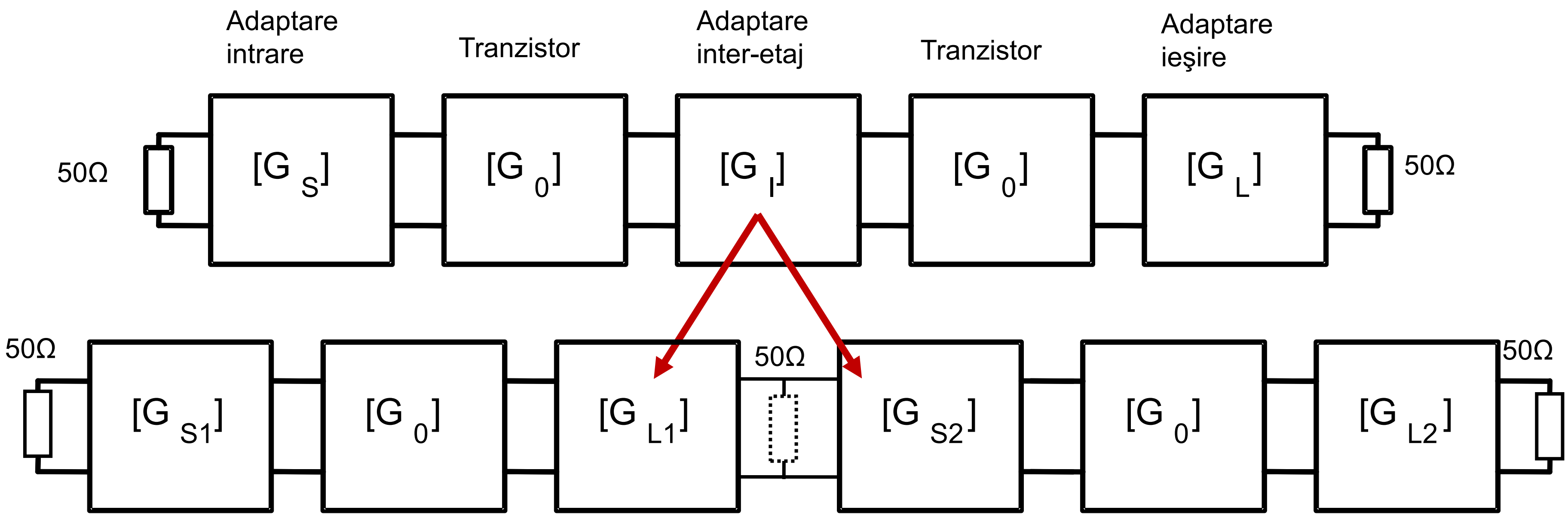
$$G_T[dB] = 1\ dB + 10\ dB + G_{L1}[dB] + G_{S2}[dB] + 10\ dB + 0.2\ dB$$

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

- 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

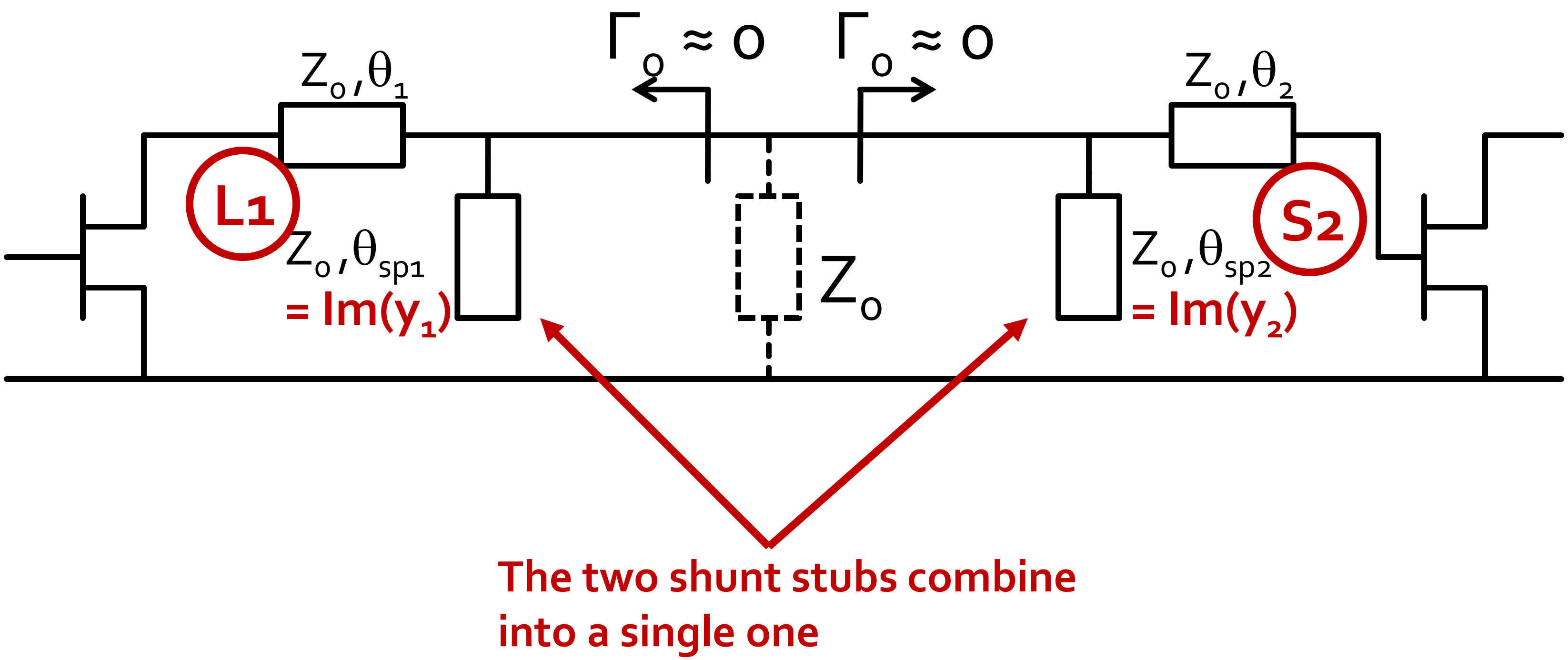
Interstage matching 2

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



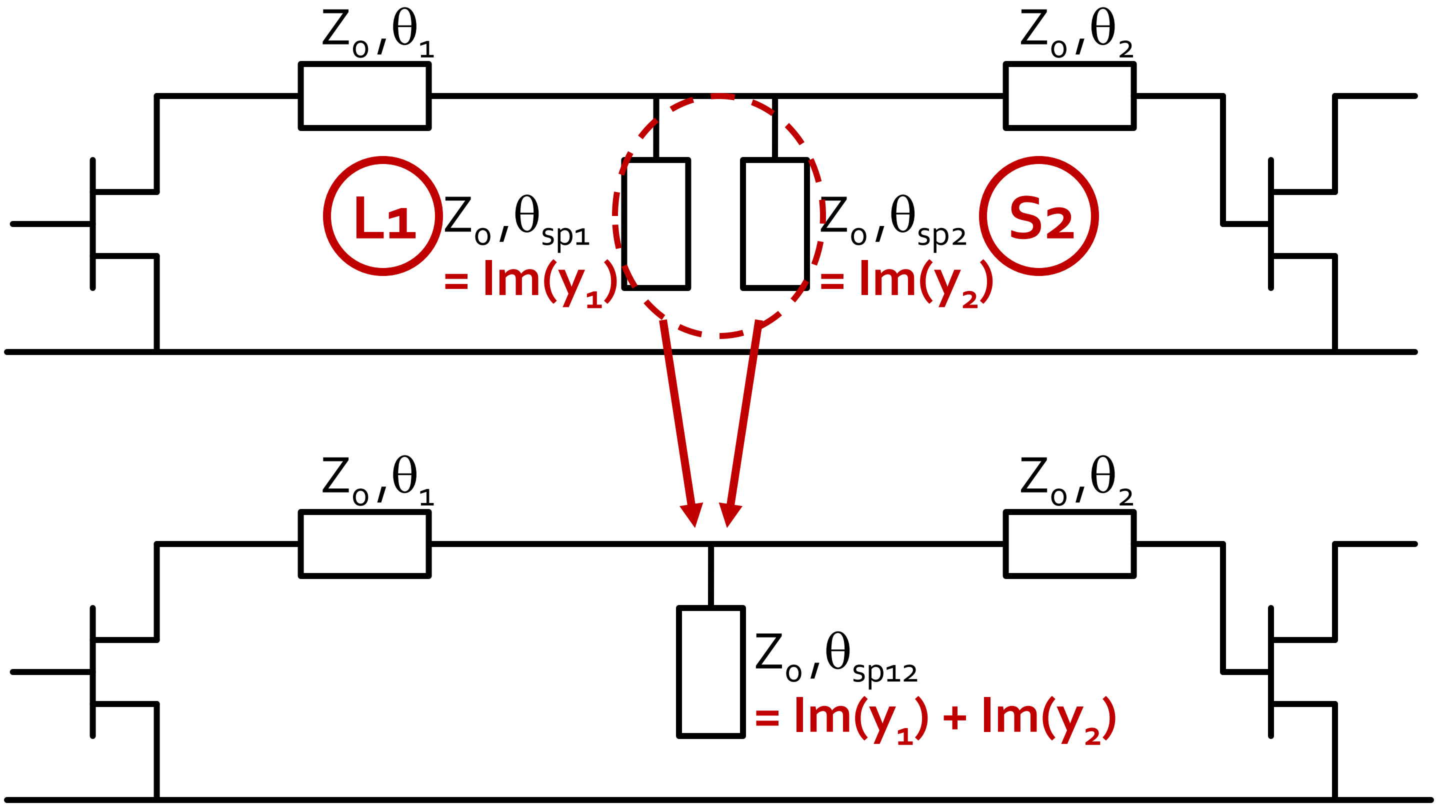
Interstage matching 2

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



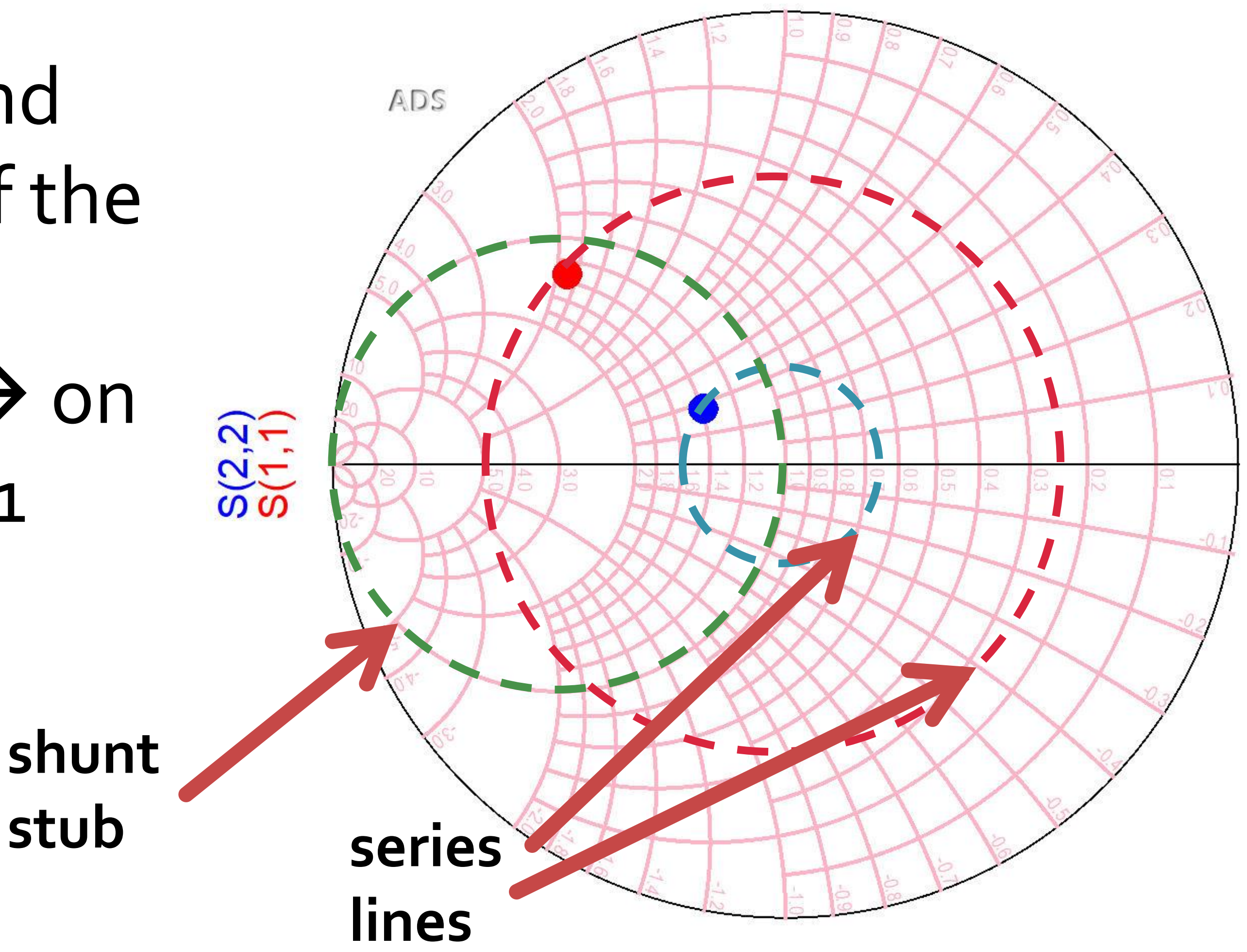
Interstage matching 2

- The two shunt stubs combine into a single one



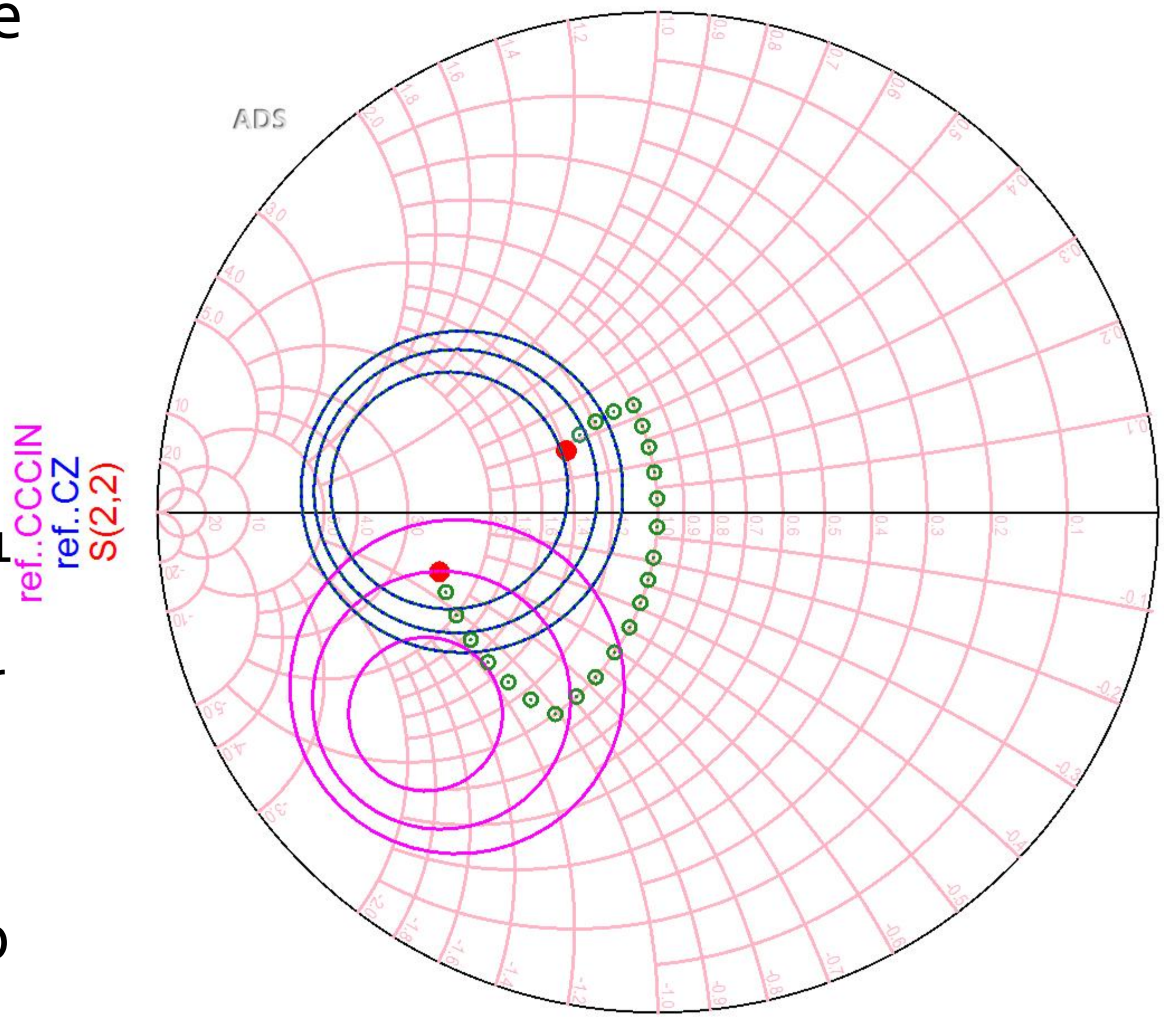
Interstage matching 2

- 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**

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

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

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

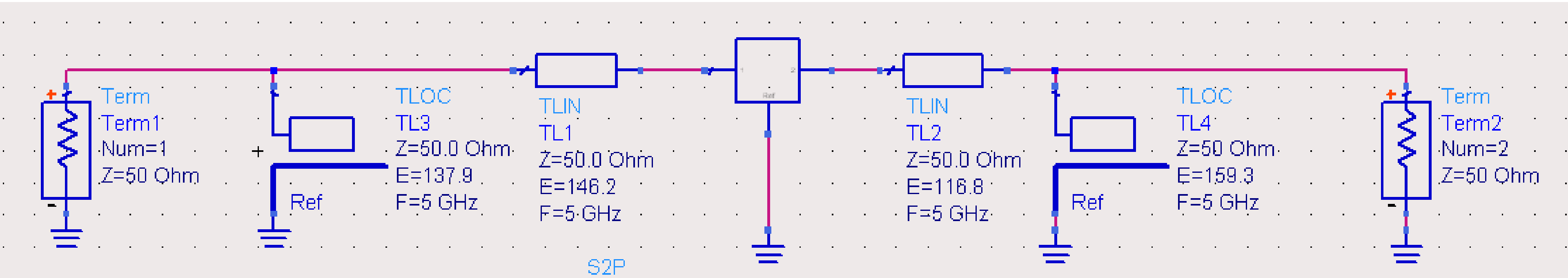
- the length of the shunt stub θ_{sp} is not calculated because it is **not** needed

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

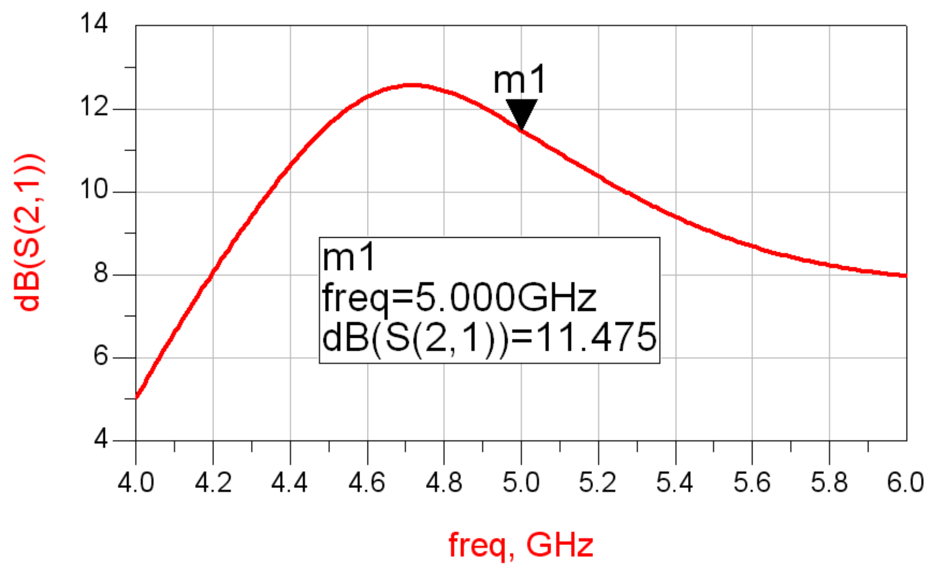
Output matching stage 1 (L1)

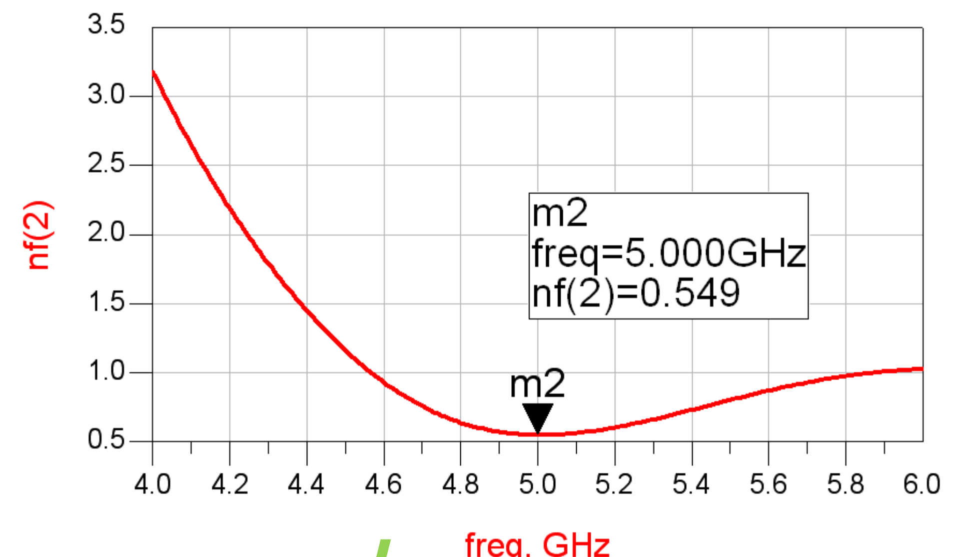
Equation	**Solution L1A**	**Solution L1B**
$\Phi+2\theta$	$+100.72°$	$-100.72°$
θ	$116.8°$	$16.1°$
$\mathrm{Im}[y(\theta)]$	-0.379	$+0.379$

