

Lecture 12

2017/2018

Microwave Devices and Circuits for Radiocommunications

Materials

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

Software

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



Date:

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

[Acceseaza ca acest student](#) | [Cere acces la licente](#)

Note obtinute

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

Nume

Email

Cod de verificare

Trimite

Software

Advanced Design System Premier High-Frequency and High Speed Design Platform

2017

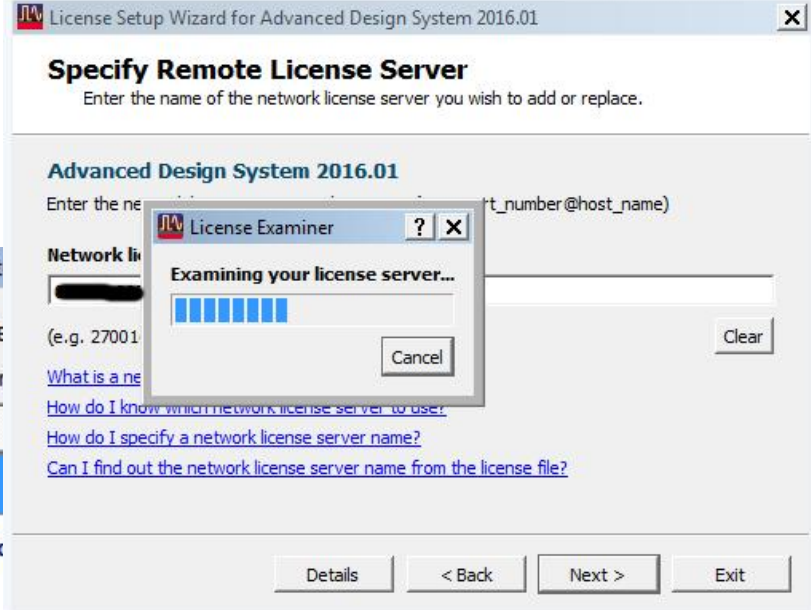
 **KEYSIGHT**
TECHNOLOGIES

Advanced Design System Premier High-Frequency and High Speed Design Platform

2016.01

 **KEYSIGHT**
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Advanced Design System

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Description

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Legend:  **License available**  License in use or not available

ADS Inclusive

▶ **License is available**

Number of licenses: 50

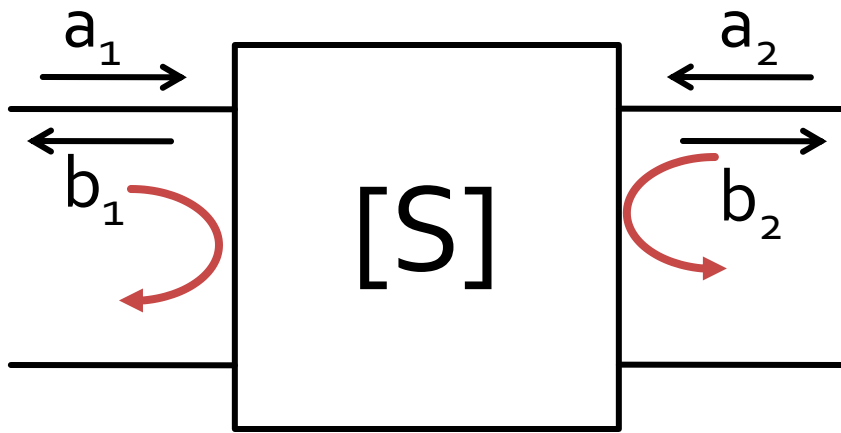
Used: 3

Version: 2018.07

Expires: 30-jul-

Microwave Network Analysis

Scattering matrix – S

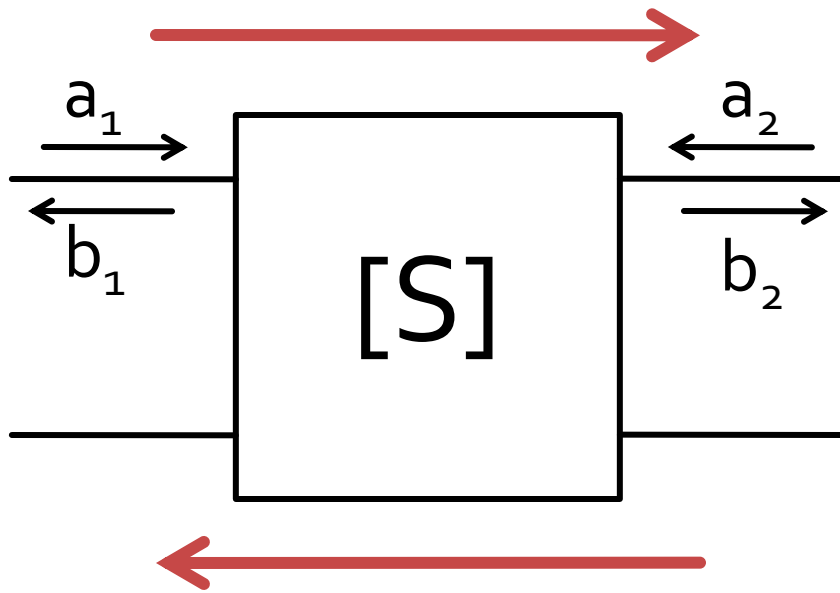


$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \cdot \begin{bmatrix} a_1 \\ a_2 \end{bmatrix}$$

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

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

Scattering matrix – S

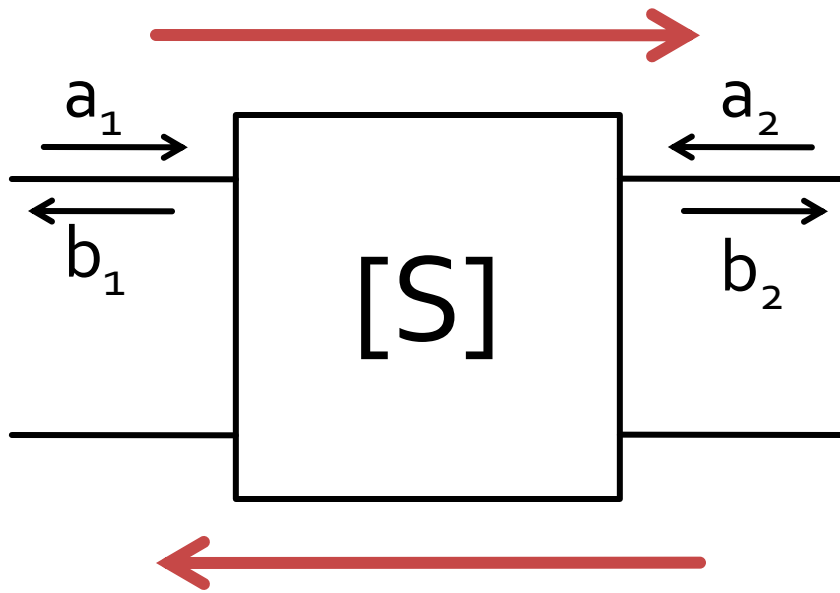


$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \cdot \begin{bmatrix} a_1 \\ a_2 \end{bmatrix}$$