Verify stage 1



G_{S1} G_o G_{L1}

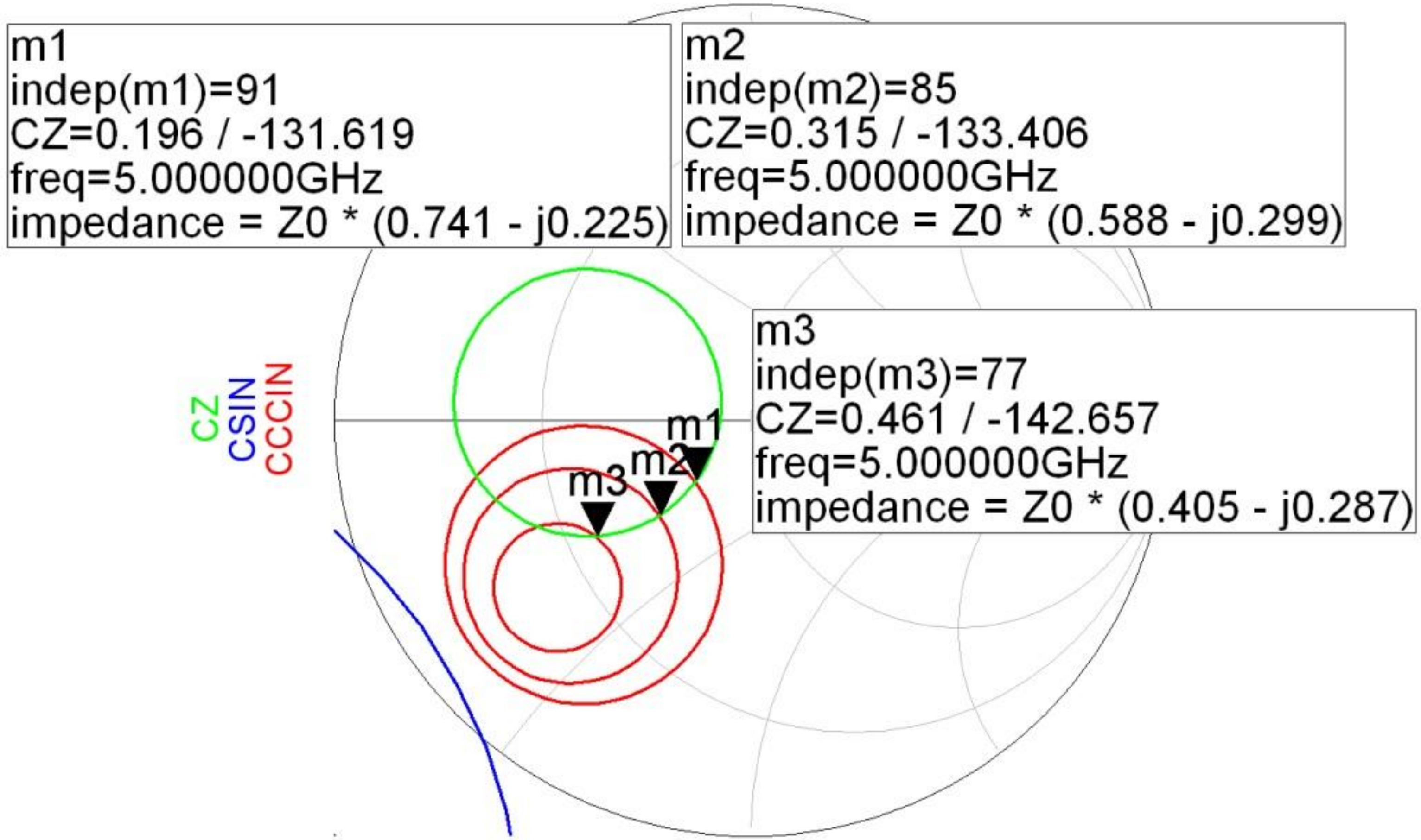




F1 < 0.7 dB, G1 > 9 dB √

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**

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

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

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

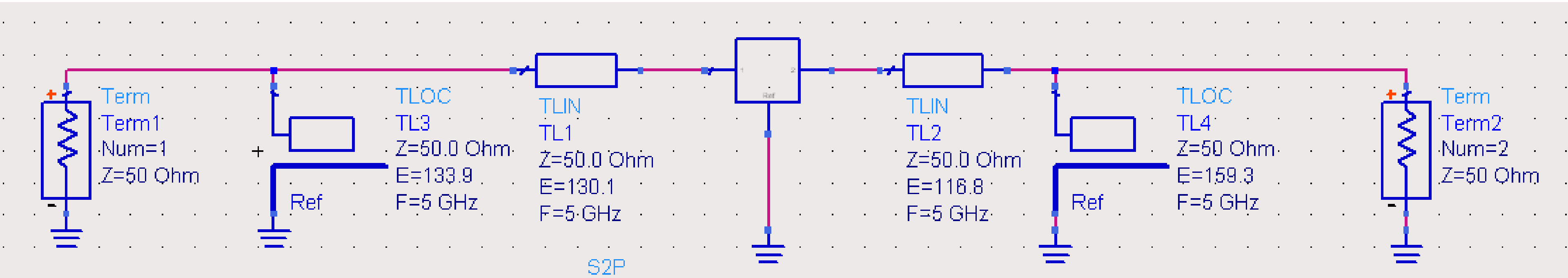
- the length of the shunt stub θ_{sp} is not calculated because it is **not** needed

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

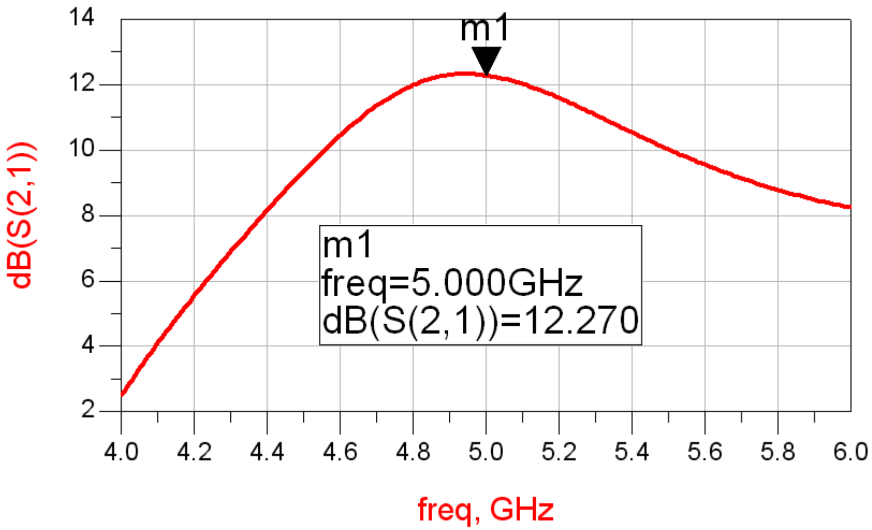
Input matching stage 2 (S2)

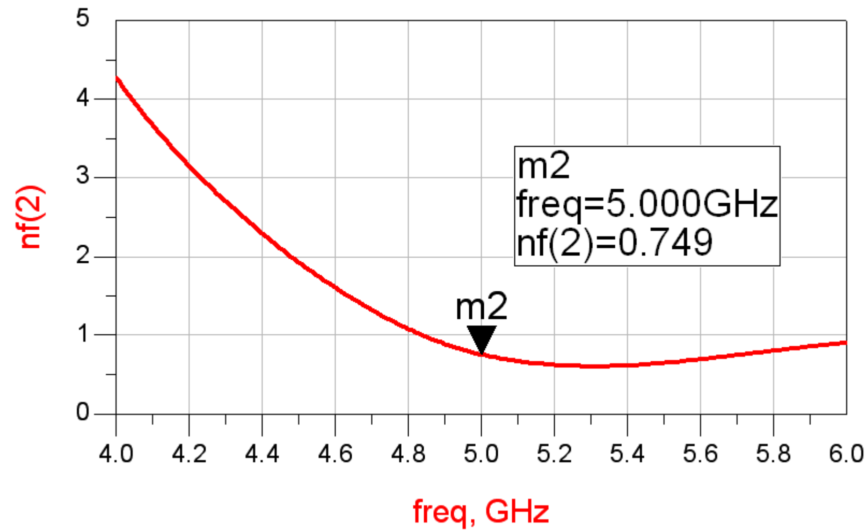
Equation	Solution S2A	Solution S2B
$\Phi+2\theta$	$+117.45°$	$-117.45°$
θ	$130.1°$	$12.6°$
$\text{Im}[y(\theta)]$	-1.039	$+1.039$

Verify stage 2



G_{S1} G_o G_{L1}

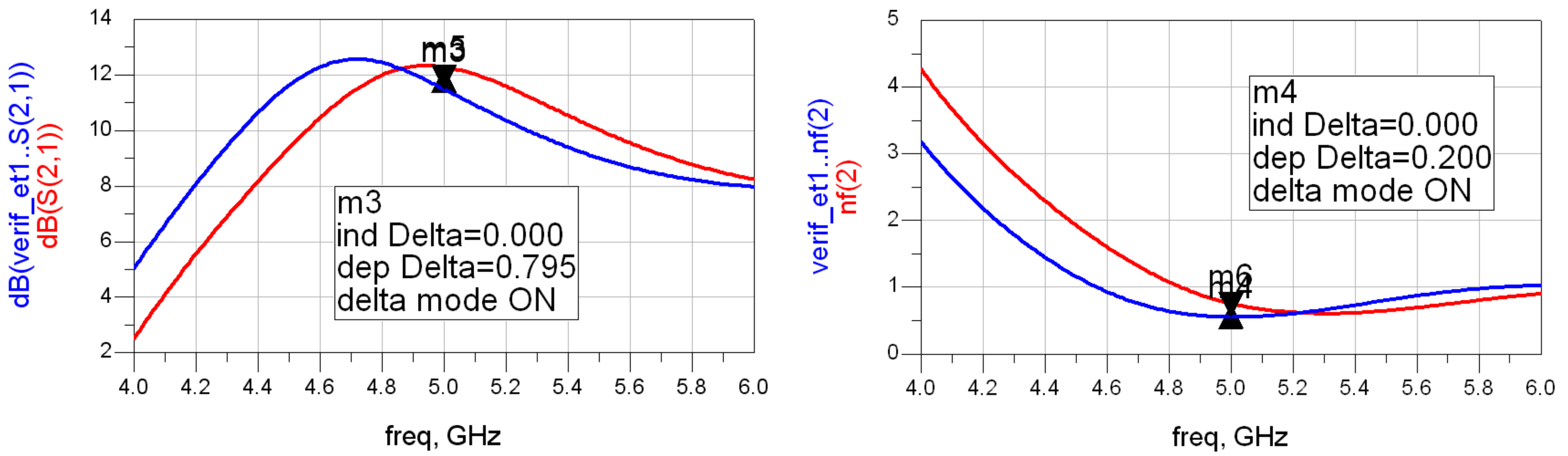




F2 < 1.2 dB, G2 < 13 dB X G1 = 11.5dB,G2 = 12.3dB,G1+G2 > 22dB √

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

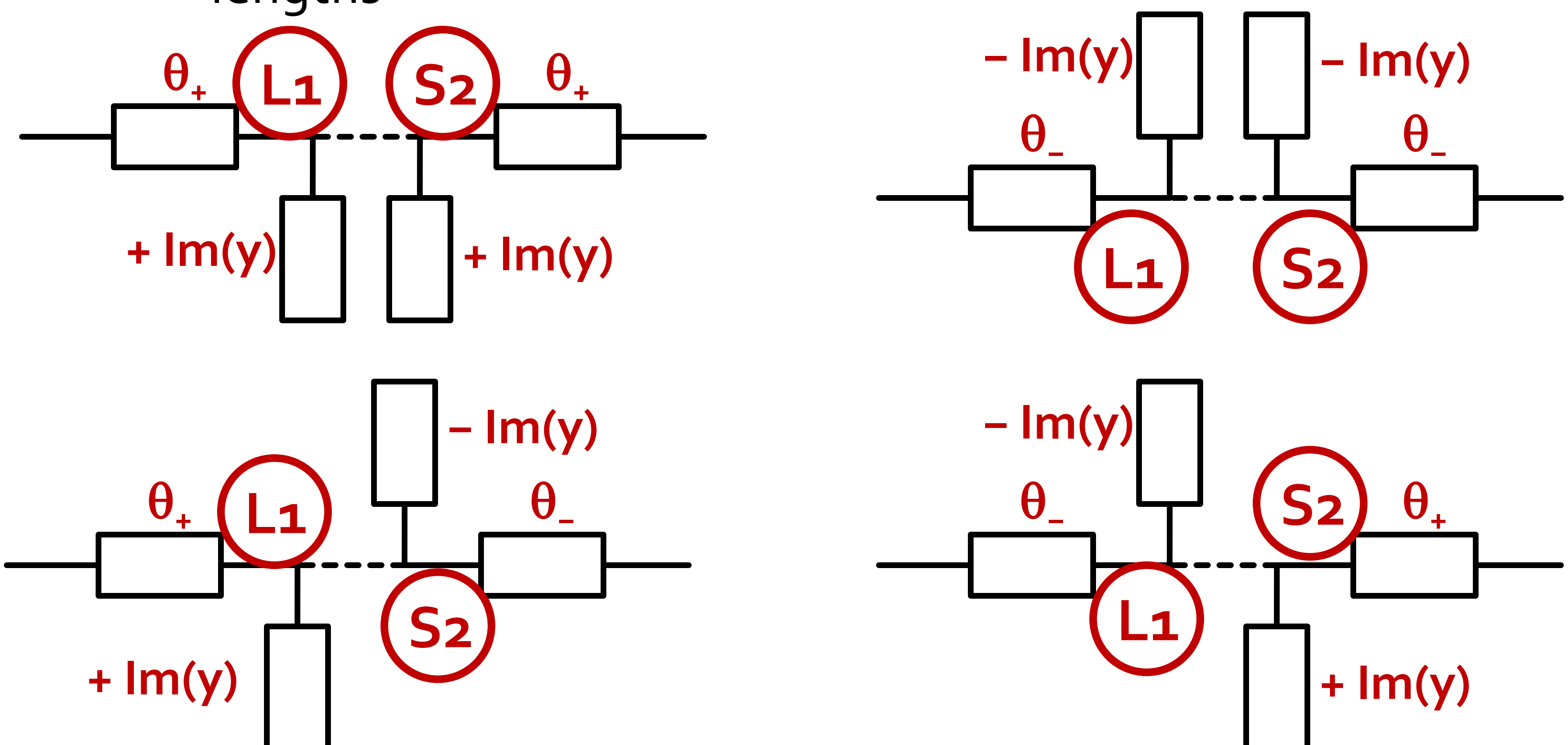
- **The two shunt stubs merge into a single one**
- There are **4 possible combinations** depending on how we chose the electrical length for the two series lines
  - for each chosen electric length (θ) the corresponding Im[y(θ)] must be used
- Ex:

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

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

Merging the two shunt stubs

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



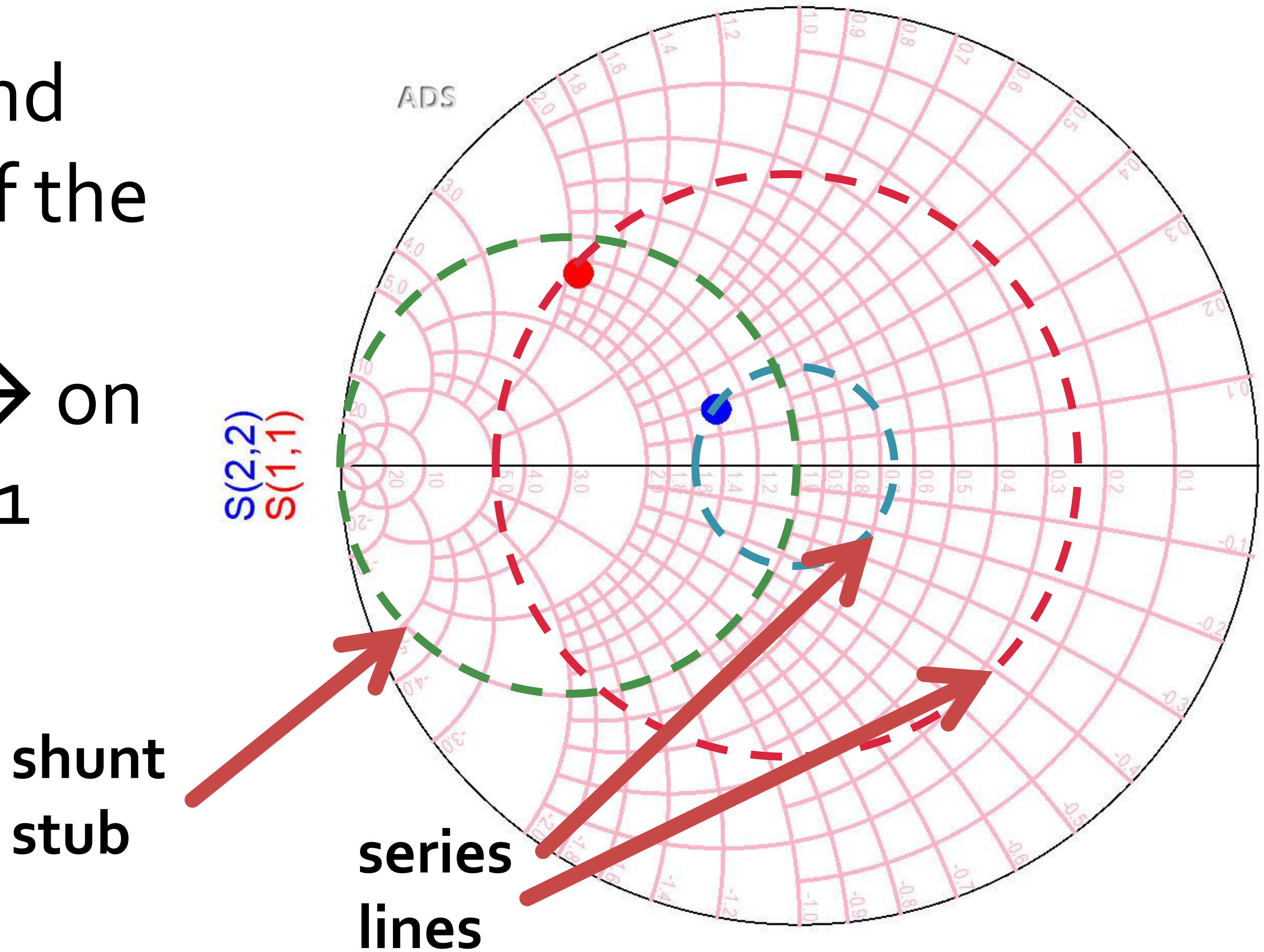
$$\mathrm{Im}\left[y_{sp}\right] = \mathrm{Im}\left[y_{L1}(\theta)\right] + \mathrm{Im}\left[y_{S2}(\theta)\right] \qquad \theta_{sp} = \tan^{-1}\left(\mathrm{Im}\left[y_{sp}\right]\right)$$

Merging the two shunt stubs

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

Smith Chart

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



Merge 1, Smith Chart

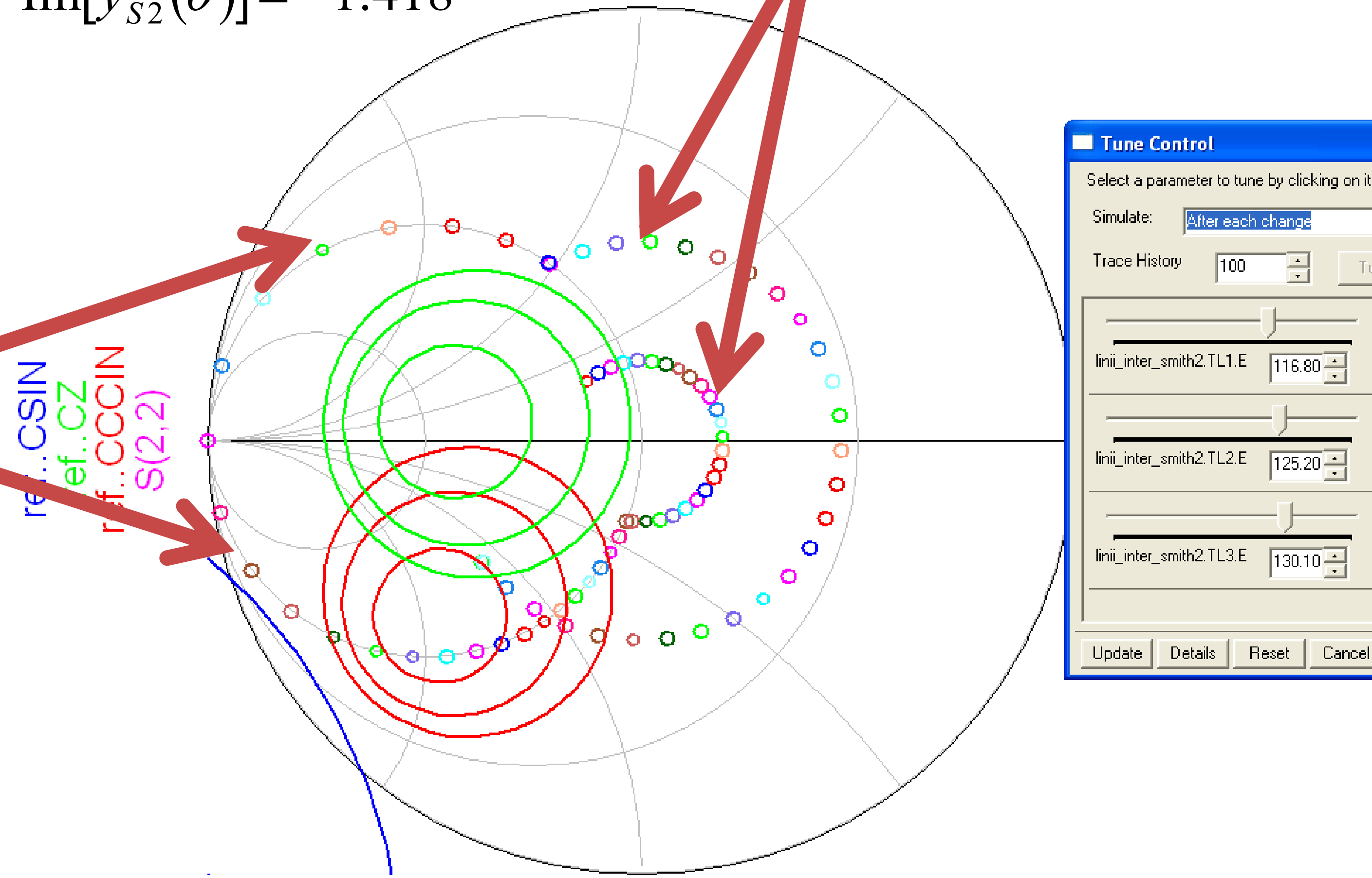
$$\theta_{L1} = 116.8^\circ \qquad \theta_{S2} = 130.1^\circ$$

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

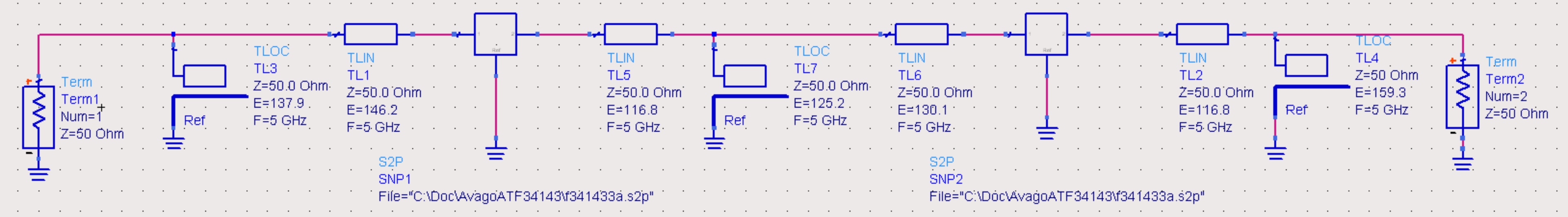
$$\theta_{sp} = 125.2^\circ$$

series lines

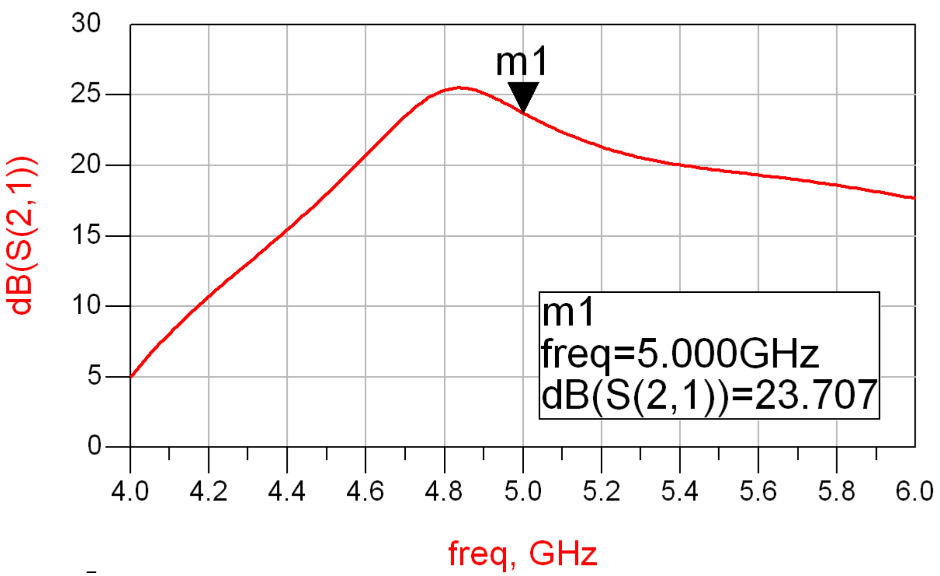
“combined” stub



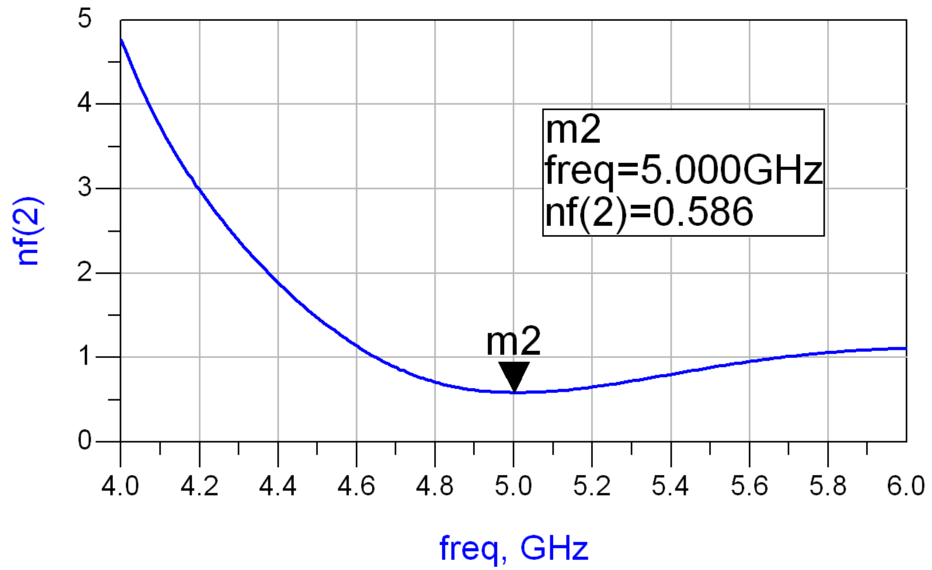
Merge 1, ADS



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))
freq, GHz



m2
freq=5.000GHz
nf(2)=0.586
m2
nf(2)
freq, GHz

Merge 2, Smith Chart

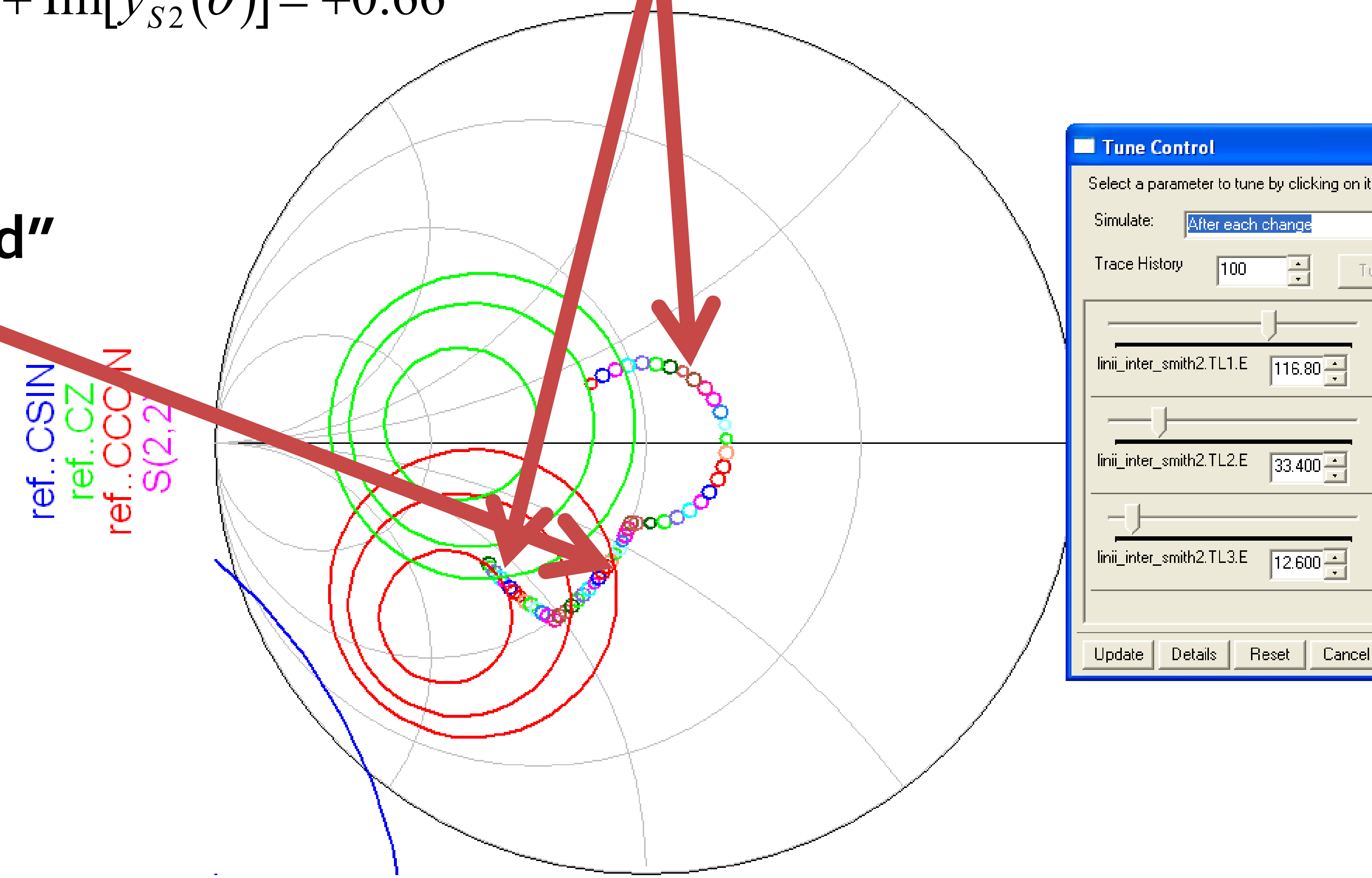
$$\theta_{L1} = 116.8° \qquad \theta_{S2} = 12.6°$$

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

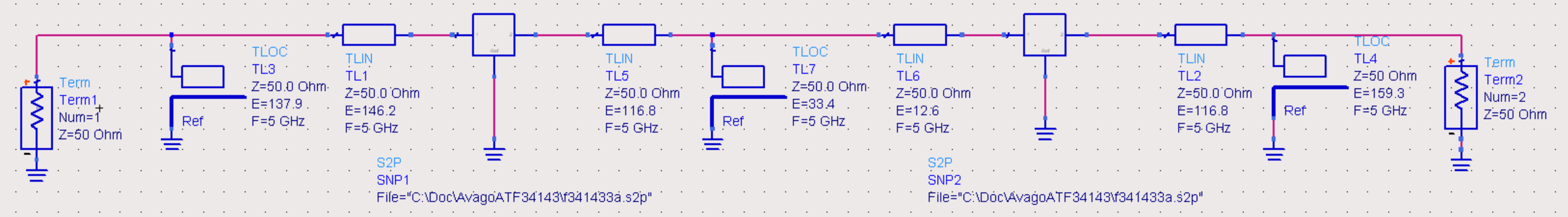
$$\theta_{sp} = 33.4°$$

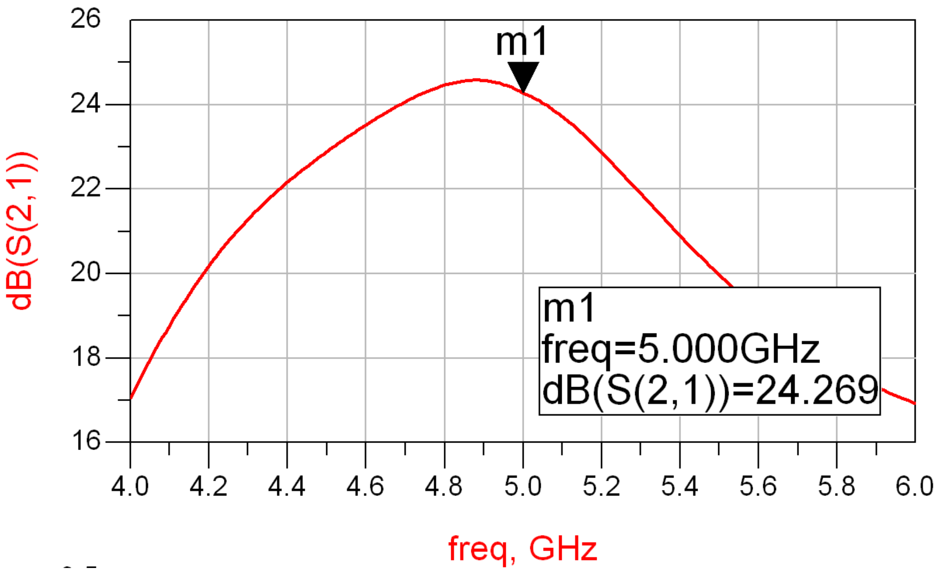
"combined" stub

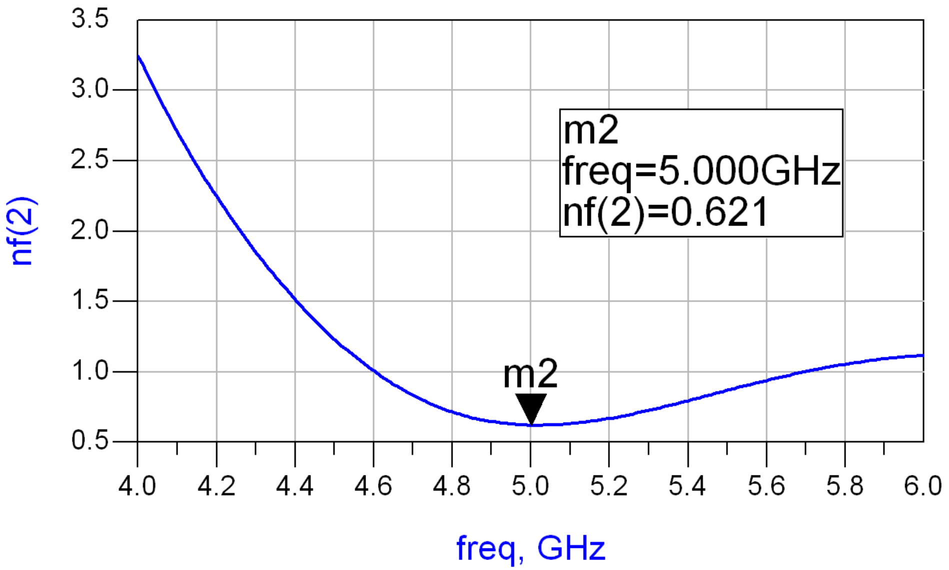
series lines



Merge 2, ADS


Term
Term1
Num=1
Z=50 Ohm
TLOC
TL3
Z=50.0 Ohm
E=137.9
F=5 GHz
Ref
TLIN
TL1
Z=50.0 Ohm
E=146.2
F=5 GHz
S2P
SNP1
File="C:\Doc\AvagoATF34143\f341433a.s2p"
TLIN
TL5
Z=50.0 Ohm
E=116.8
F=5 GHz
TLOC
TL7
Z=50.0 Ohm
E=33.4
F=5 GHz
Ref
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
TLOC
TL4
Z=50 Ohm
E=159.3
F=5 GHz
Ref
Term
Term2
Num=2
Z=50 Ohm


m1
m1
freq=5.000GHz
dB(S(2,1))=24.269
dB(S(2,1))
freq, GHz
16
18
20
22
24
26
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)=0.621
m2
nf(2)
freq, GHz
0.5
1.0
1.5
2.0
2.5
3.0
3.5
4.0
4.2
4.4
4.6
4.8
5.0
5.2
5.4
5.6
5.8
6.0

Merge 3, Smith Chart

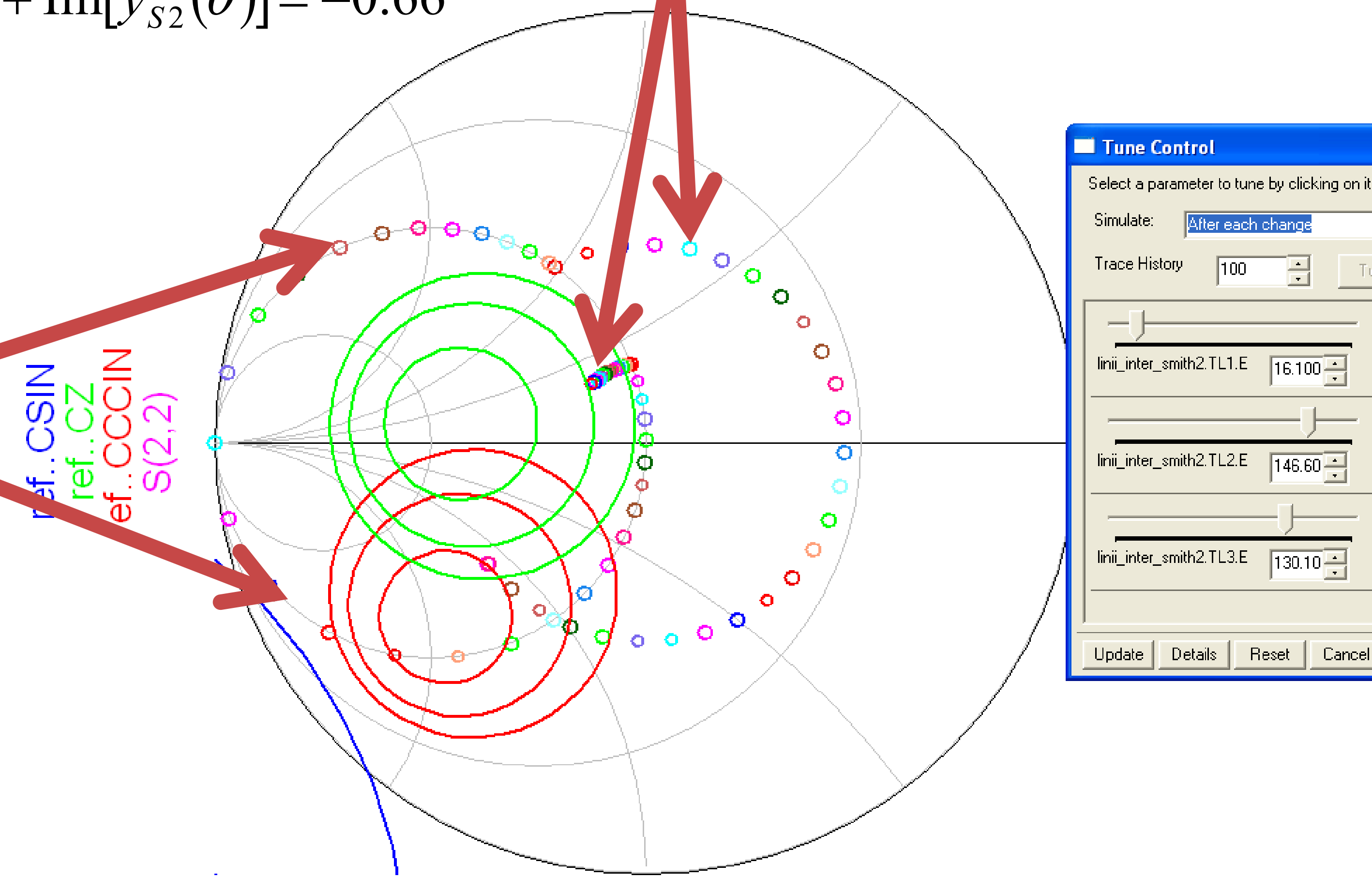
$$\theta_{L1} = 16.1^\circ \qquad \theta_{S2} = 130.1^\circ$$

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

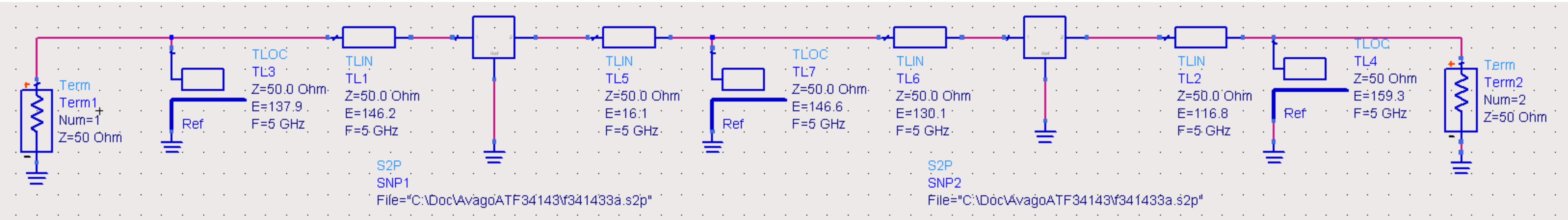
$$\theta_{sp} = 146.6^\circ$$

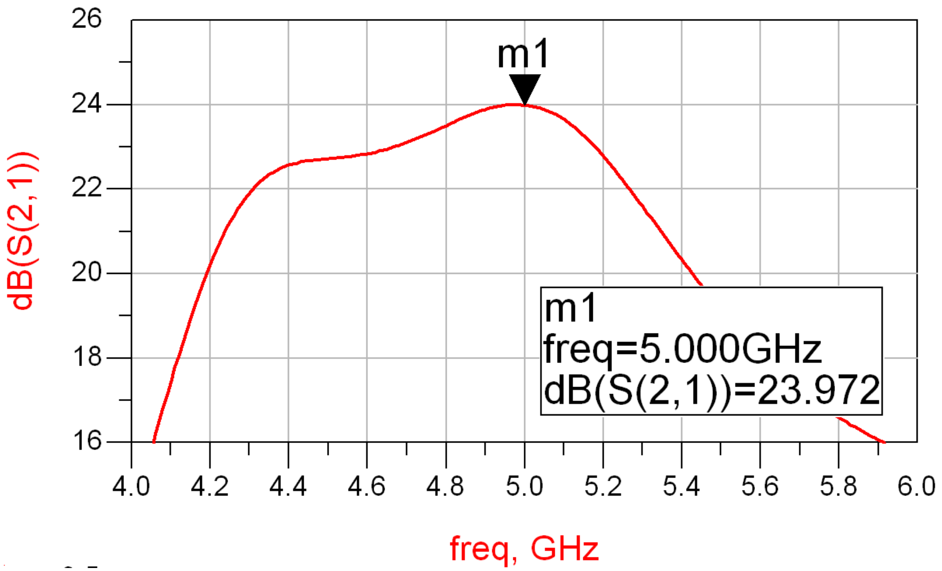
series lines

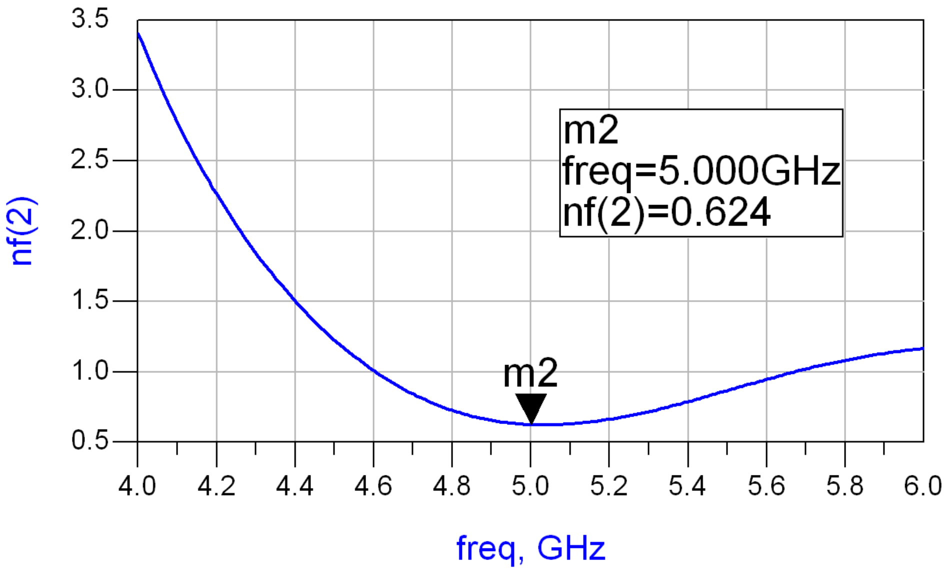
“combined” stub



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


m2
freq=5.000GHz
nf(2)=0.624
m2
nf(2)
freq, GHz

Merge 4, Smith Chart

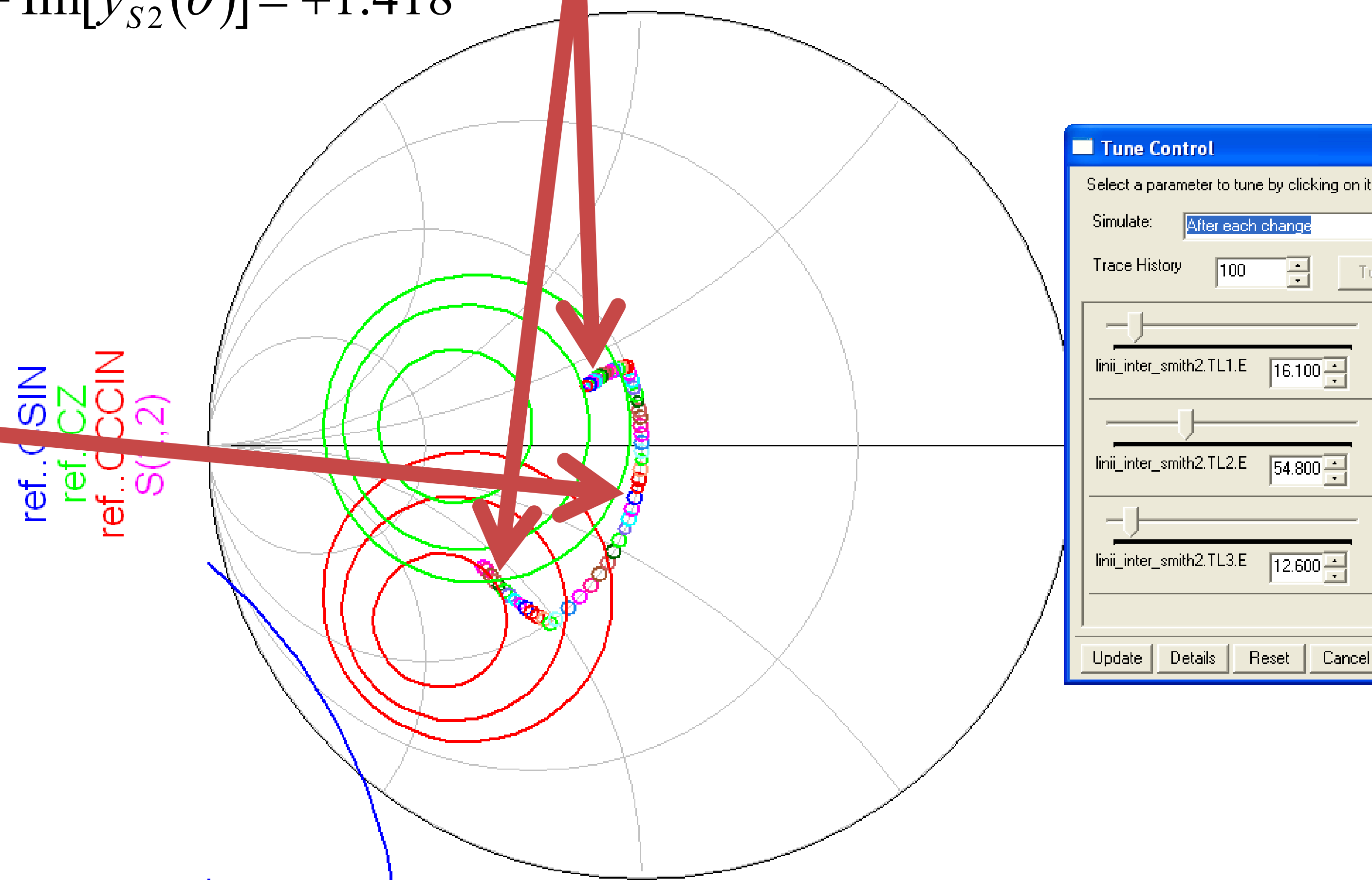
$$\theta_{L1} = 16.1° \qquad \theta_{S2} = 12.6°$$