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

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

Scattering matrix – S



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

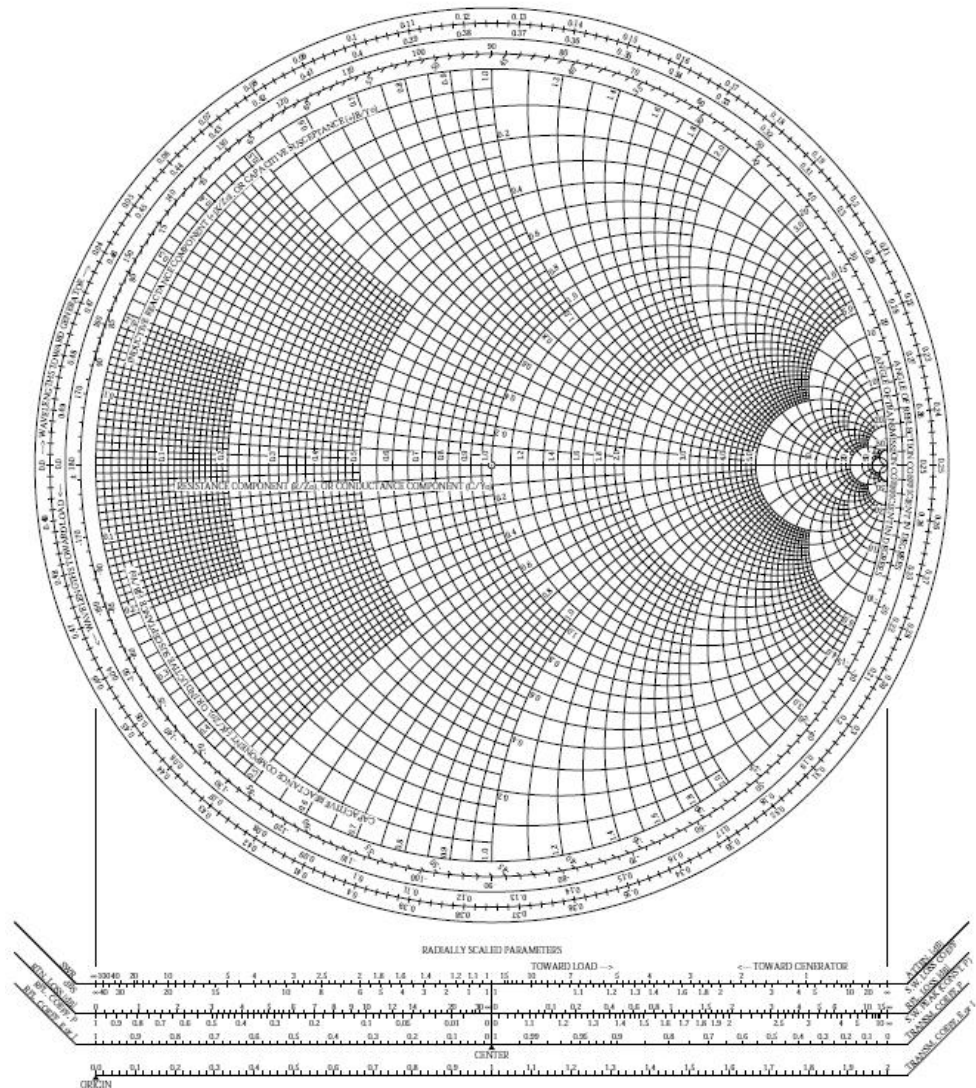
$$|S_{21}|^2 = \frac{\text{Power in } Z_0 \text{ load}}{\text{Power from } Z_0 \text{ source}}$$

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

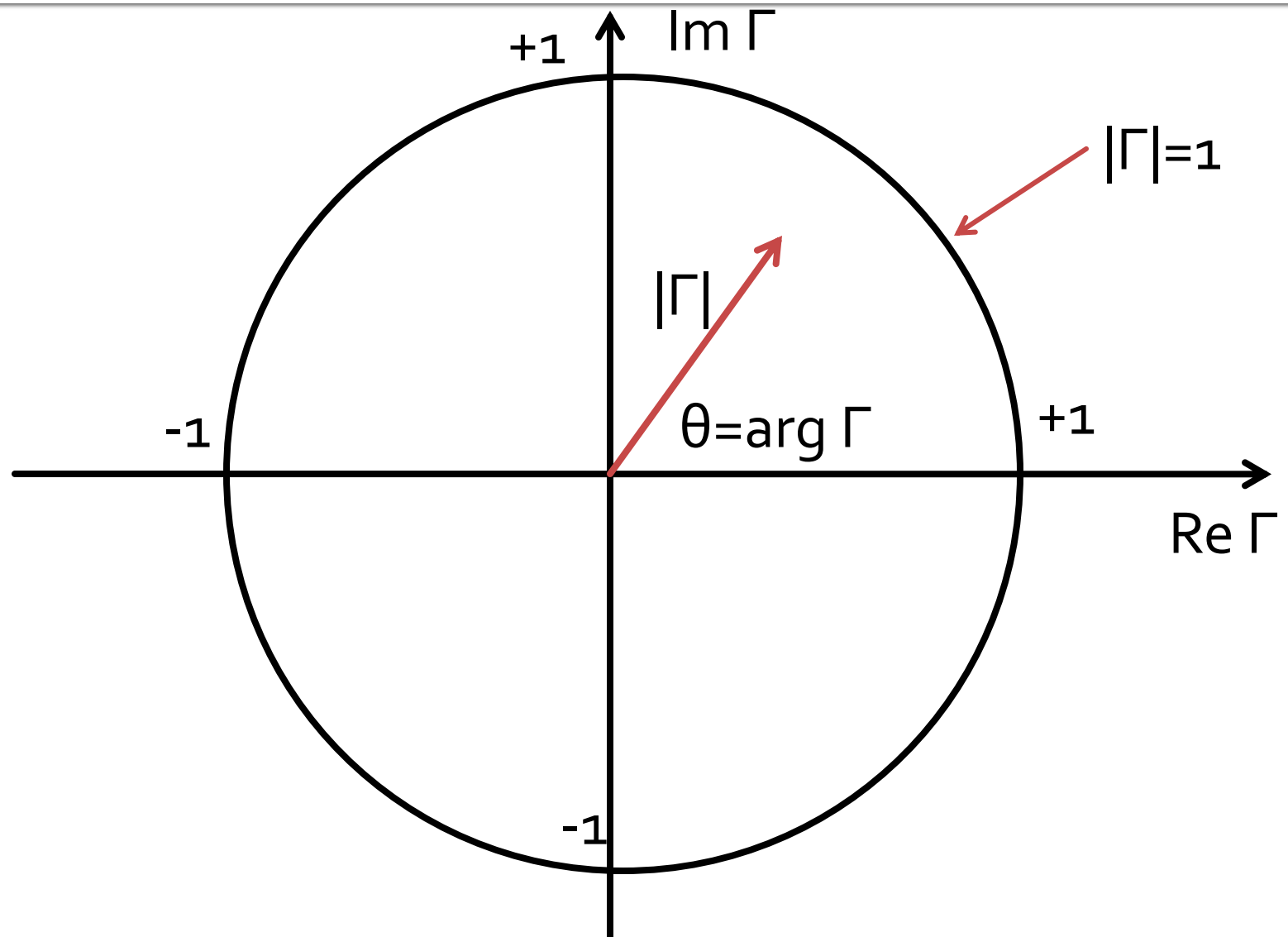
Impedance Matching

The Smith Chart

The Smith Chart



The Smith Chart

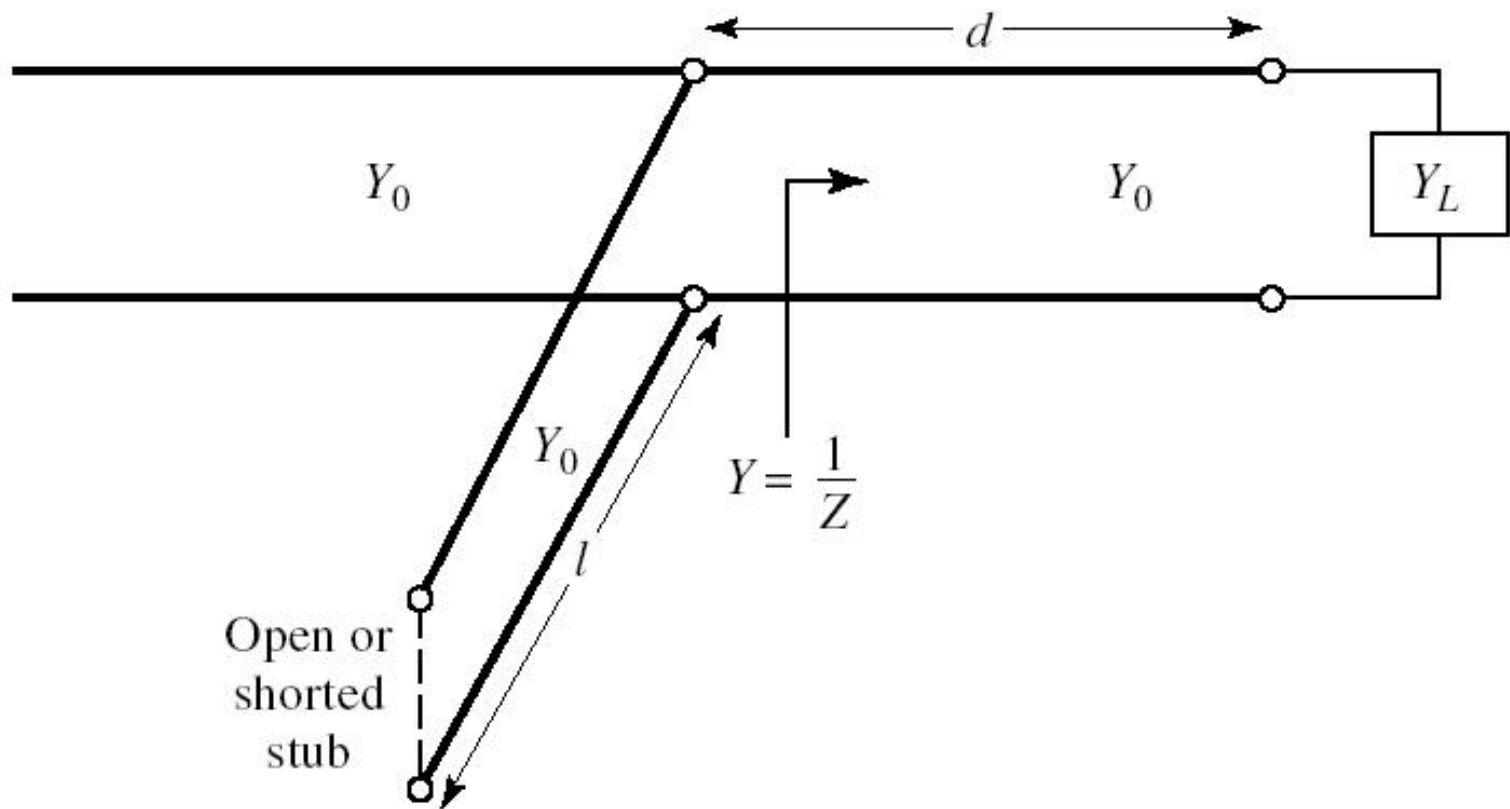


Impedance Matching with Stubs

Impedance Matching

Shunt Stub

- Shunt Stub



Analytical solution, Γ

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

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

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

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

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

- **"+" solution** ↓

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

$$\theta_{sp} = \tan^{-1}(\text{Im } y_s) = -55.8^\circ (+180^\circ) \rightarrow \theta_{sp} = 124.2^\circ$$