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

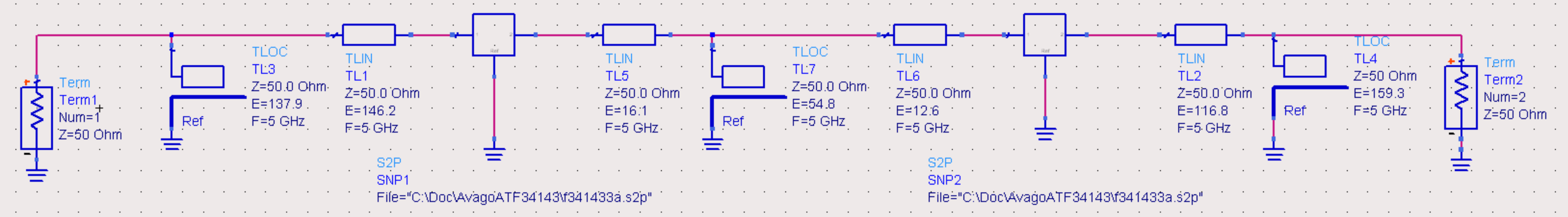
$$\theta_{sp} = 54.8°$$

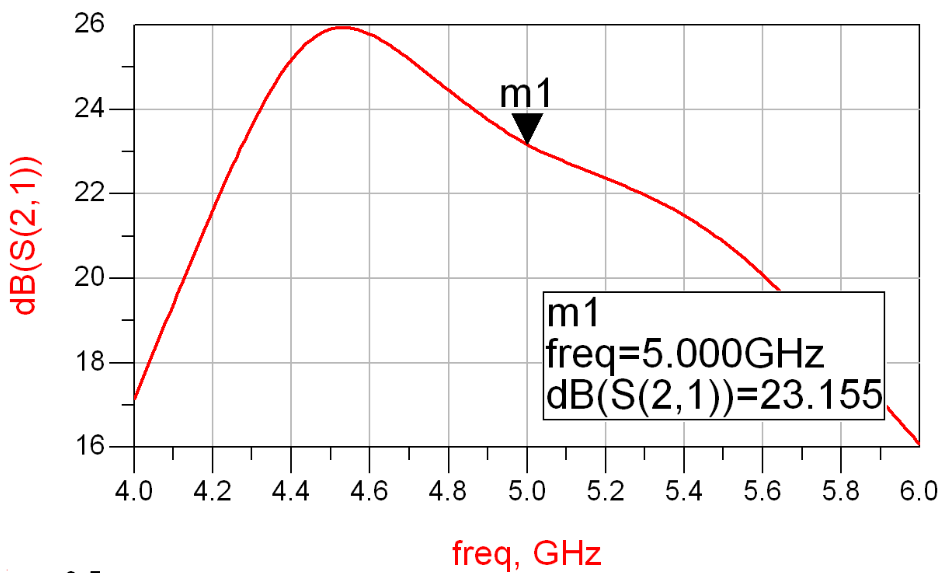
"combined" stub

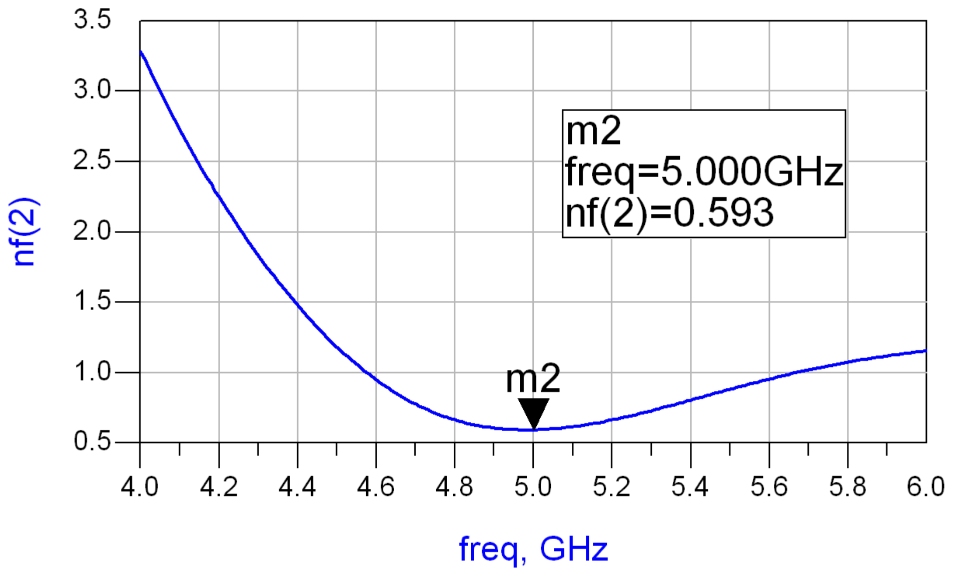
series lines



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
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))=23.155


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
freq=5.000GHz
nf(2)=0.593
m2

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 $l = \frac{\theta}{360°} \cdot \lambda$
  - 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

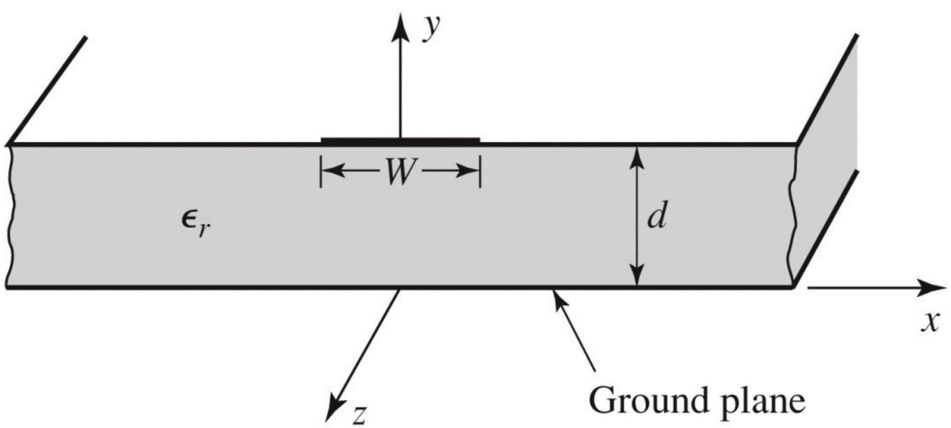


Figure 3.25a

Implementation in microstrip technology

- quasi TEM line

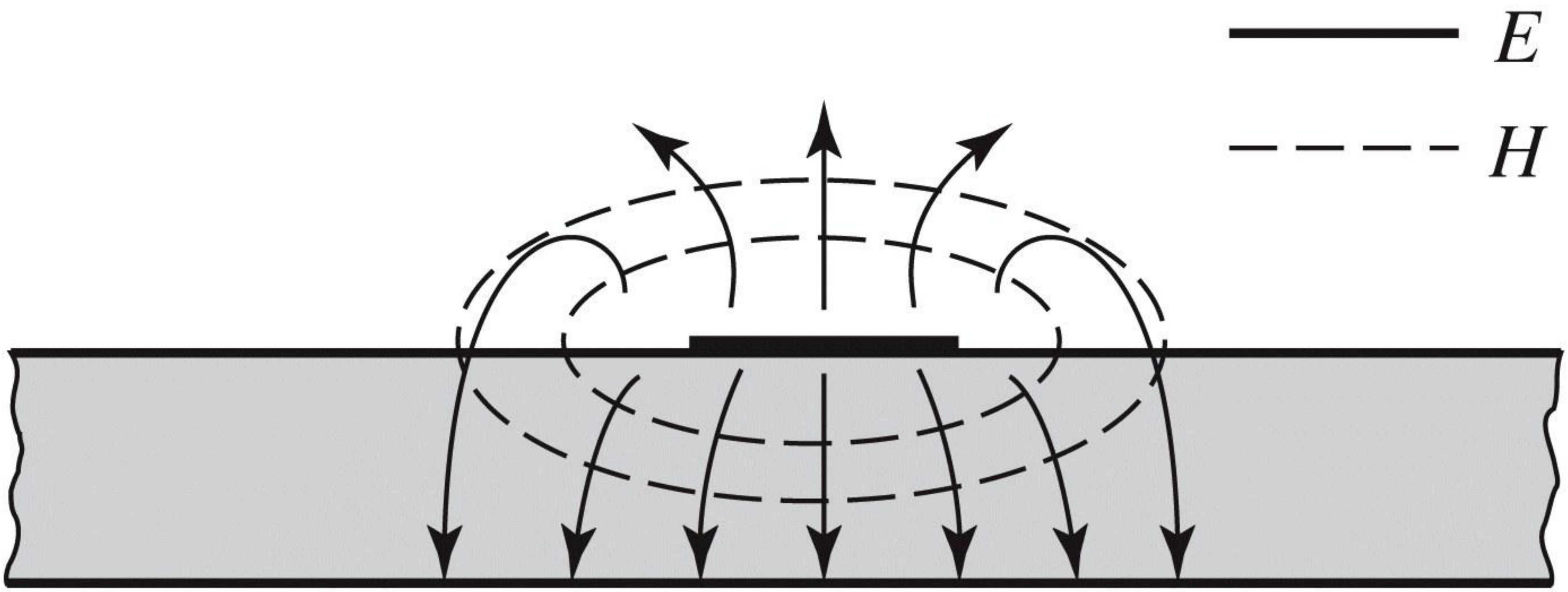
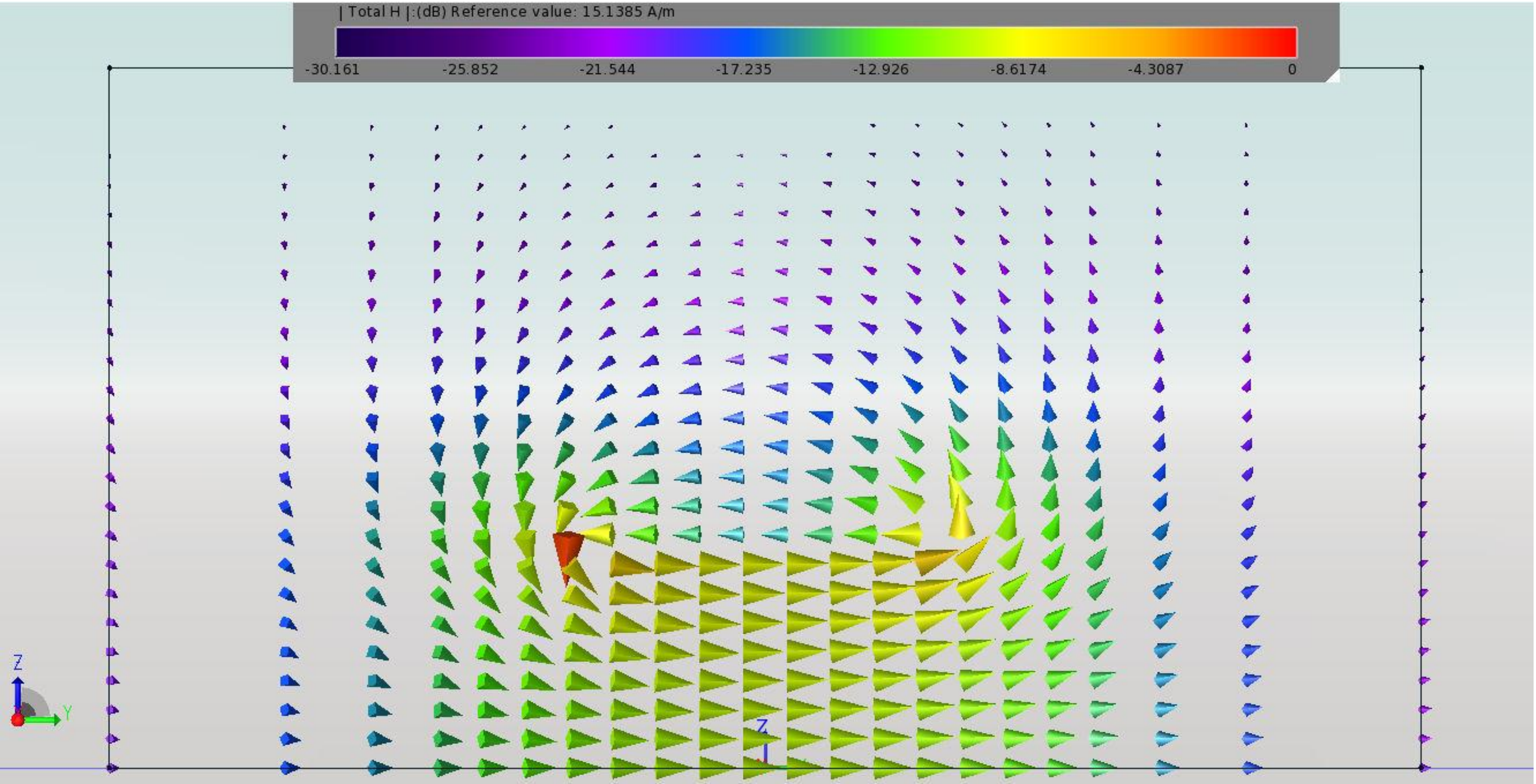