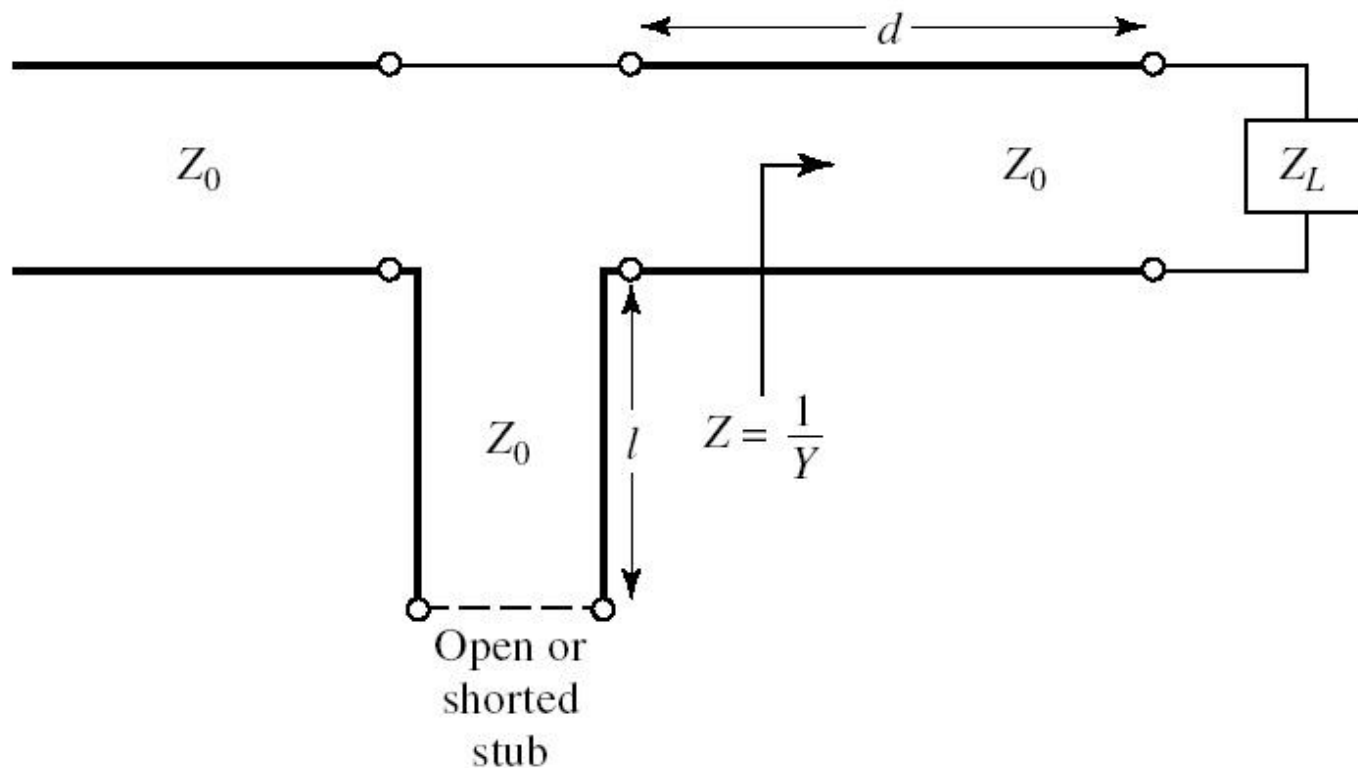
- **"-" solution** ↓

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

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

Series Stub

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



Analytical solution, Γ

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

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

$$\Gamma_s = 0.555 \angle -29.92^\circ$$

$$|\Gamma_s| = 0.555; \quad \varphi = -29.92^\circ \quad \cos(\varphi + 2\theta) = 0.555 \Rightarrow (\varphi + 2\theta) = \pm 56.28^\circ$$

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

- **"+" solution**

$$(-29.92^\circ + 2\theta) = +56.28^\circ \quad \theta = 43.1^\circ \quad \text{Im } z_s = \frac{+2 \cdot |\Gamma_s|}{\sqrt{1 - |\Gamma_s|^2}} = +1.335$$

$$\theta_{ss} = -\cot^{-1}(\text{Im } z_s) = -36.8^\circ (+180^\circ) \rightarrow \theta_{ss} = 143.2^\circ$$

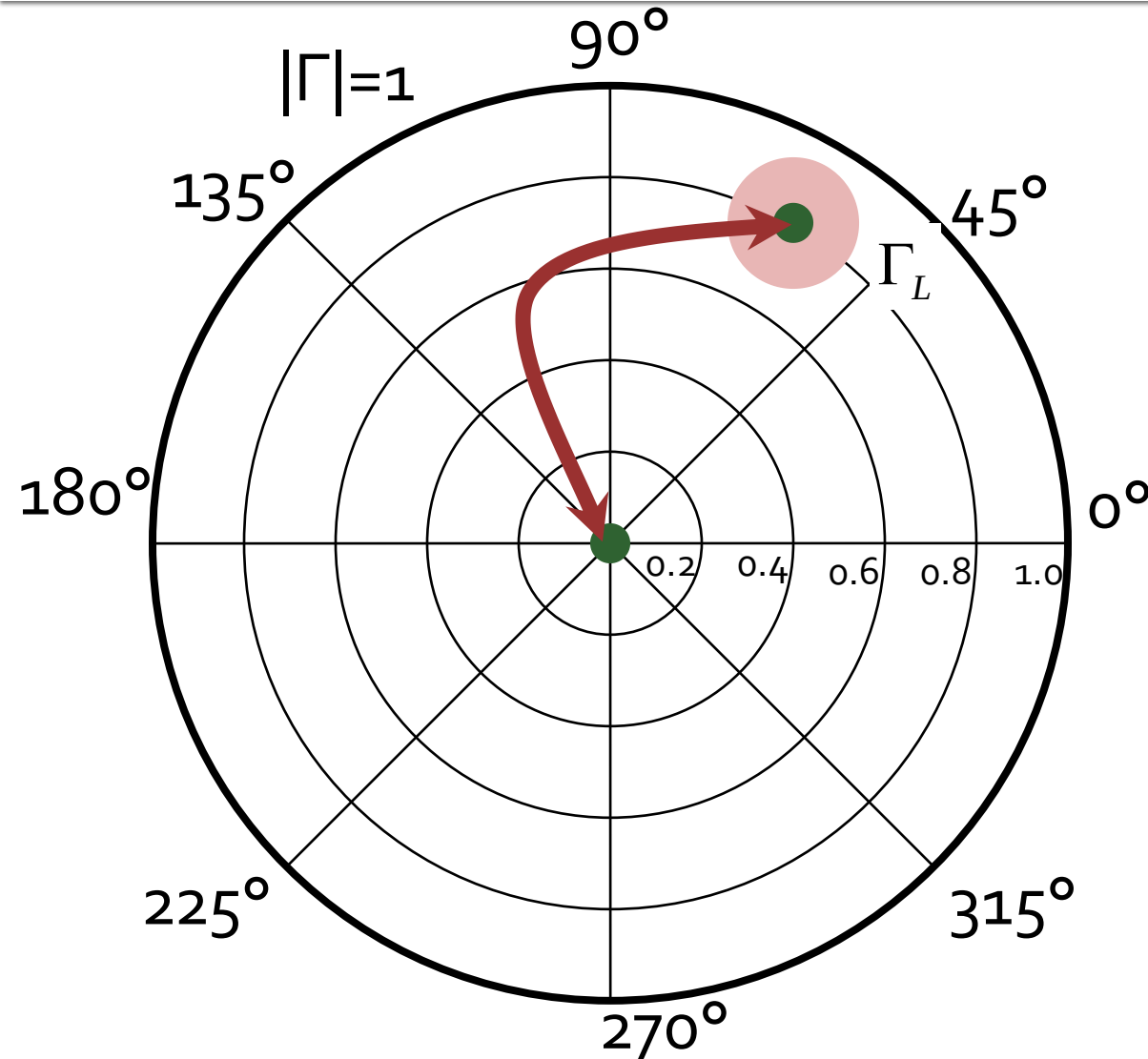
- **"-" solution**

$$(-29.92^\circ + 2\theta) = -56.28^\circ \quad \theta = -13.2^\circ (+180^\circ) \rightarrow \theta = 166.8^\circ$$

$$\text{Im } z_s = \frac{-2 \cdot |\Gamma_s|}{\sqrt{1 - |\Gamma_s|^2}} = -1.335 \quad \theta_{ss} = -\cot^{-1}(\text{Im } z_s) = 36.8^\circ$$