Figure 3.25b

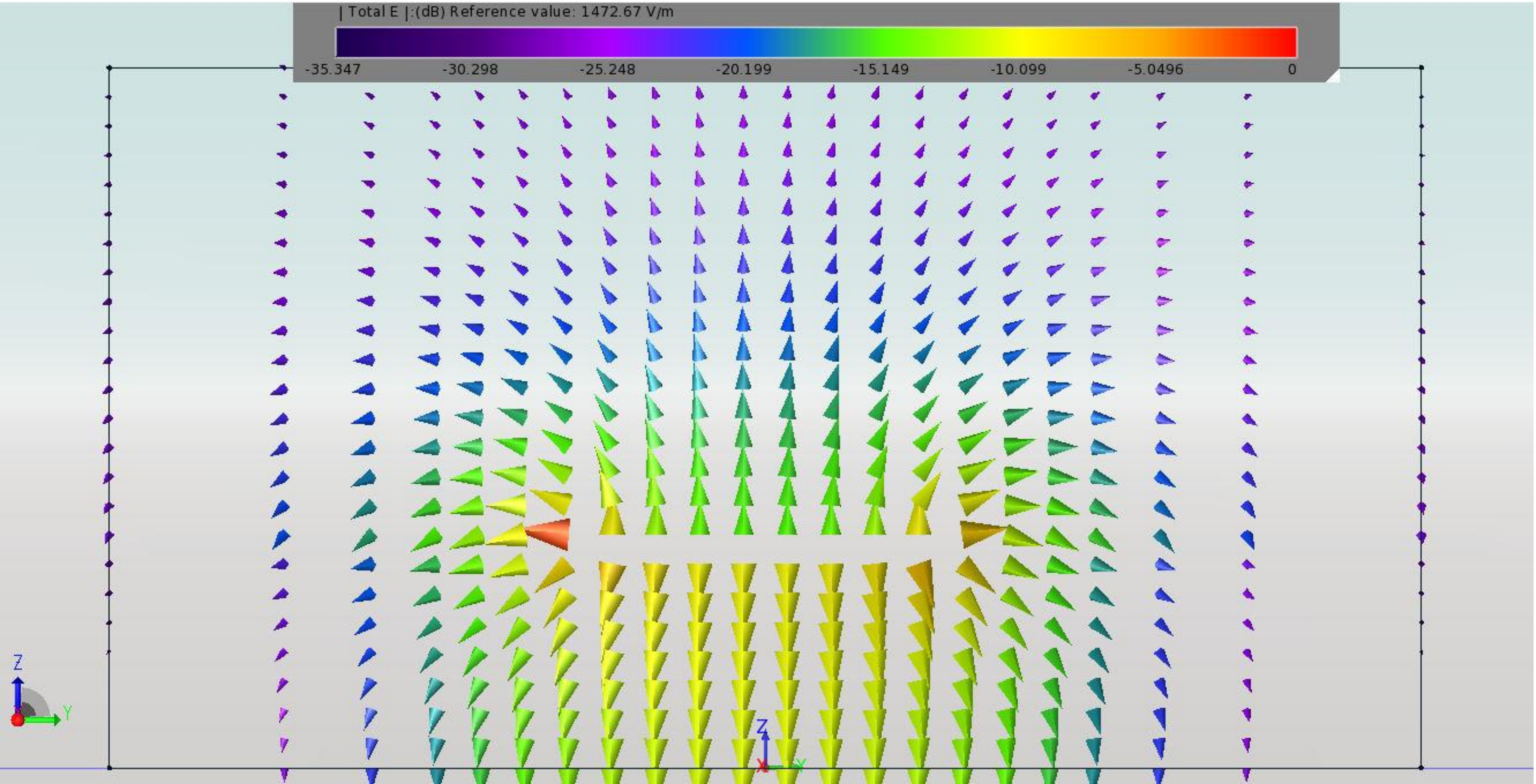
Implementation in microstrip technology

- quasi TEM line, EmPro



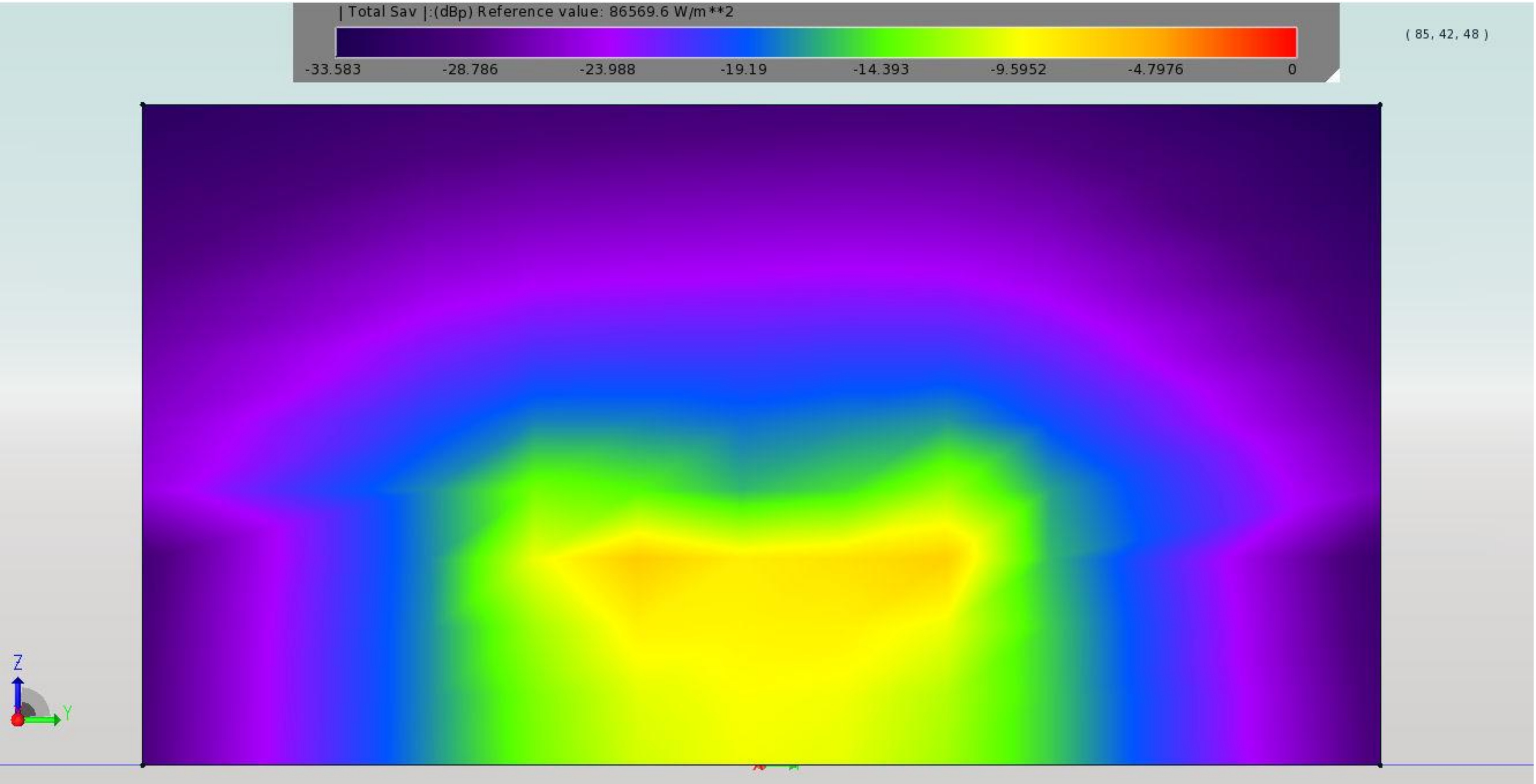
Implementation in microstrip technology

- quasi TEM line, EmPro



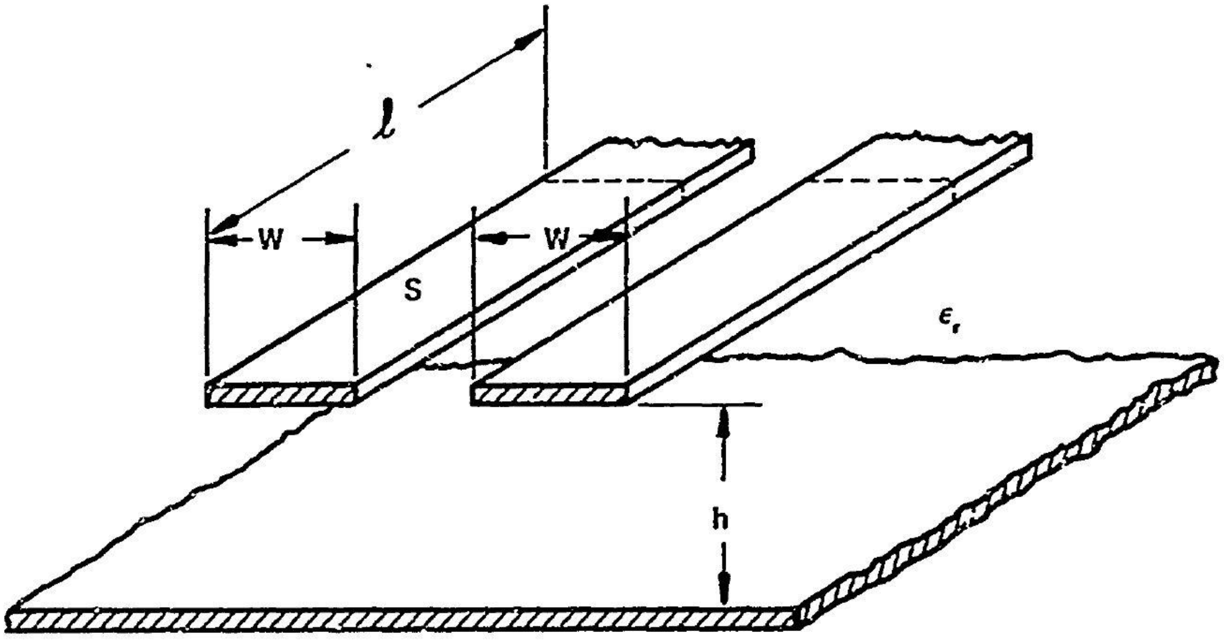
Implementation in microstrip technology

- quasi TEM line, EmPro



Implementation in microstrip technology

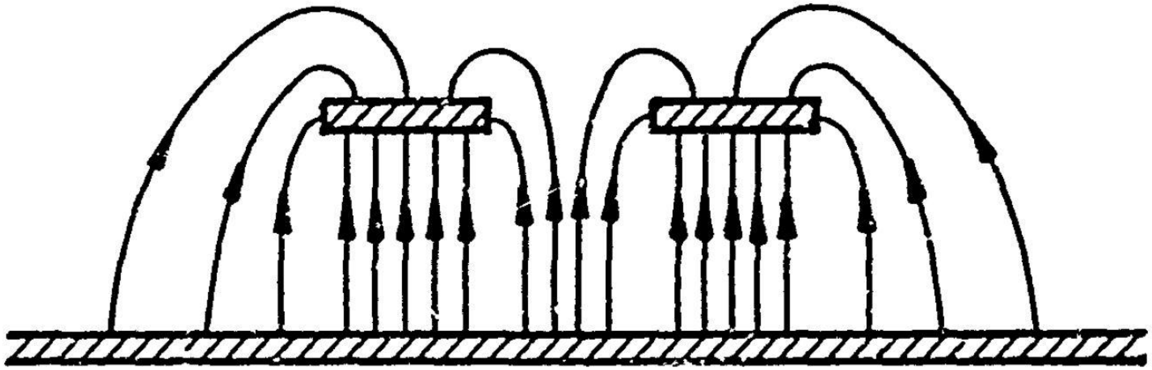
- ~ quasi TEM



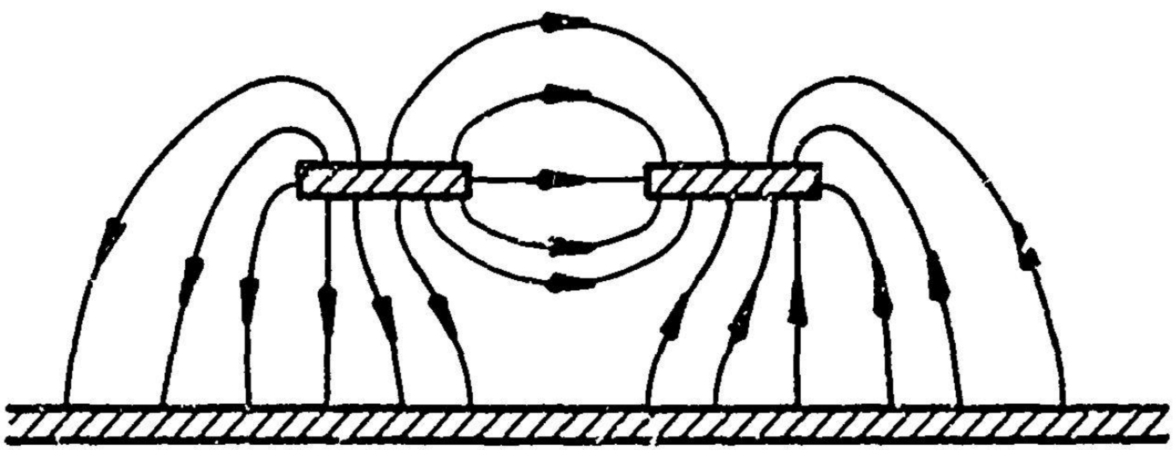
a) COUPLED STRIP GEOMETRY

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

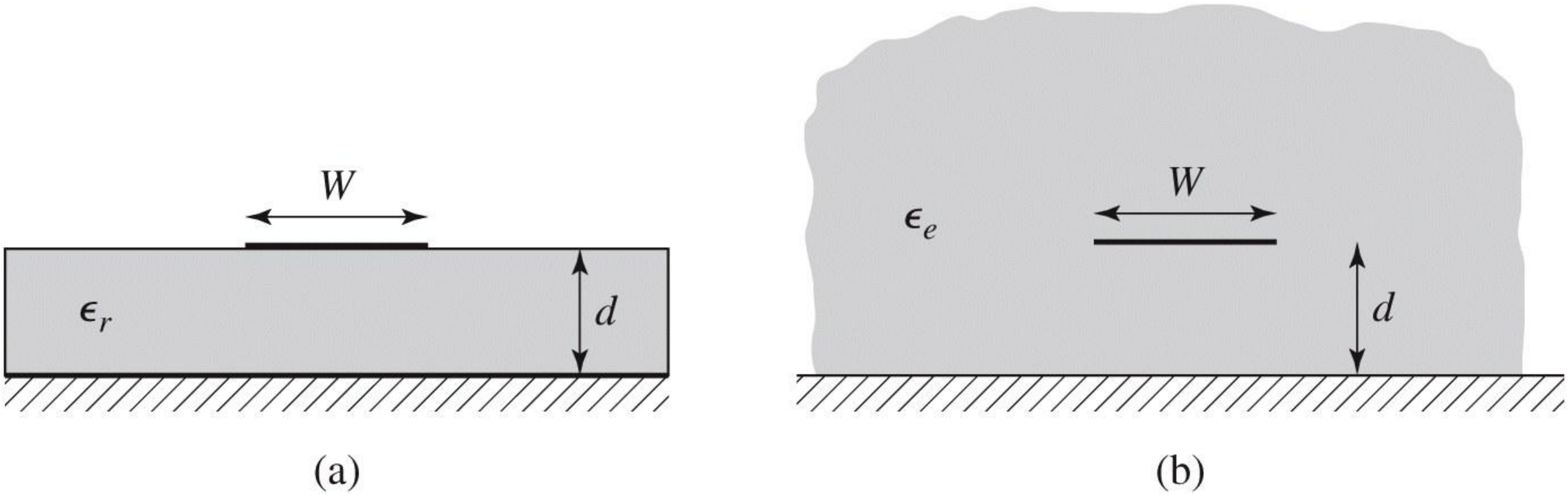


Figure 3.26

Design

- Empirical formulas

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

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

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

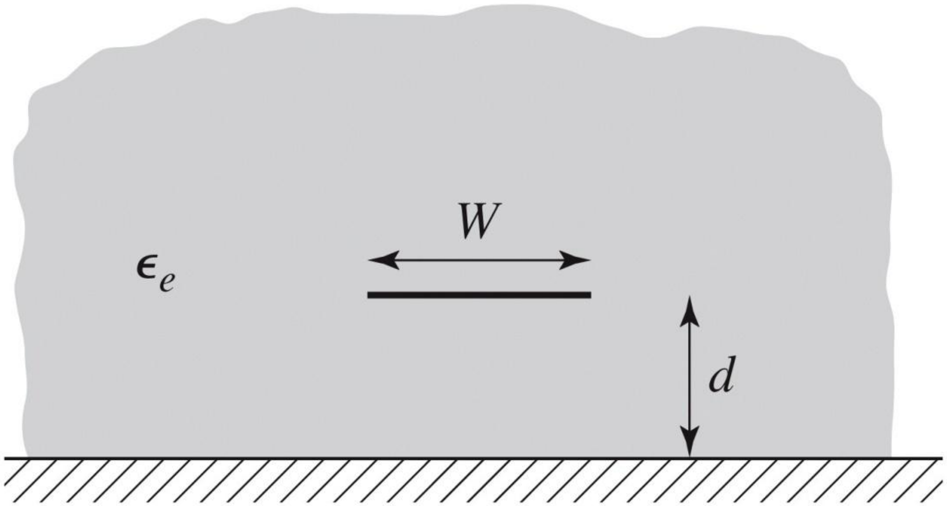


Figure 3.26b

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

Design

- Empirical formulas

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

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

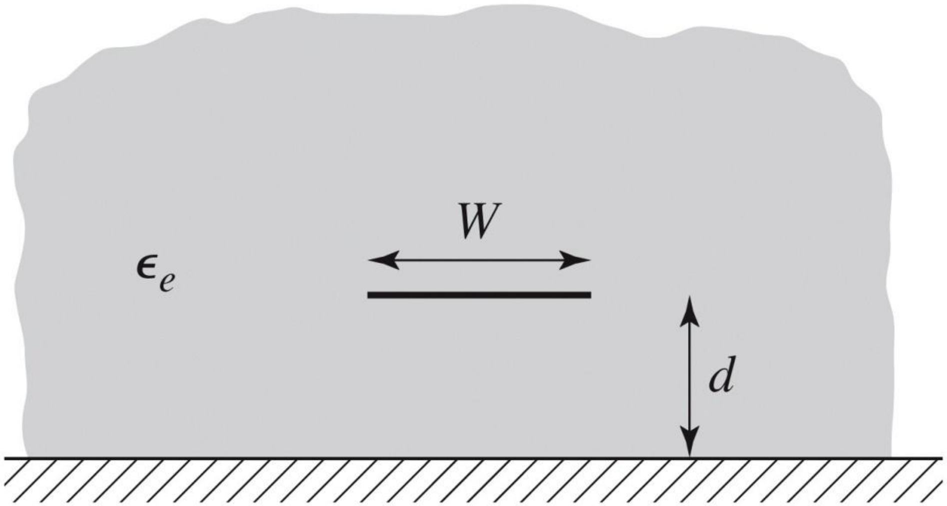
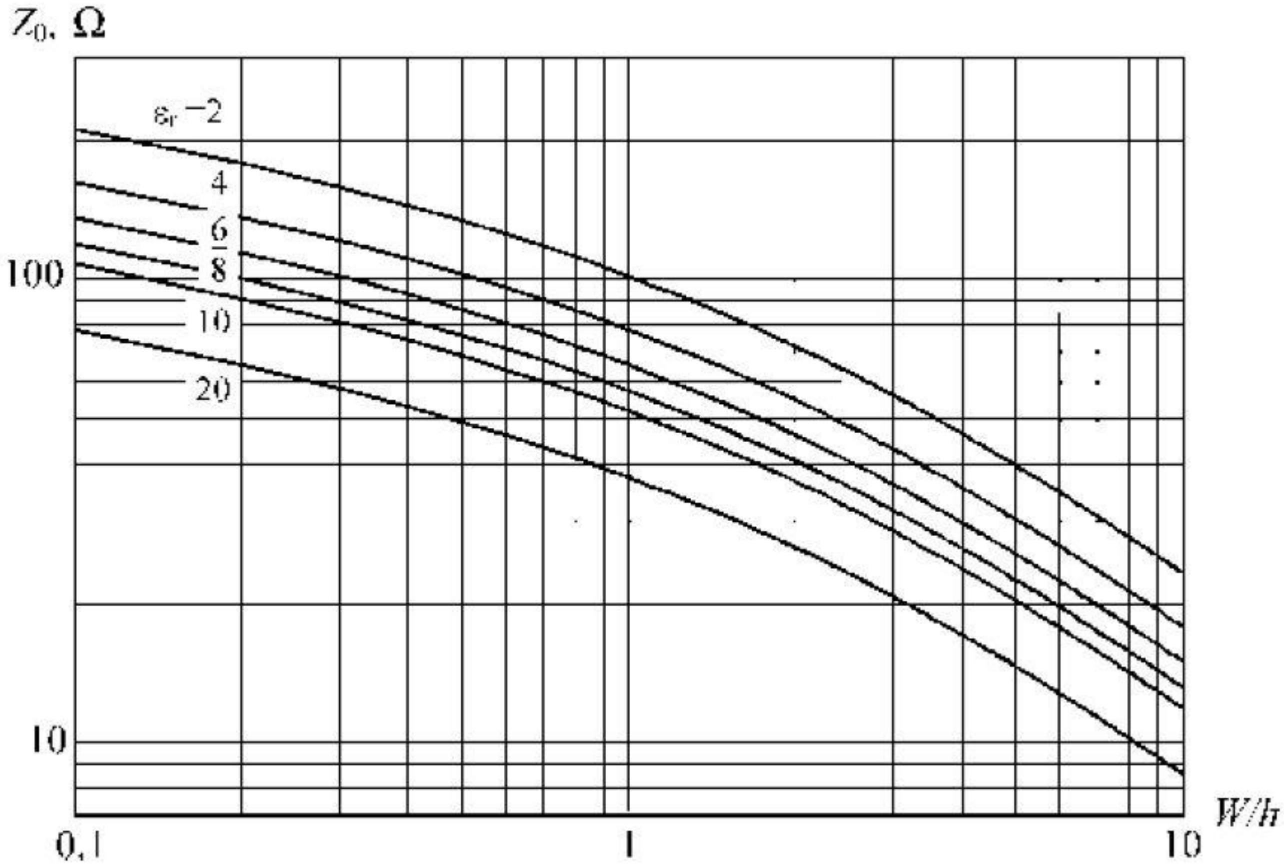


Figure 3.26b

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

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

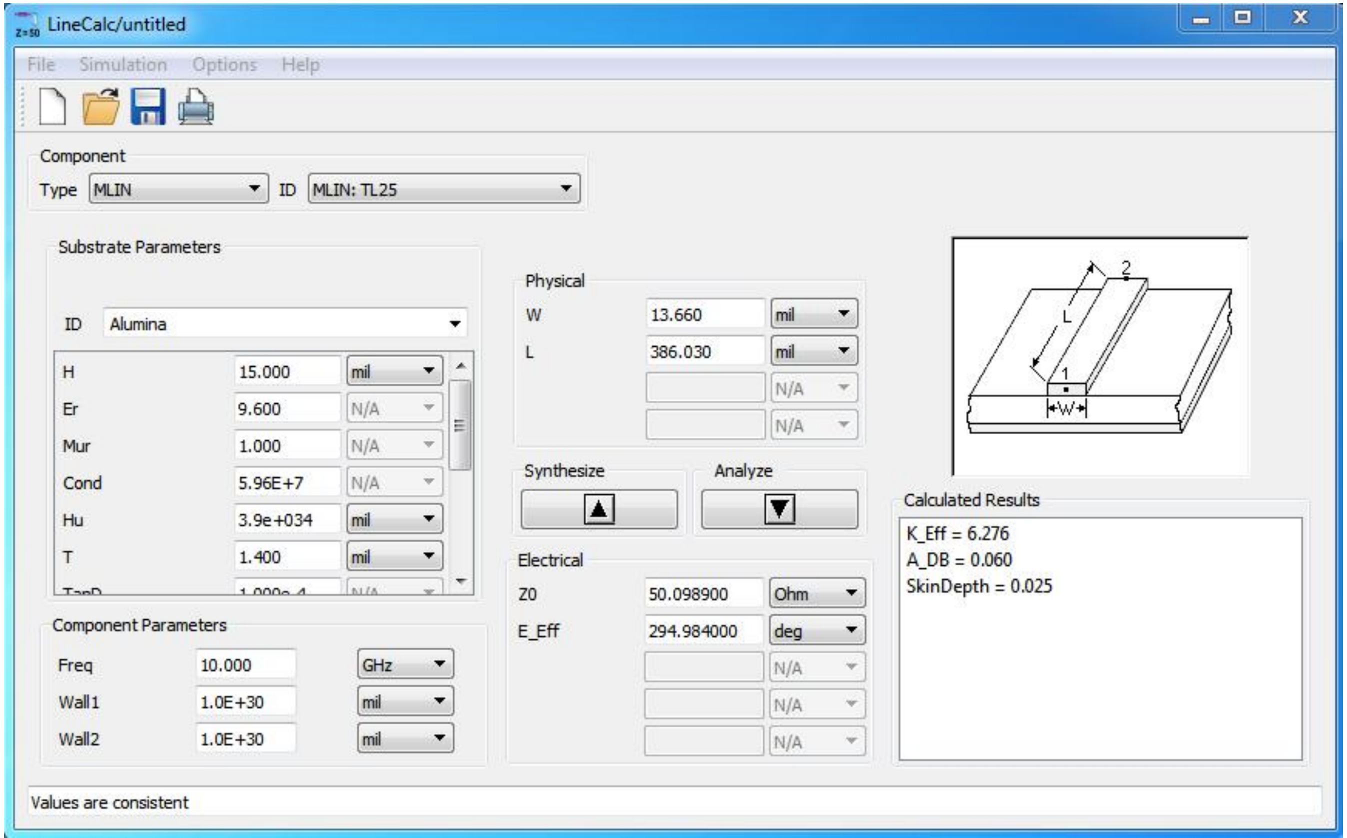
Microstrip standardization

- 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)

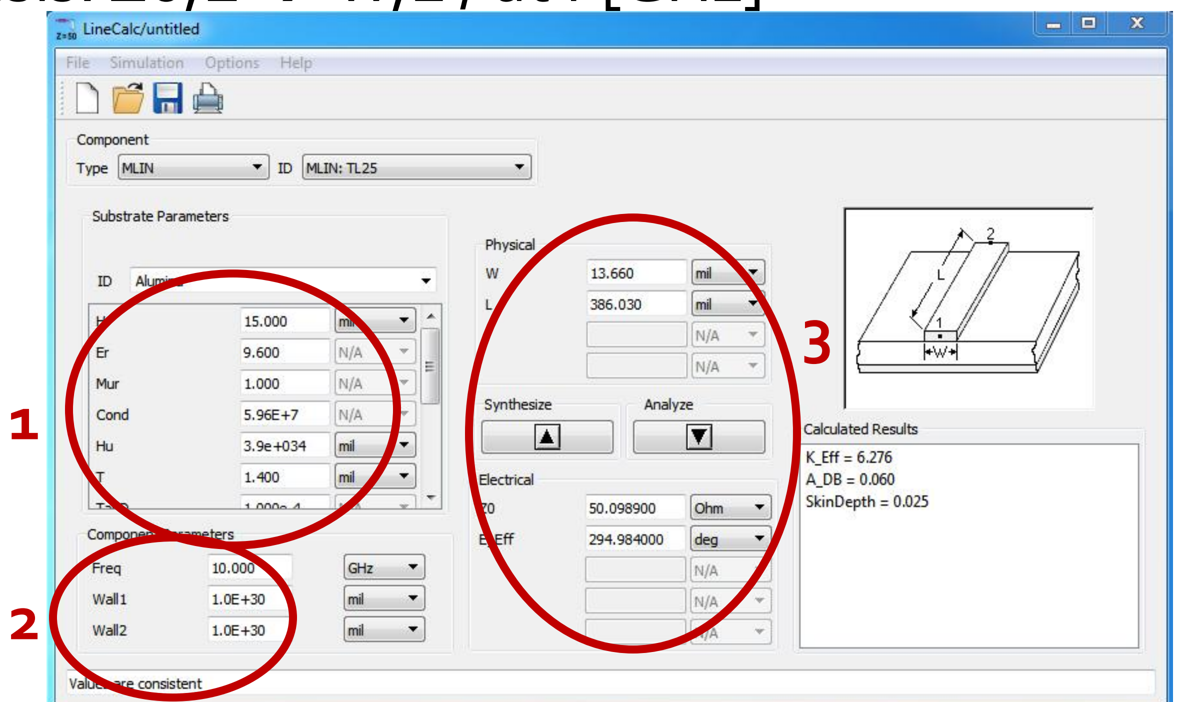
ADS linecalc

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



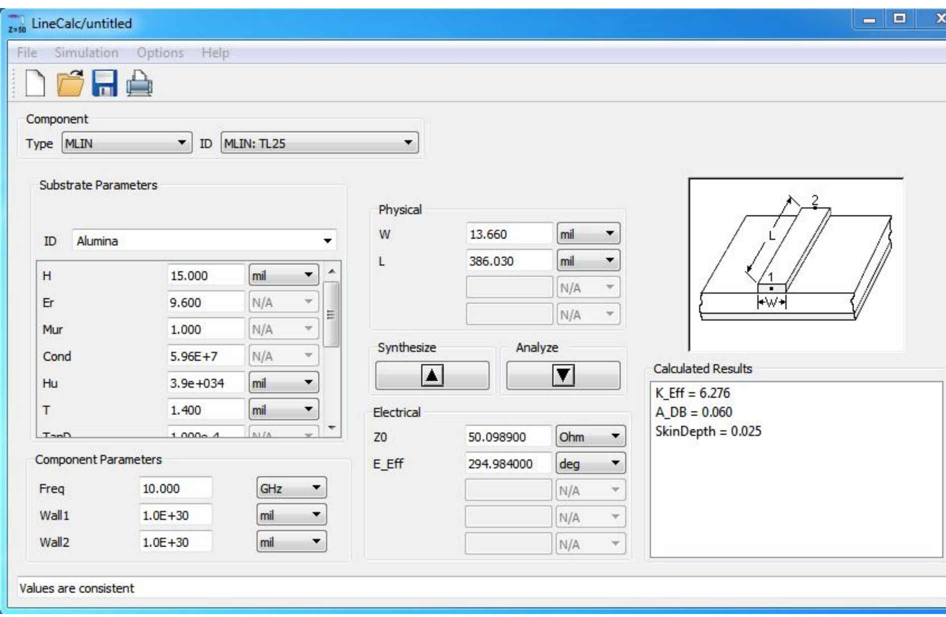
ADS linecalc

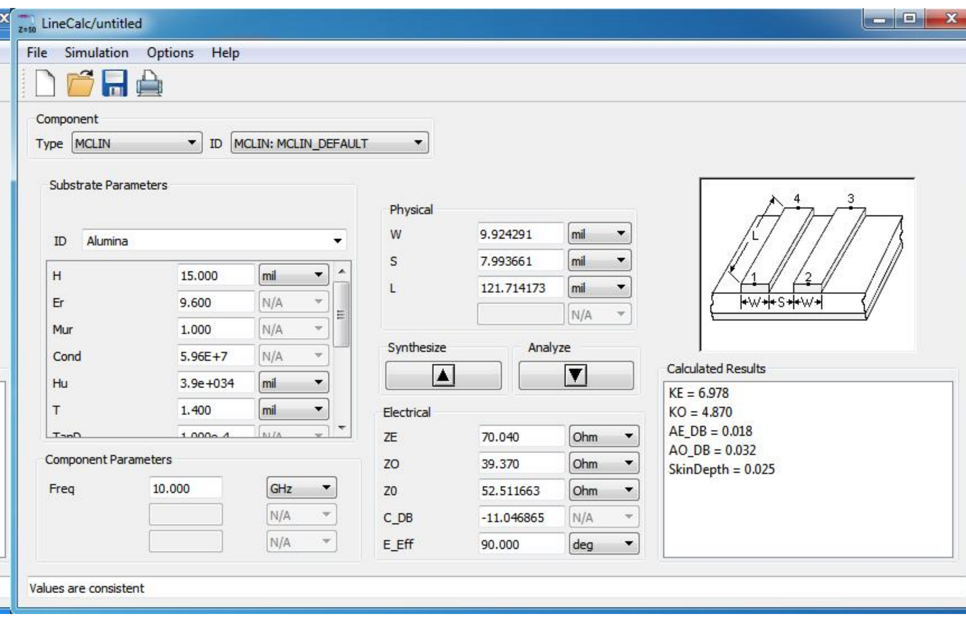
- 1. Define substrate (receive from schematic)
- 2. Insert frequency
- 3. Insert input data
  - Analyze: W,L → Zo,E or Ze,Zo,E / at f [GHz]
  - Synthesis: Zo,E → W,L / at f [GHz]



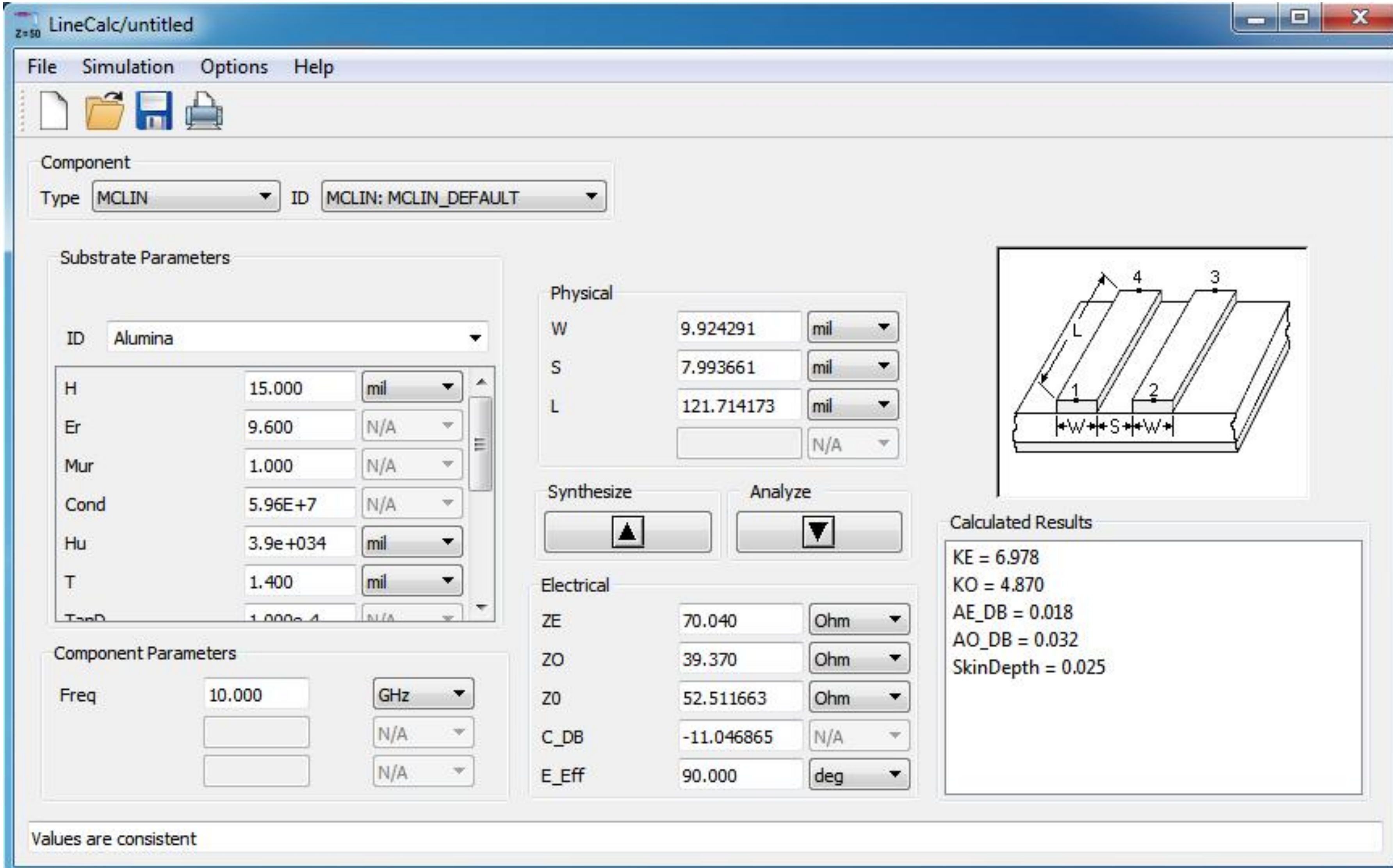
ADS linecalc

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





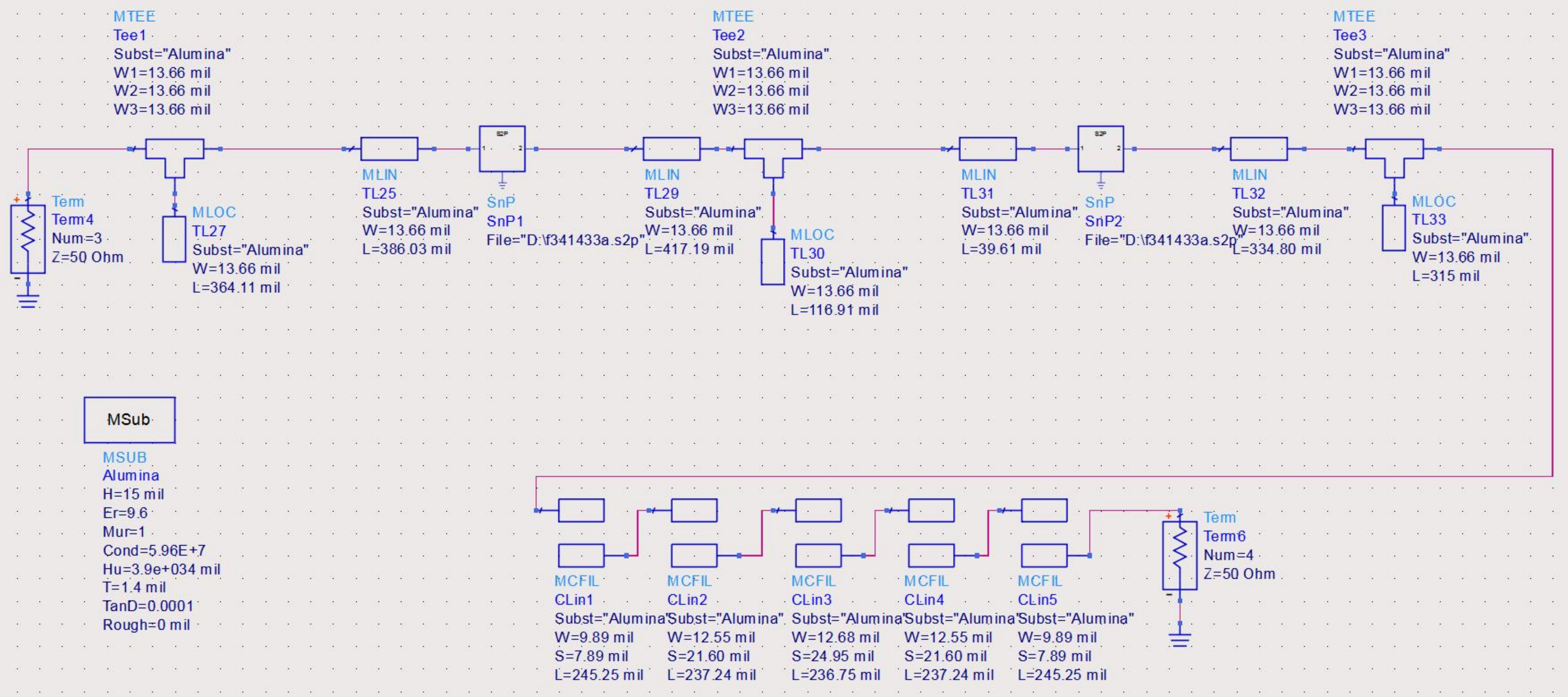
ADS linecalc



Transmission lines

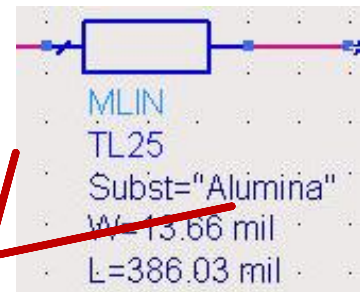
- http://rf-opto.etti.tuiasi.ro
- Transmission lines / Rogers
  - more precise formulas including
    - t, trace thickness
    - f, frequency
  - formulas for
    - microstrip
    - strip
    - coupled lines