The Smith Chart, matching ,

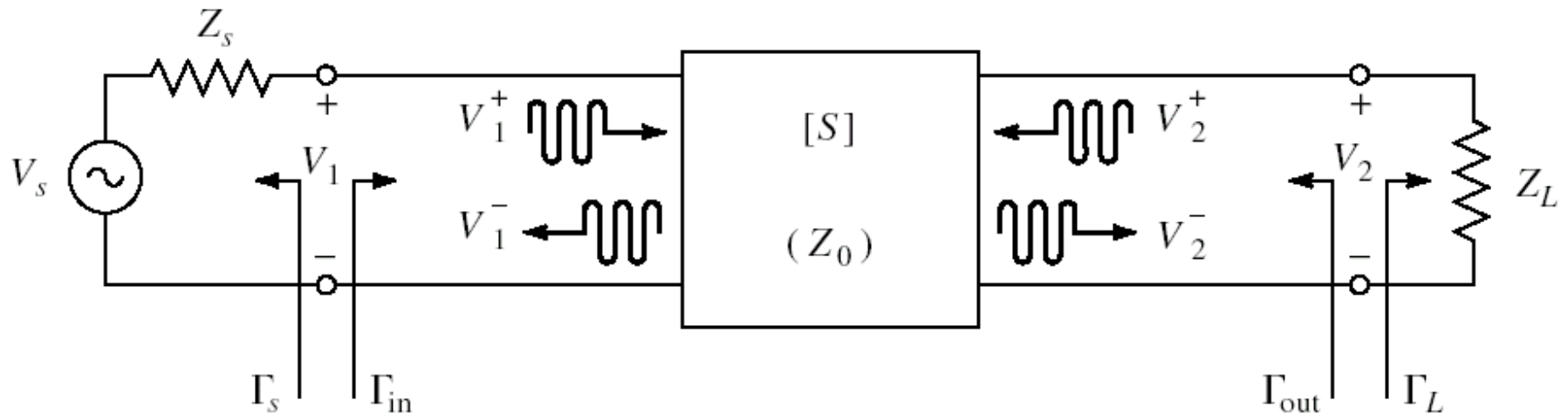
$Z_L \neq Z_o, Z_L = Z_o$



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

Microwave Amplifiers

Amplifier as two-port



- Characterized with S parameters
- normalized at Z_0 (implicit 50Ω)
- Datasheets: S parameters for specific bias conditions

Datasheets

NE46100

VCE = 5 V, Ic = 50 mA

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

VCE = 5 V, Ic = 100 mA

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

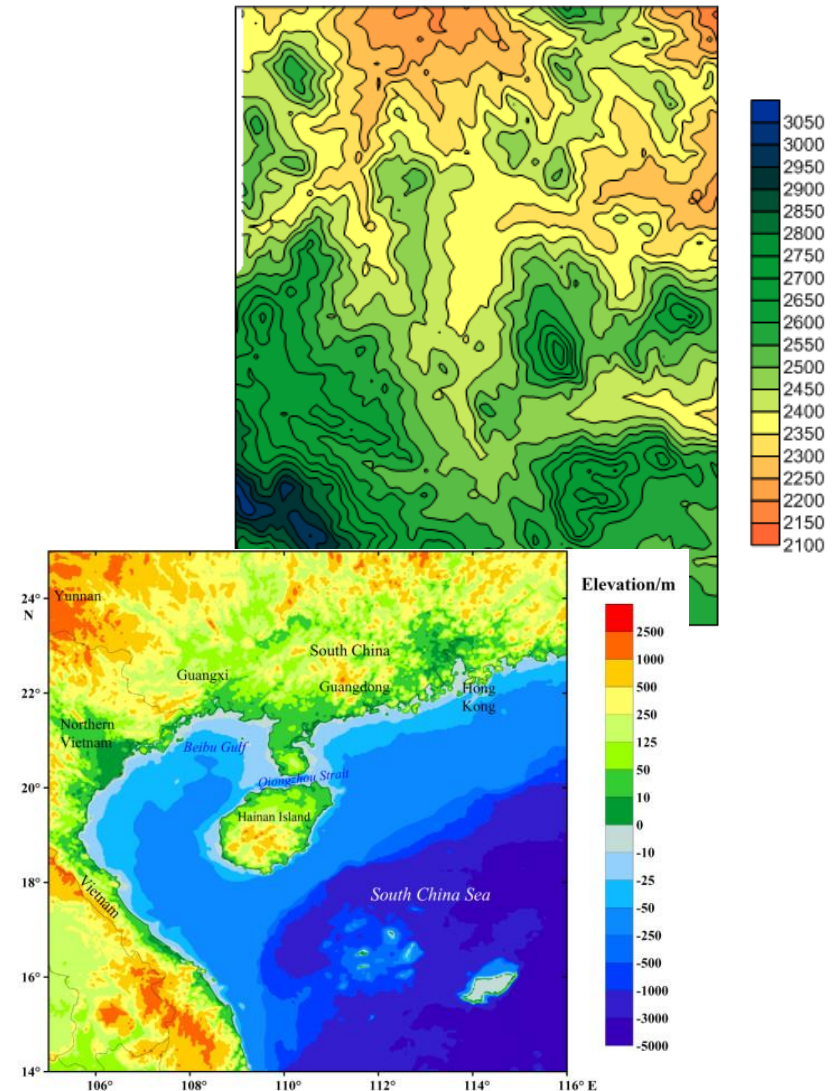
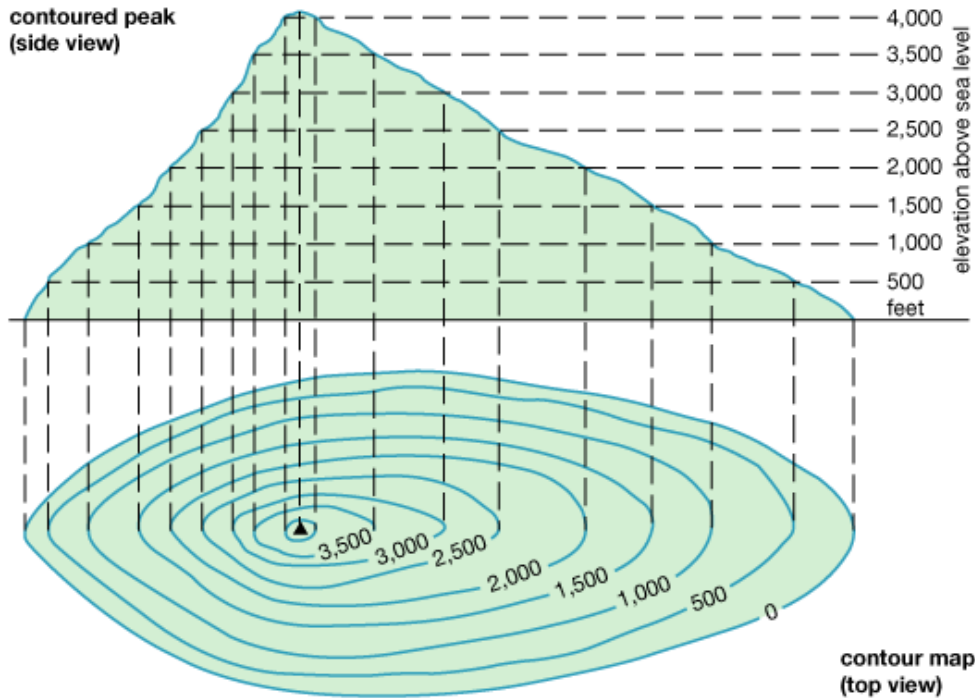
S2P - Touchstone