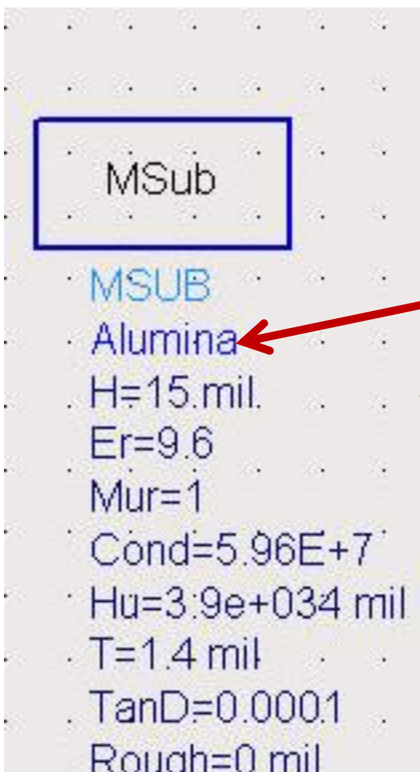
Implementation in microstrip technology

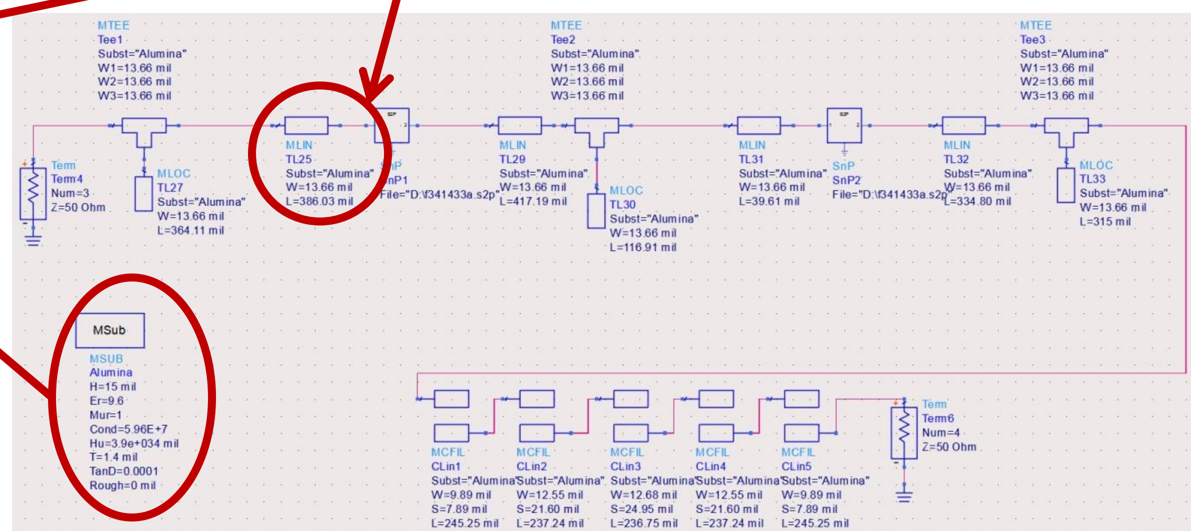


Implementation in microstrip technology

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





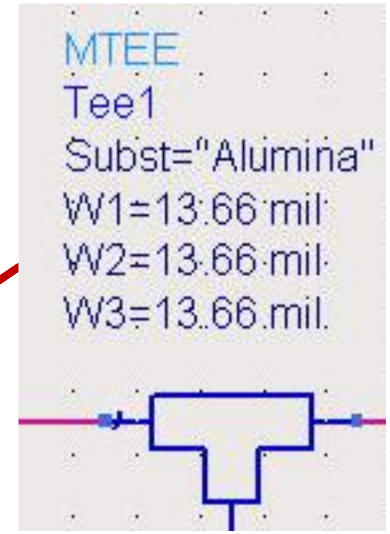


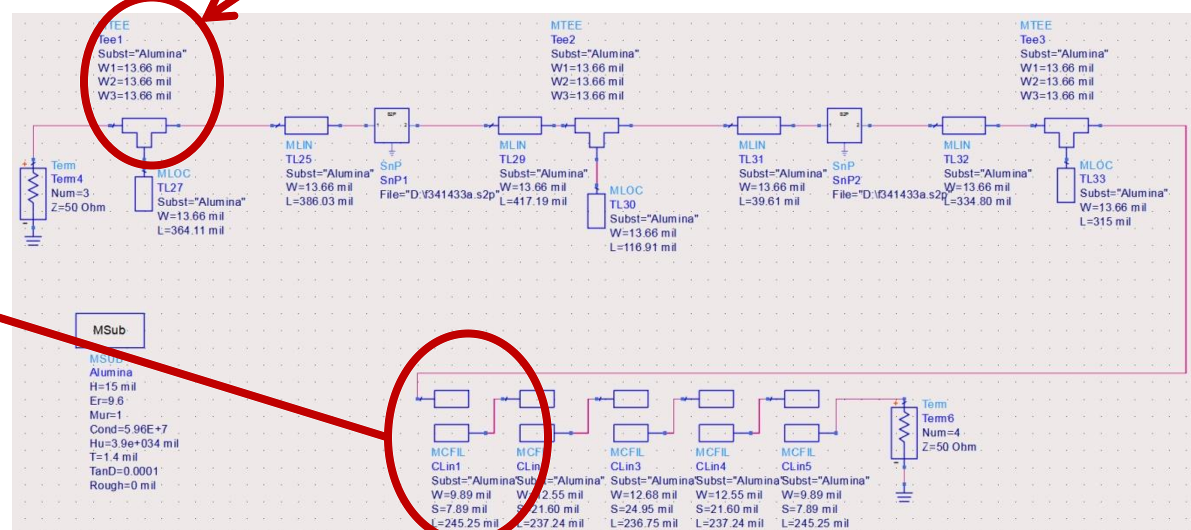
Implementation in microstrip technology

- 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

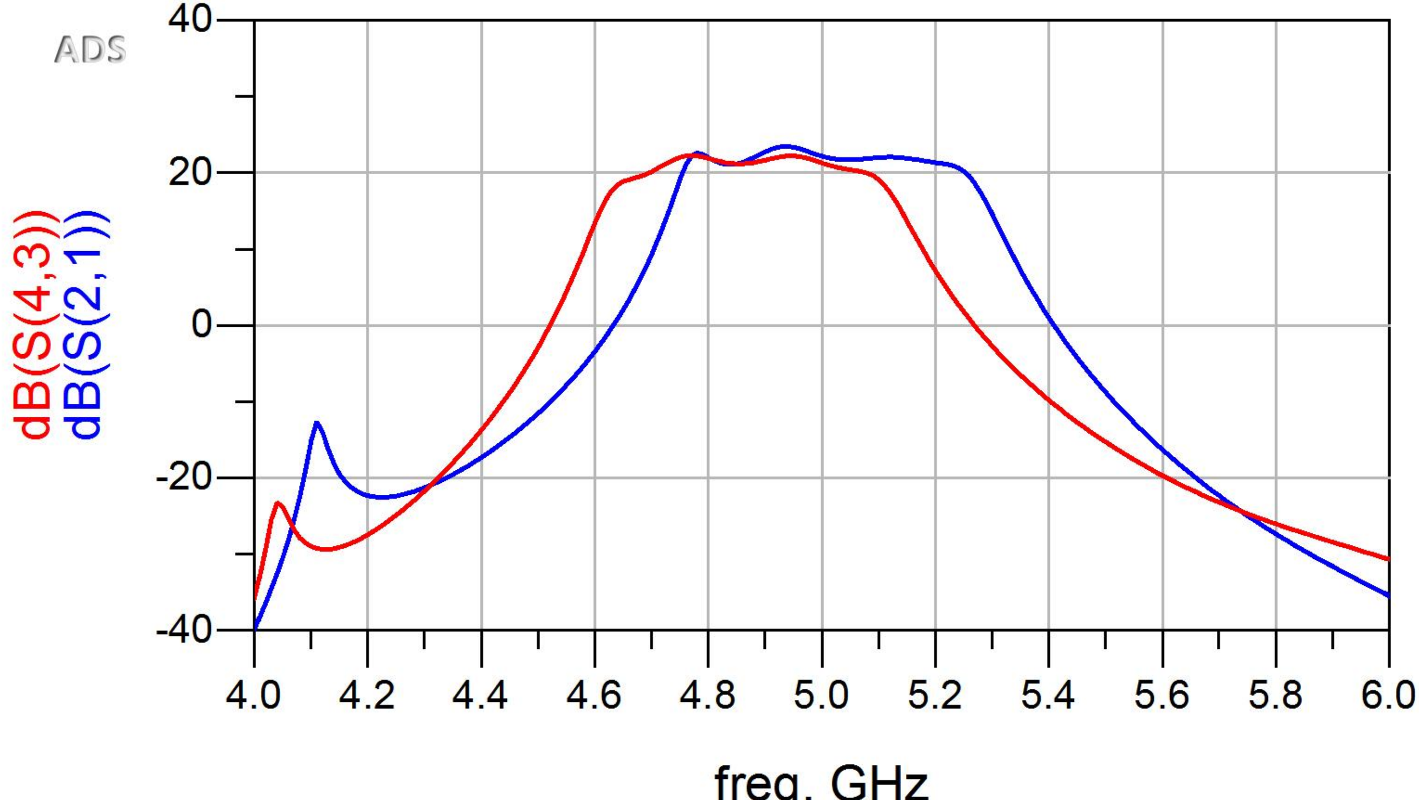
- Attention is required when inserting parameters for MTEE and MCFIL by checking in the schematic the width of the lines connected to each port.





Implementation in microstrip technology

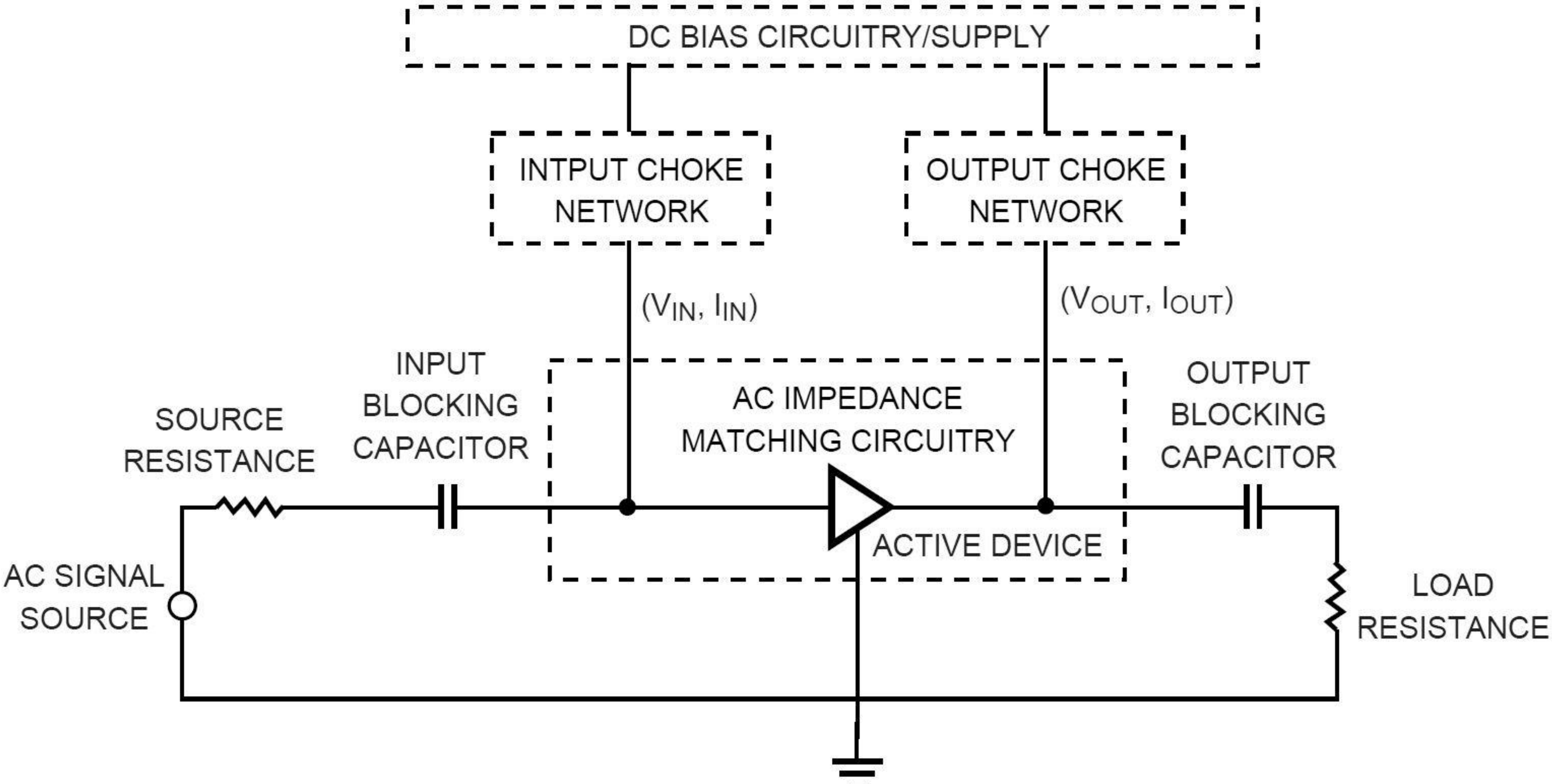
- Usually there is a shift of the transfer function (red) towards lower frequencies compared to the ideal model (blue)
  - due to the MCFIL/MCLIN difference
- Tune the length of filter elements to move the filter bandwidth around fo = 5GHz



DC Bias

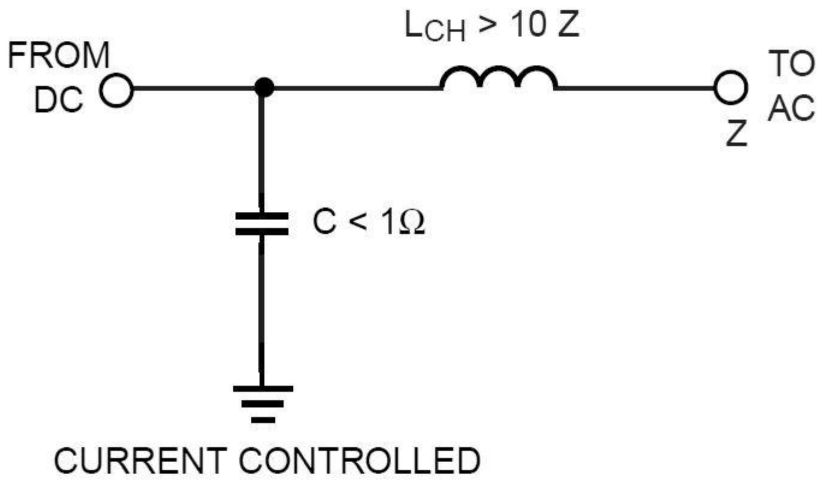
- https://rf-opto.etti.tuiasi.ro
- Agilent Application Notes
  - decoupling signal from DC Bias circuitry
  - DC Bias circuits for microwave transistors
- Appcad has tools for designing DC Bias circuits

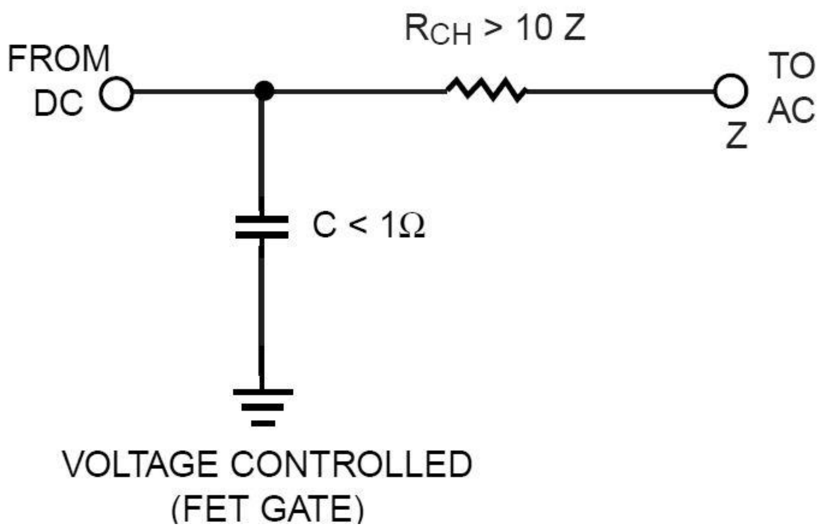
DC Bias



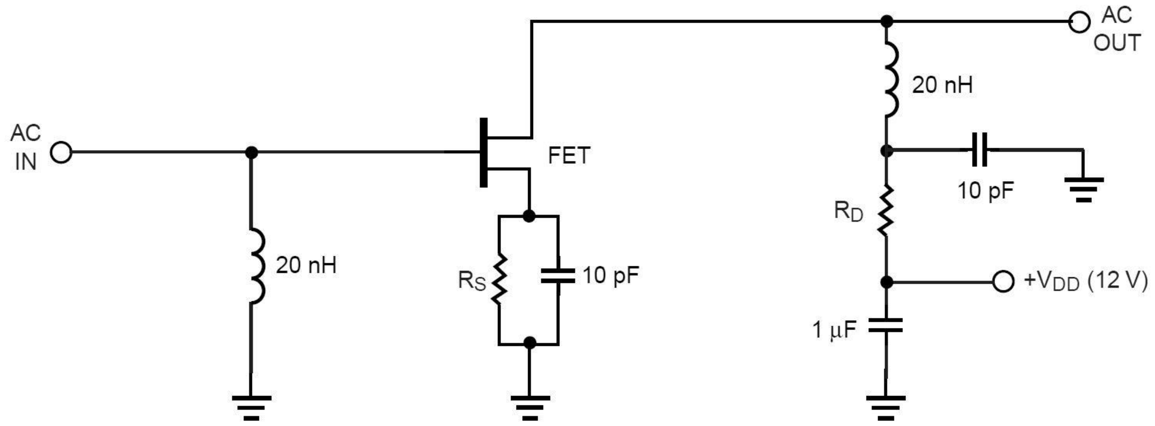
DC BIAS CIRCUITRY/SUPPLY
INTPUT CHOKE NETWORK
OUTPUT CHOKE NETWORK
(VIN, IIN)
(VOUT, IOUT)
INPUT BLOCKING CAPACITOR
SOURCE RESISTANCE
AC IMPEDANCE MATCHING CIRCUITRY
OUTPUT BLOCKING CAPACITOR
ACTIVE DEVICE
AC SIGNAL SOURCE
LOAD RESISTANCE

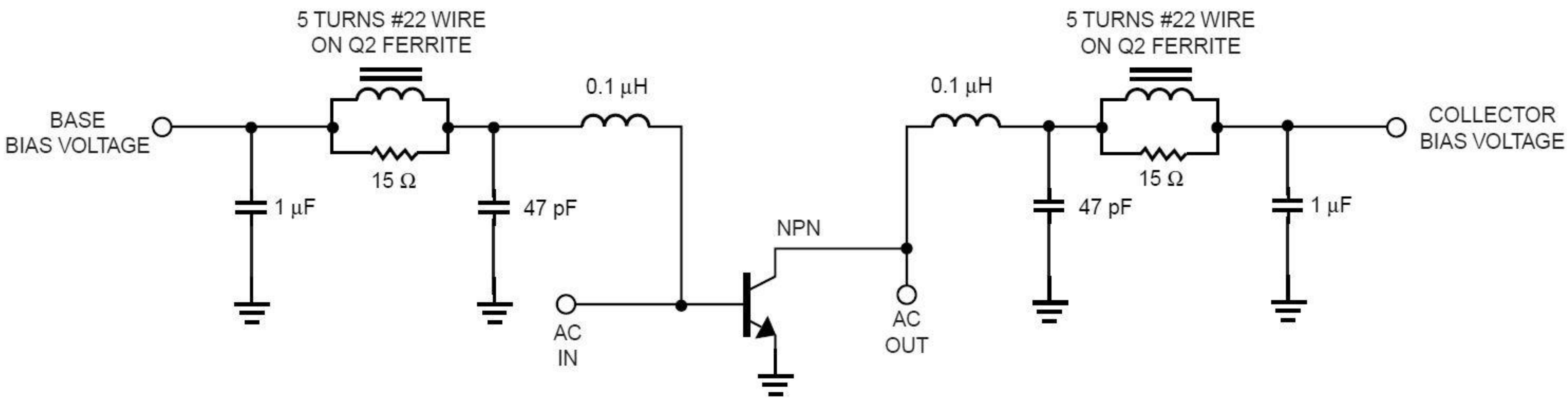
DC Bias, typical choke



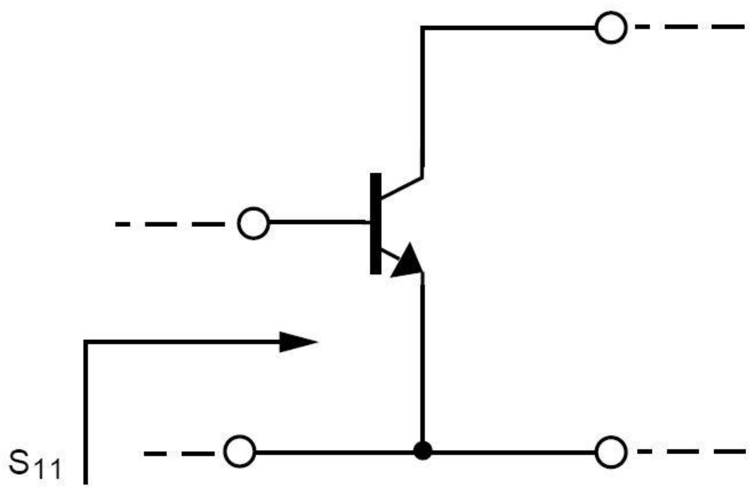


DC Bias, typical schematics/values



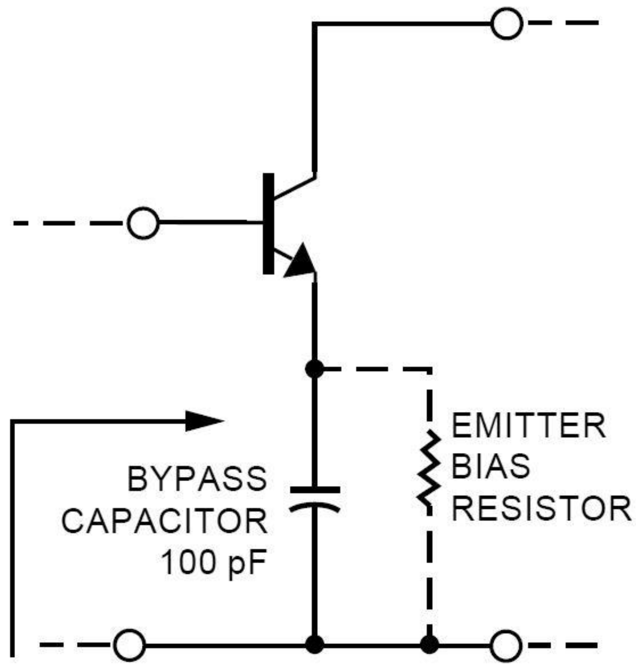


DC Bias, elements in E/S



S_{11} (AT 4 GHz) = 0.52 $\angle$ 154°

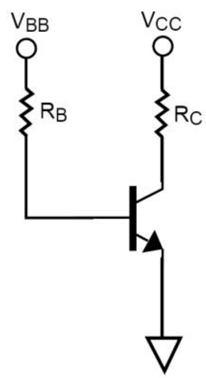
S_{11} (AT 0.1 GHz) = 0.901 $\angle$ −14.9°