- Touchstone file format (*.s2p)

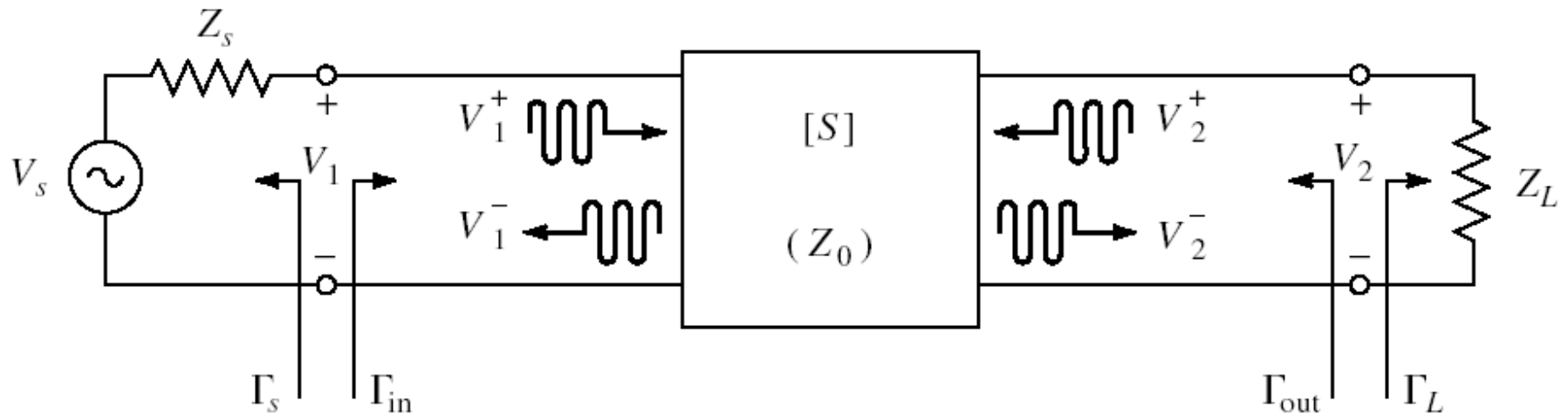
```
! SIEMENS Small Signal Semiconductors
! VDS = 3.5 V   ID = 15 mA
# GHz S MA R 50
! f      S11      S21      S12      S22
! GHz   MAG ANG   MAG ANG   MAG ANG   MAG ANG
1.000 0.9800 -18.0 2.230 157.0 0.0240 74.0 0.6900 -15.0
2.000 0.9500 -39.0 2.220 136.0 0.0450 57.0 0.6600 -30.0
3.000 0.8900 -64.0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

Contour map/lines

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



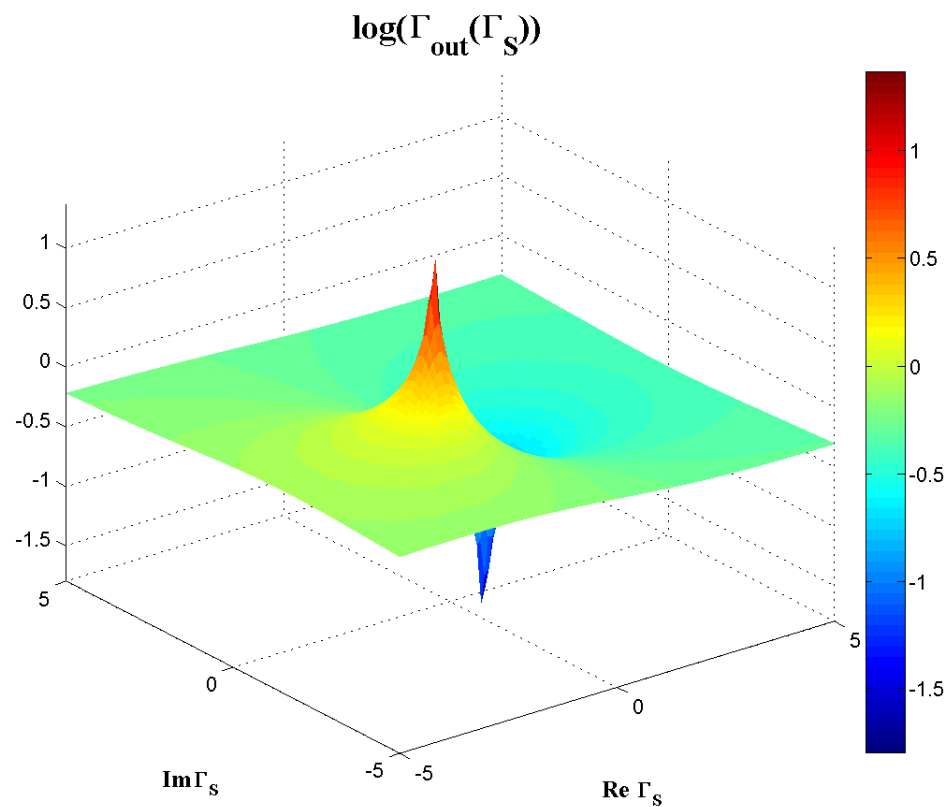
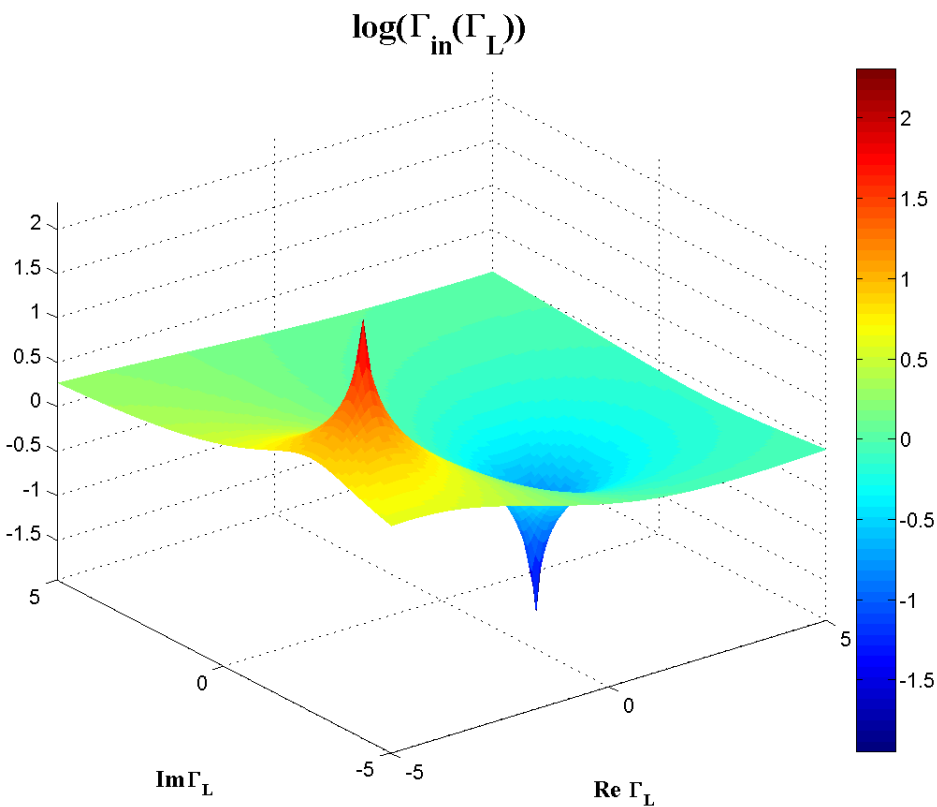
Amplifier as two-port



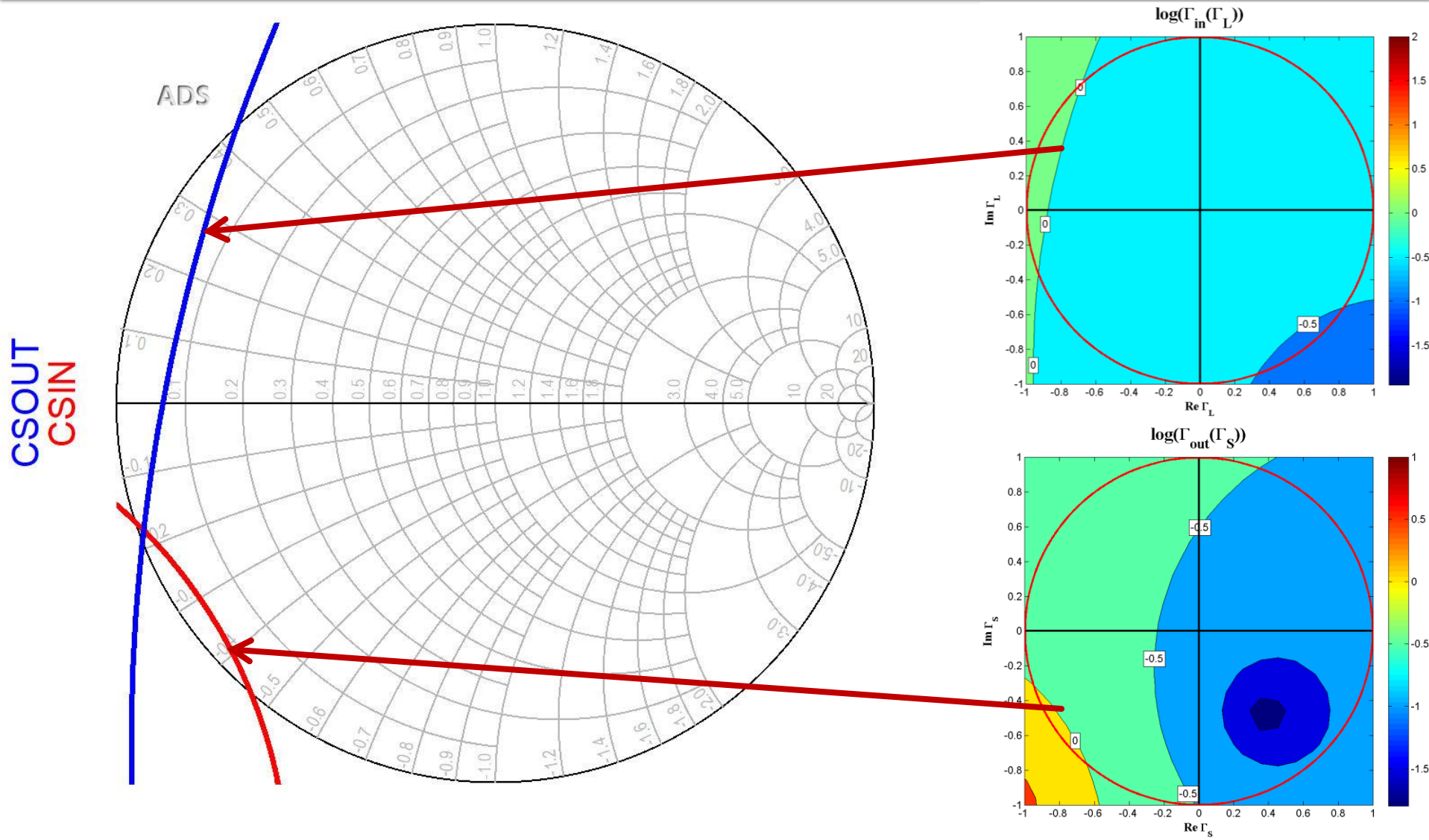
- For an amplifier two-port we are interested in:
 - **stability**
 - power gain
 - noise (sometimes – small signals)
 - linearity (sometimes – large signals)

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

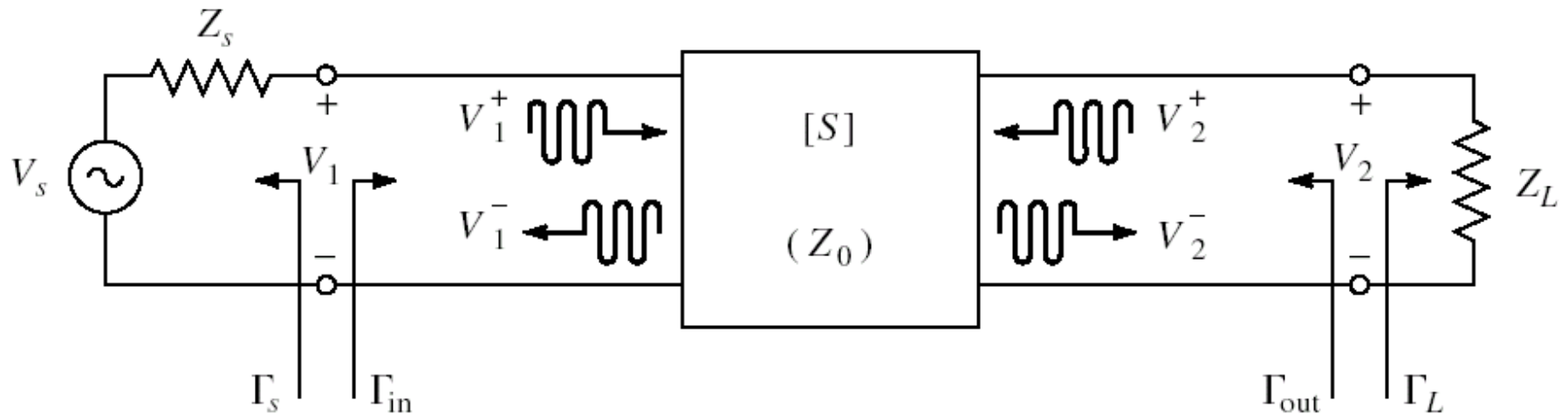
■ $\log_{10}|\Gamma_{in}|, \log_{10}|\Gamma_{out}|$



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