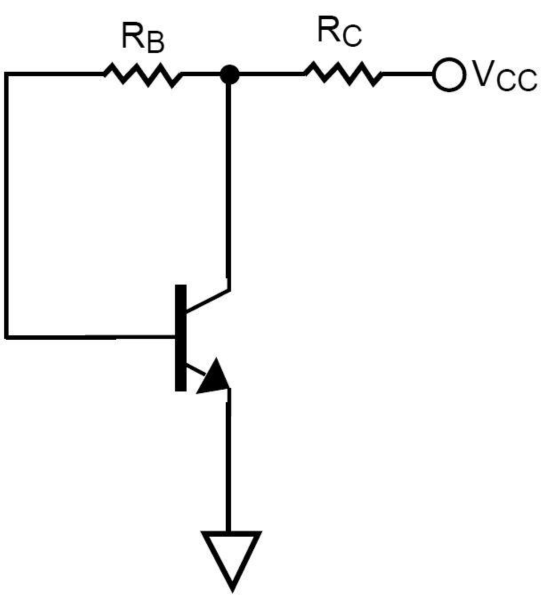
S'_{11} (AT 4 GHz) = 0.52 $\angle$ 154° UNCHANGED AT 4 GHz

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

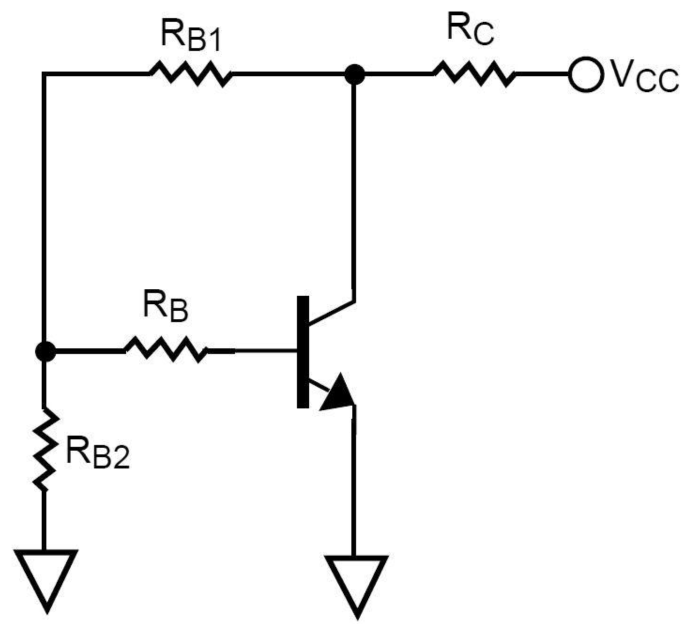
DC Bias, bipolar transistors



NON-STABILIZED



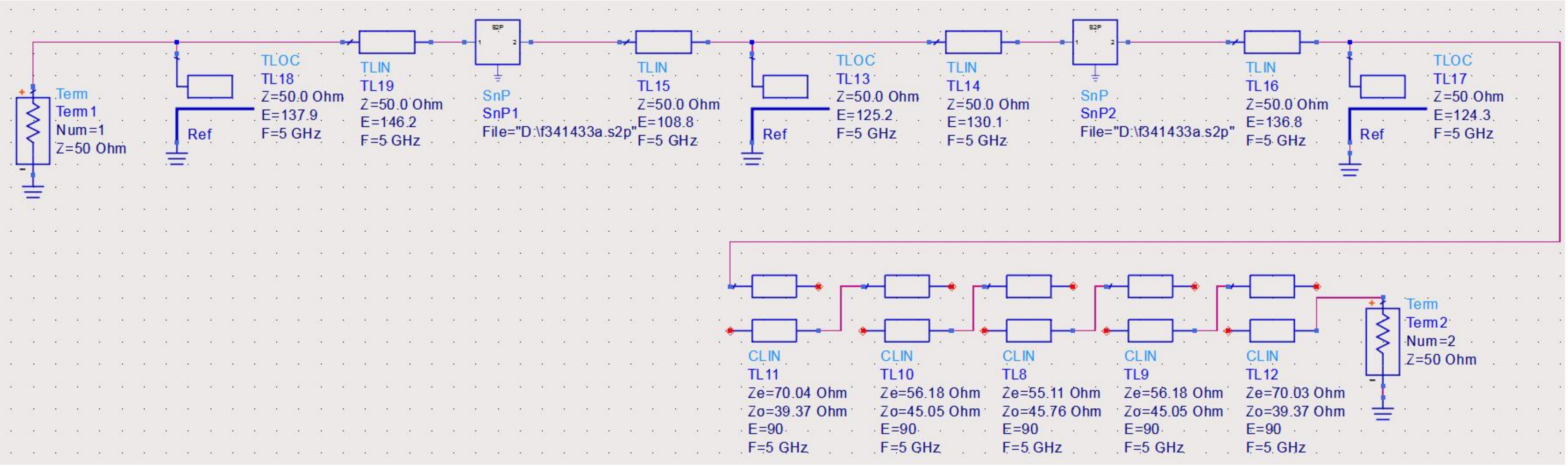
VOLTAGE FEEDBACK



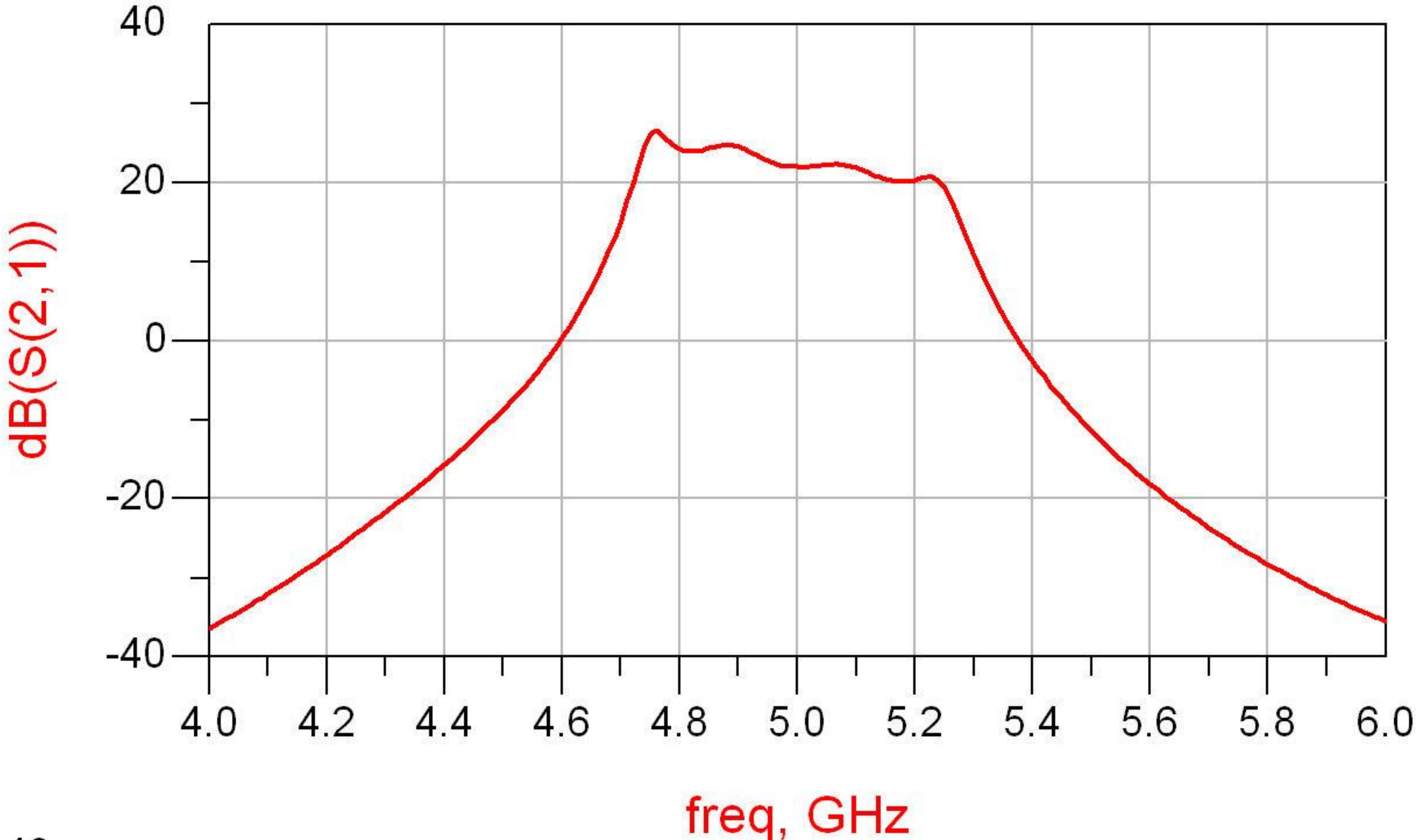
VOLTAGE FEEDBACK AND CONSTANT BASE CURRENT SOURCE

Example project

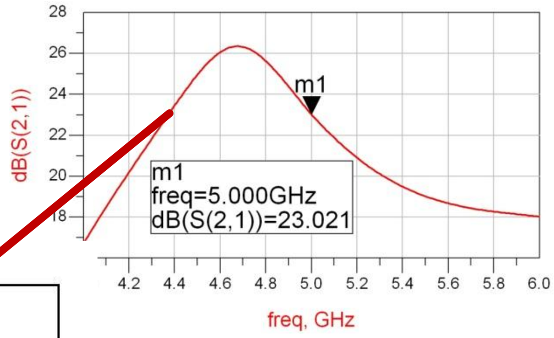
- Unify the two schematics
  - L10 – amplifier
  - L12 – filter

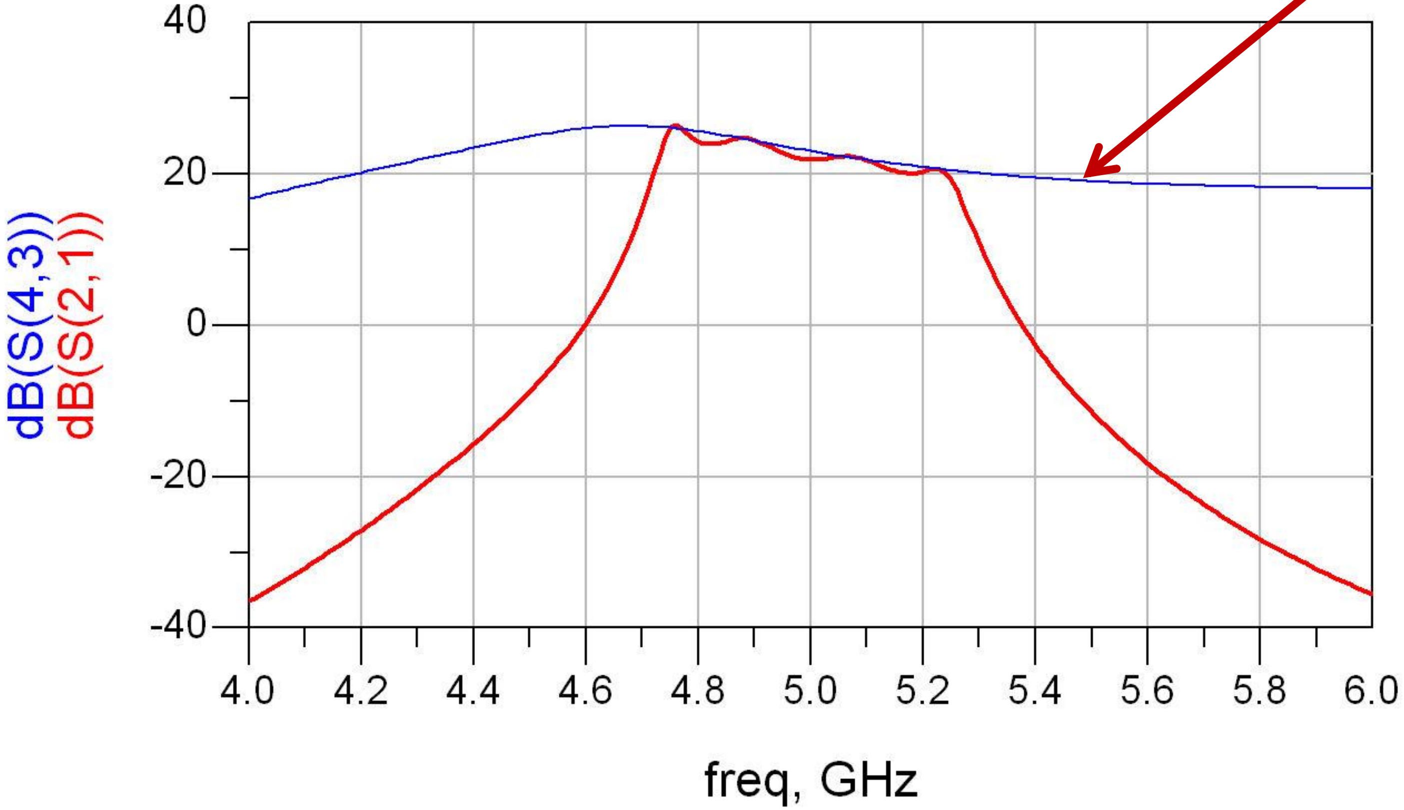


Result (unbalanced)

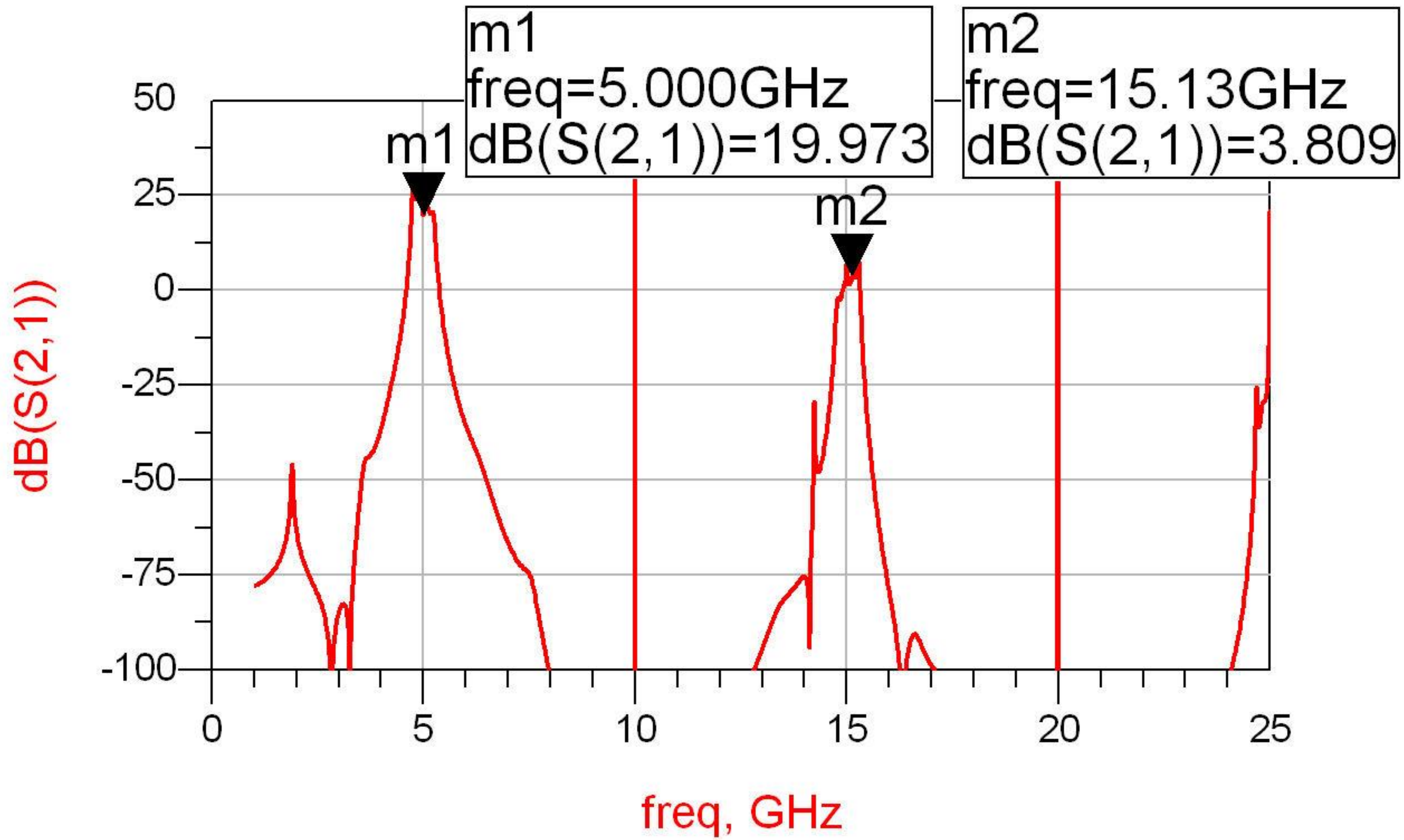


Result (unbalanced)



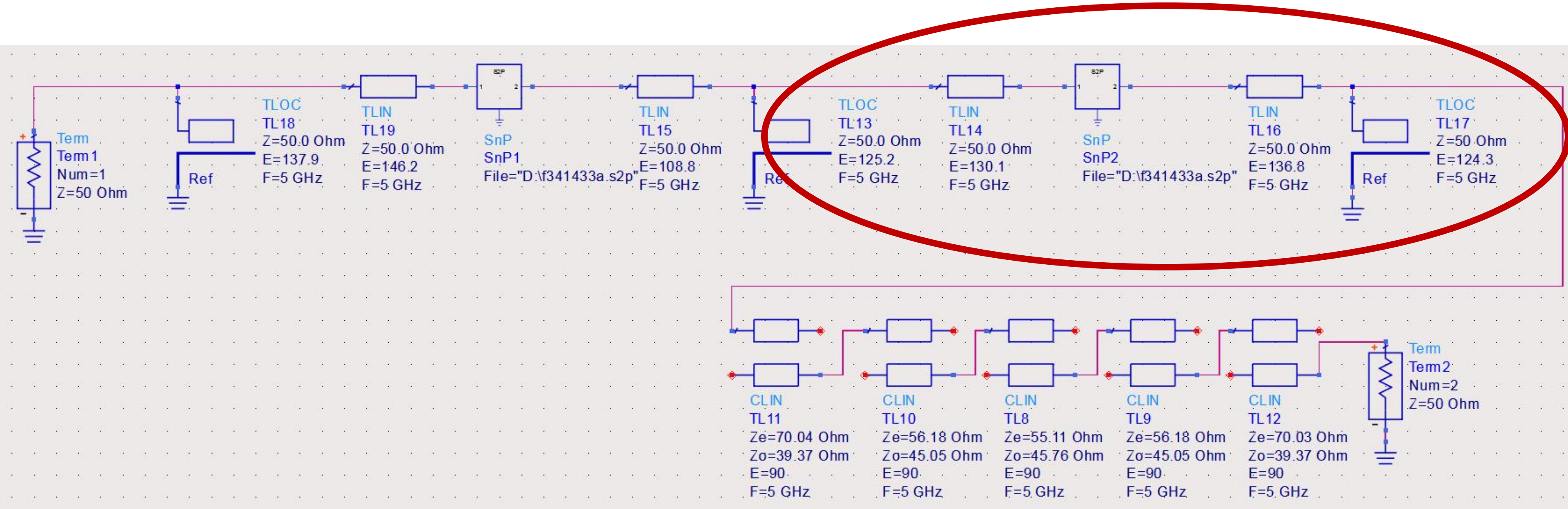


Result (~periodic in frequency)

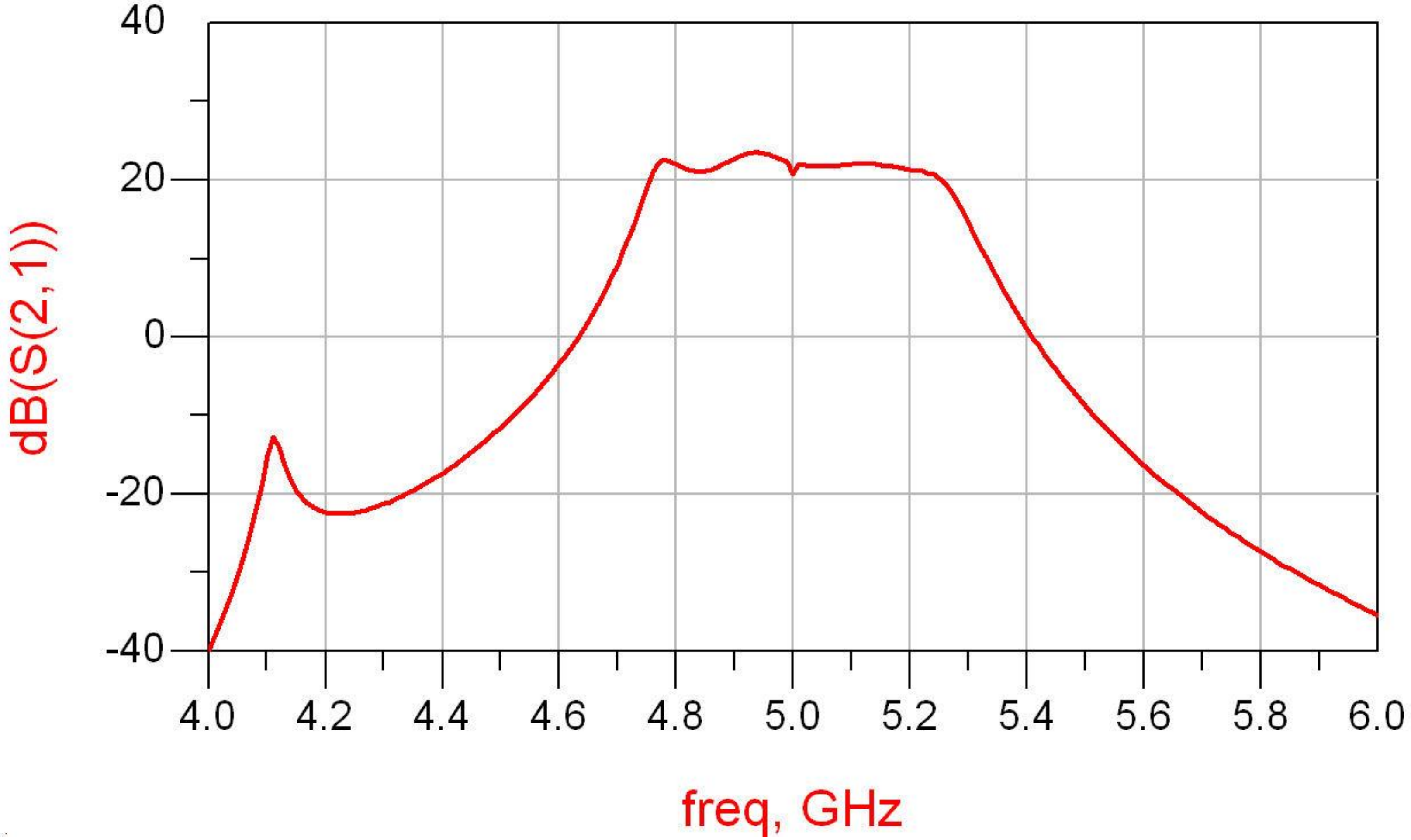


Tune -> balance

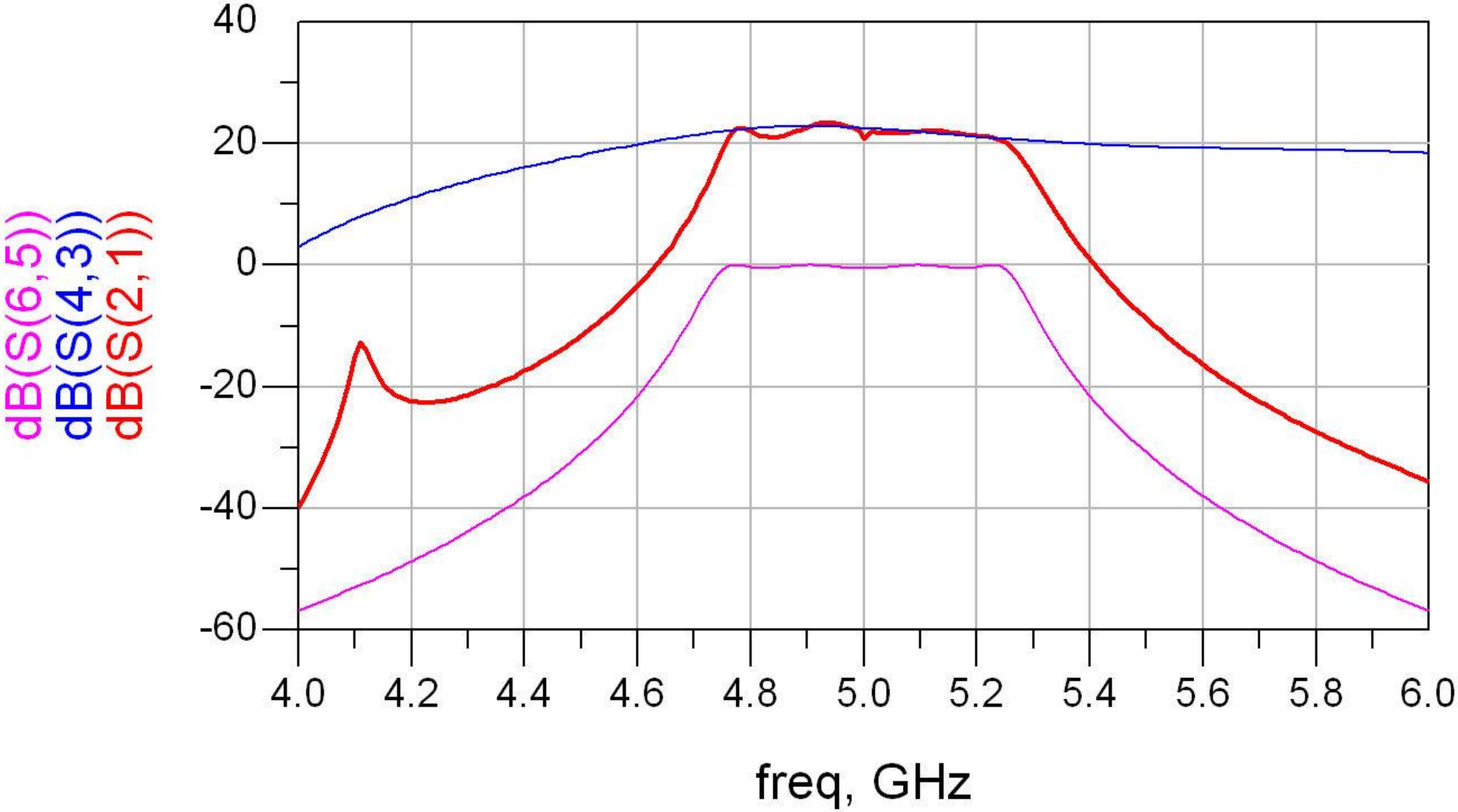
- purpose: balance the gain characteristic of the amplifier (maximum at design frequency)
  - favor tuning lines at the end of the amplifier
    - eliminate/minimize effect of the tune on noise



Tune -> balance, result

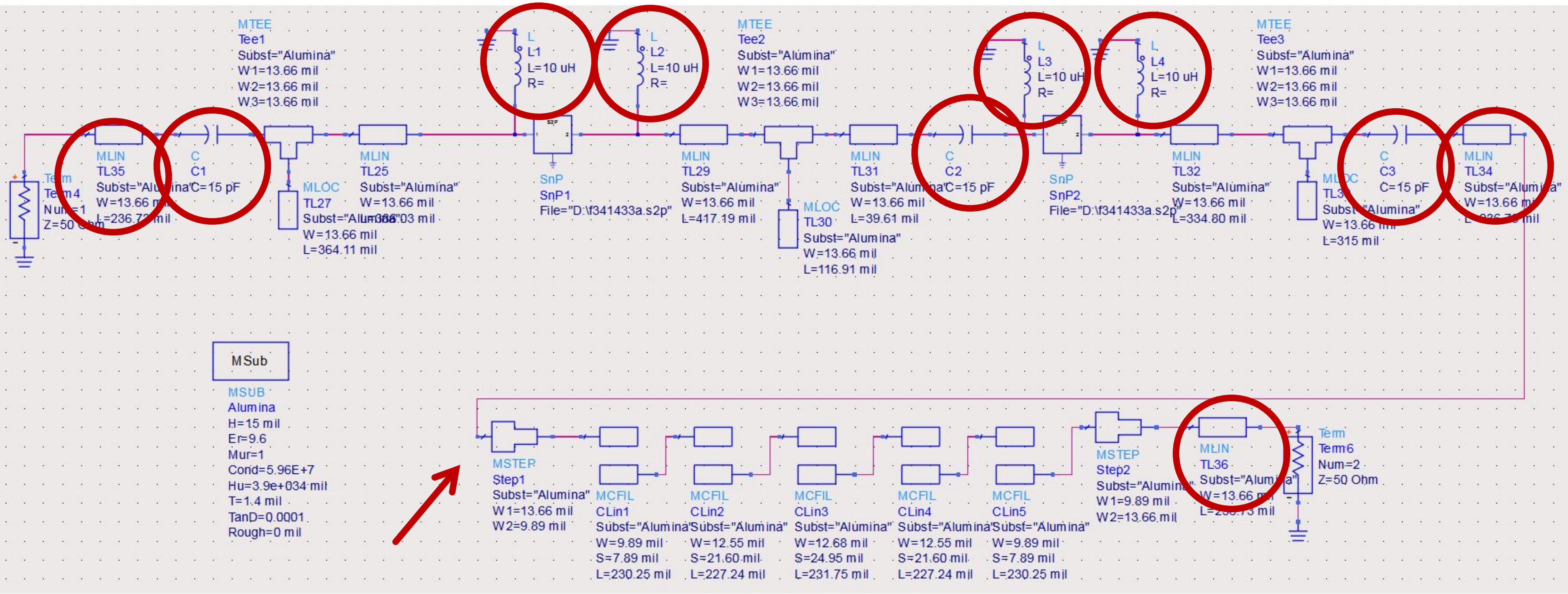


Amplifier, Filter, Total

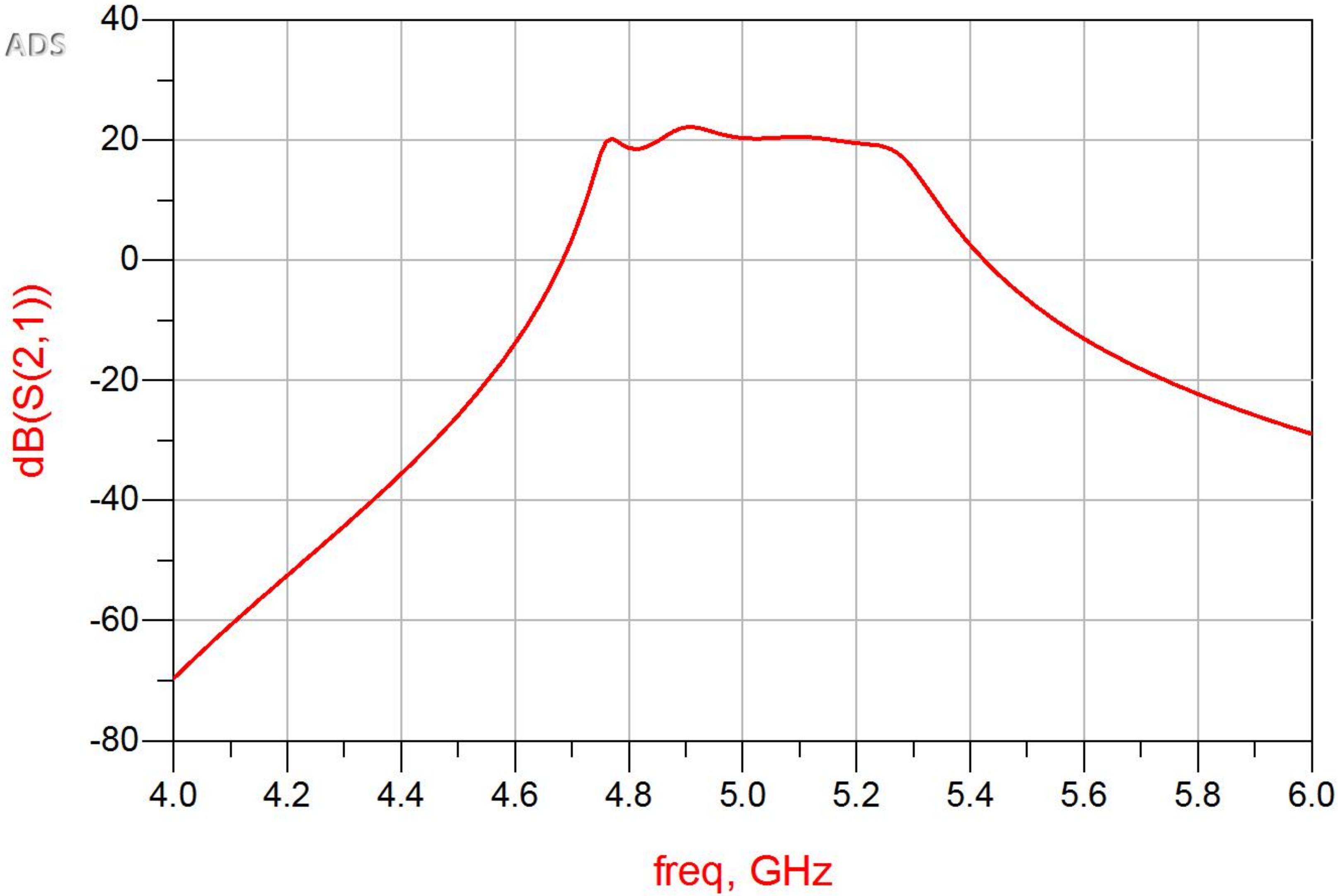


DC Bias elements in ADS schematic

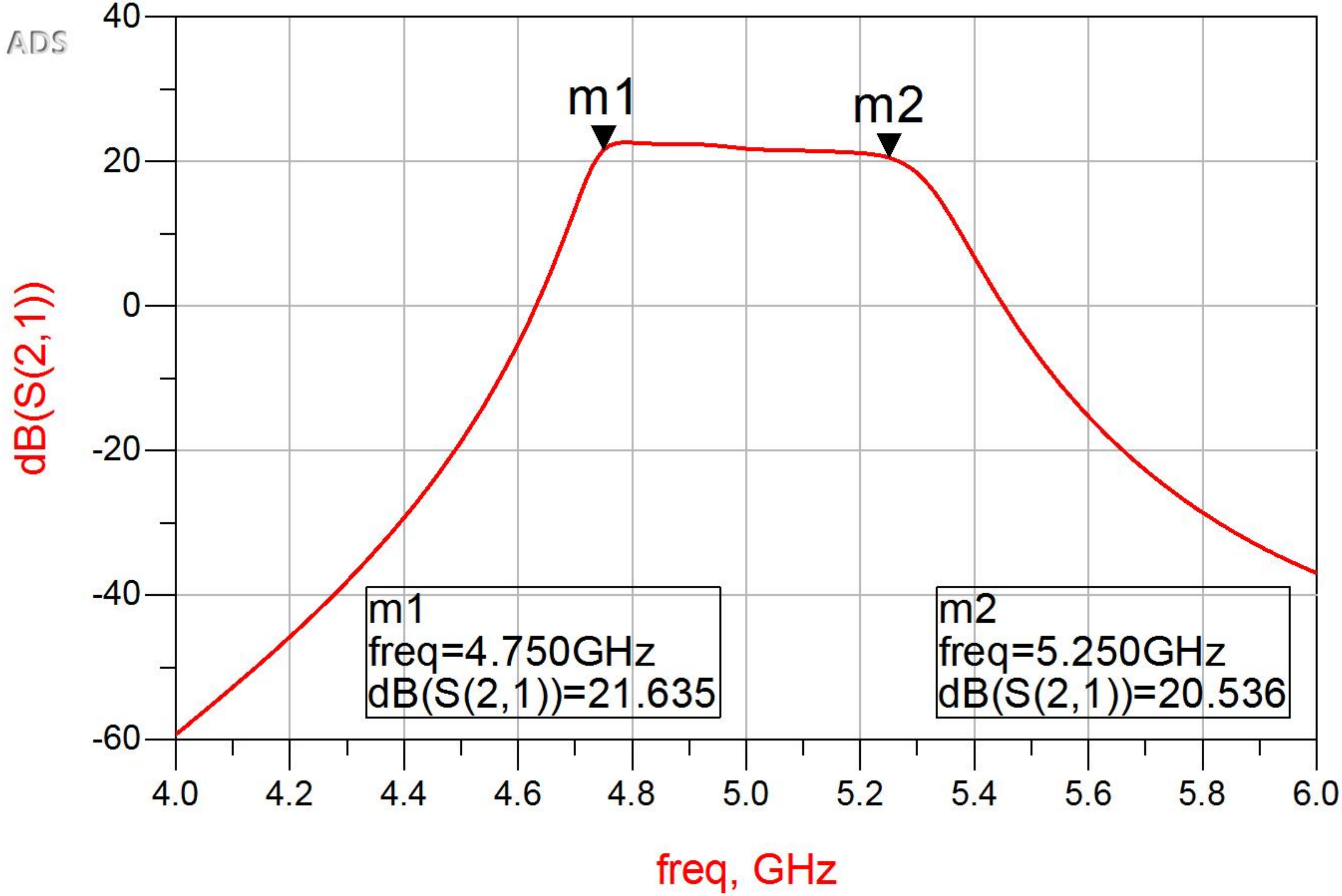
- Insert L (RF chokes) and C (decoupling)
- additional 50Ω connection lines
  - source
  - load
  - between blocks



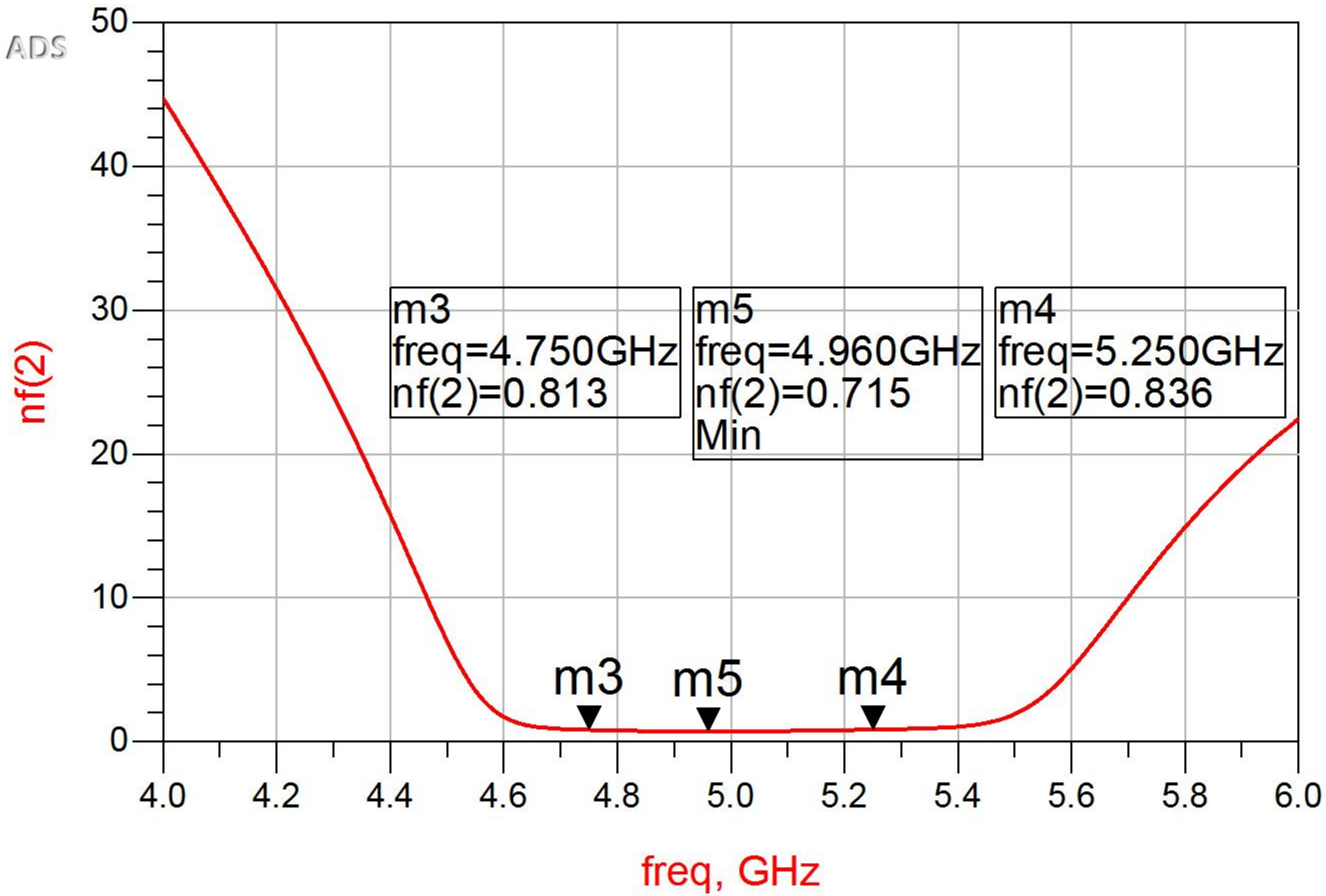
Gain -> Tune/Optimization



Final result (Gain)

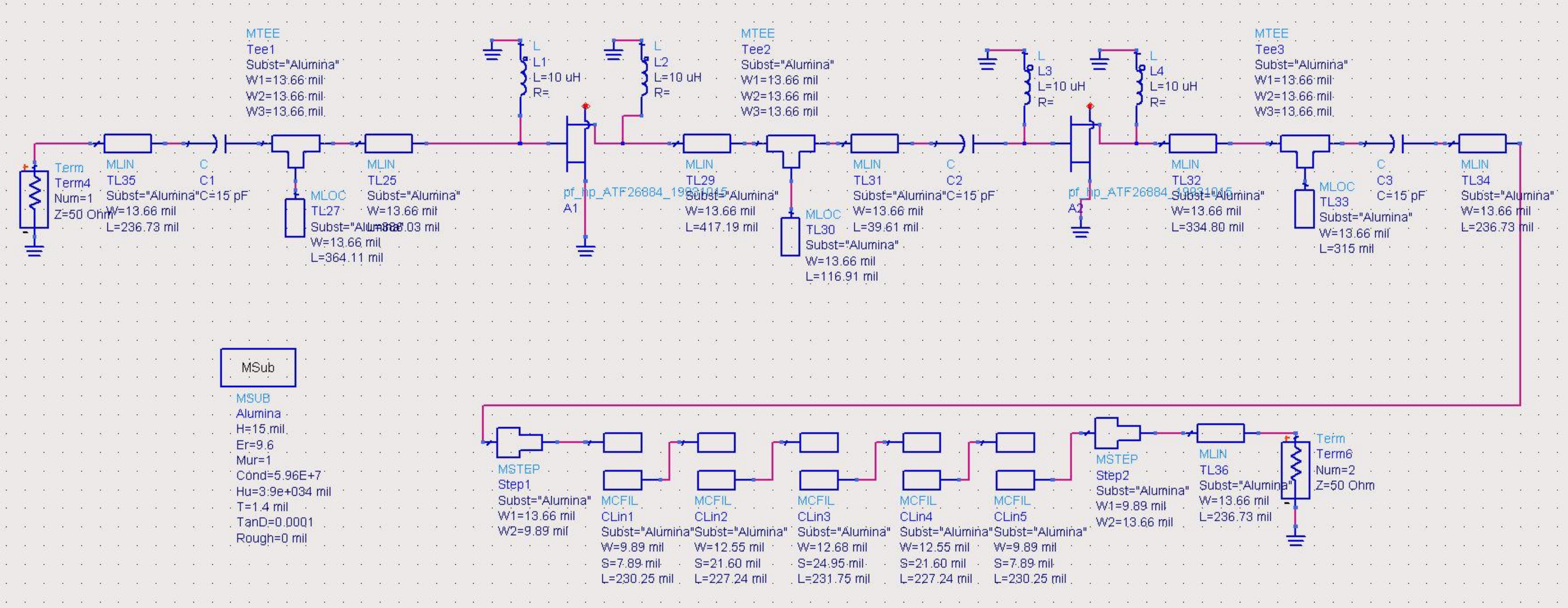


Final result (Noise)

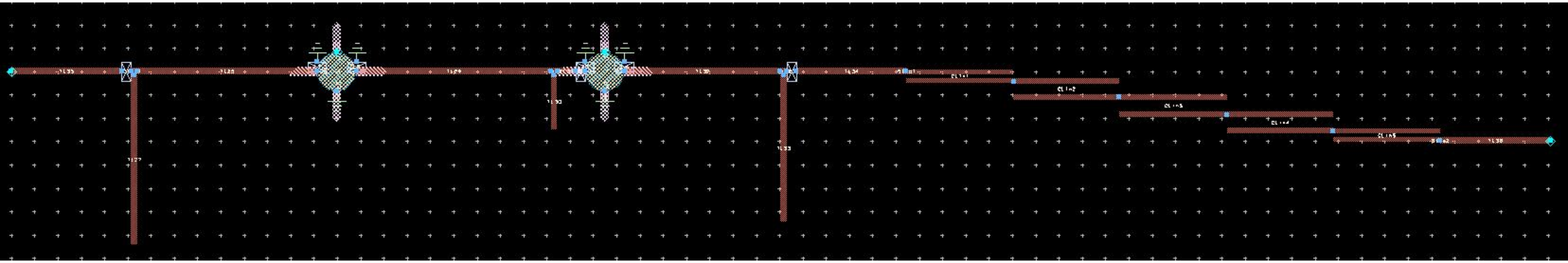


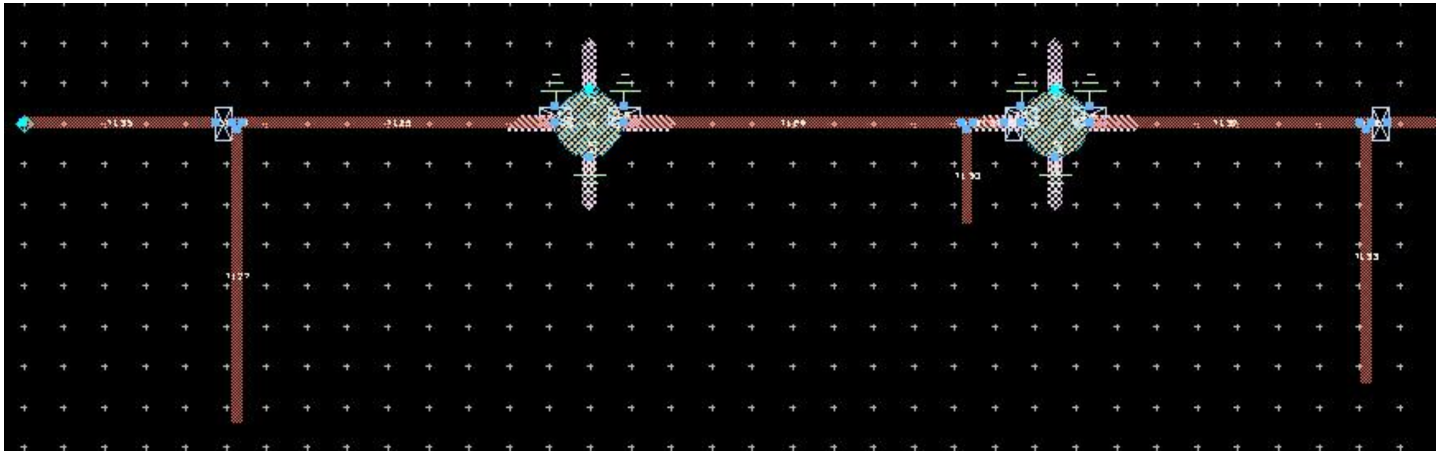
Layout (Example)

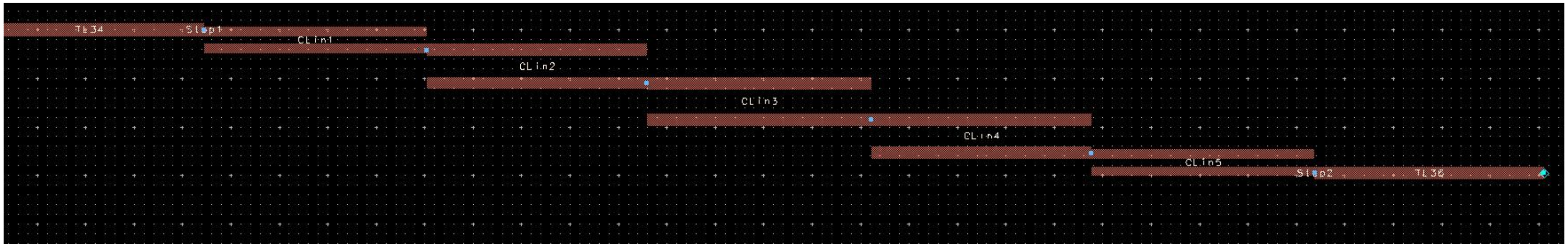
- Temporary replacement of the transistors and lumped elements (LC) with elements for which ADS has case information



Layout (Example)







Contact

- Microwave and Optoelectronics Laboratory
- **https://rf-opto.etti.tuiasi.ro**
- **rdamian@etti.tuiasi.ro**