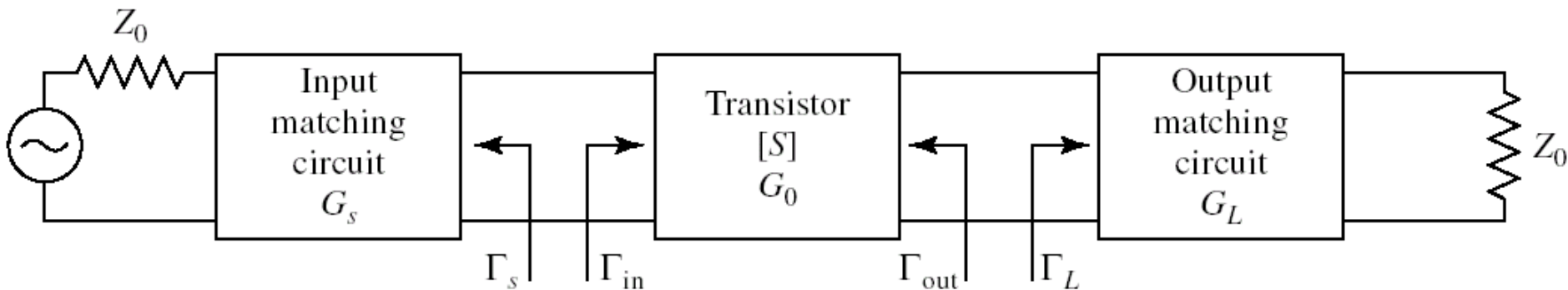


Amplifier as two-port



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

Design for Specified Gain



- In the unilateral assumption:

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

$$G_s = \frac{1 - |\Gamma_s|^2}{|1 - S_{11} \cdot \Gamma_s|^2}$$

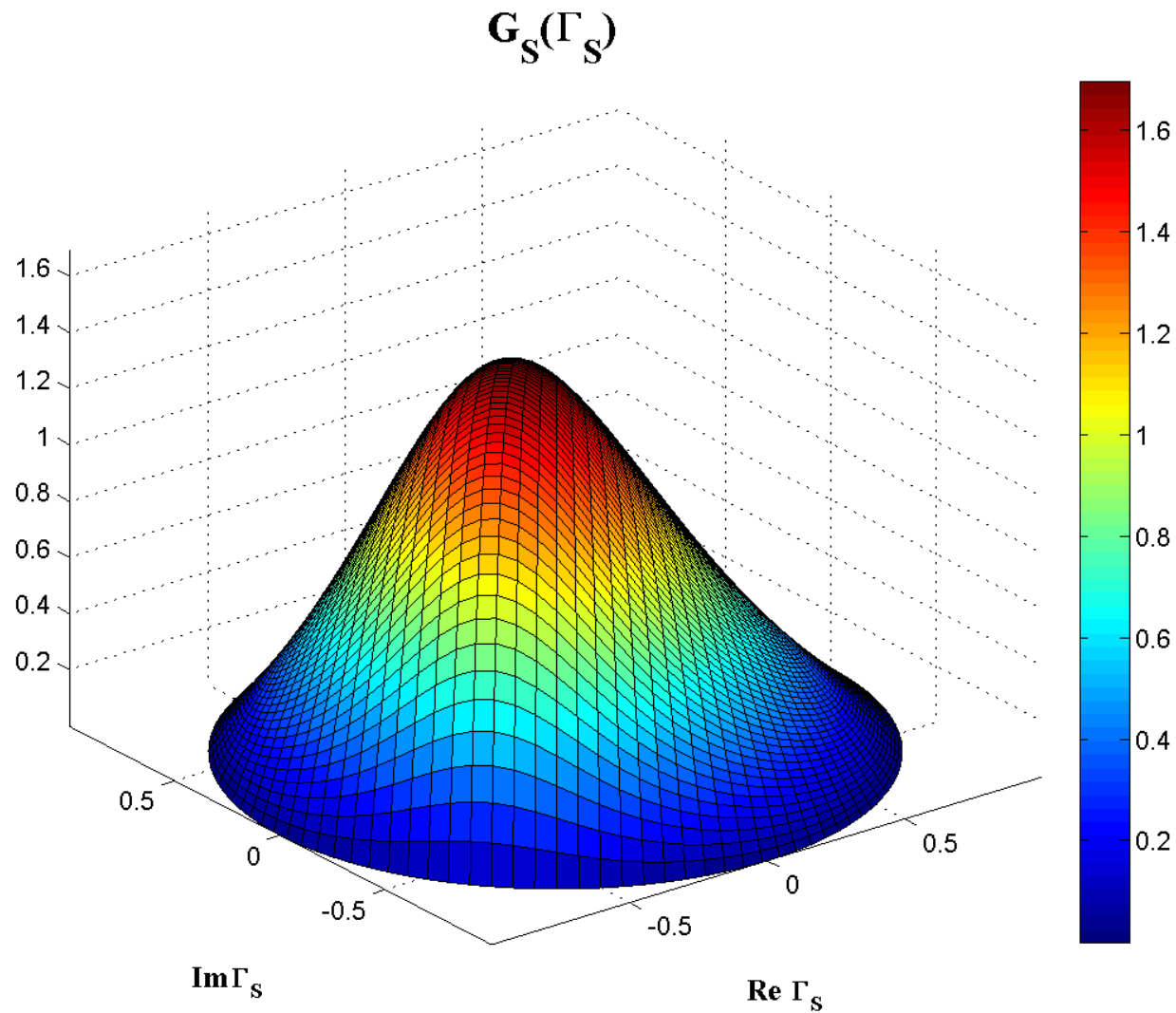
$$G_s = G_s(\Gamma_s)$$

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

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

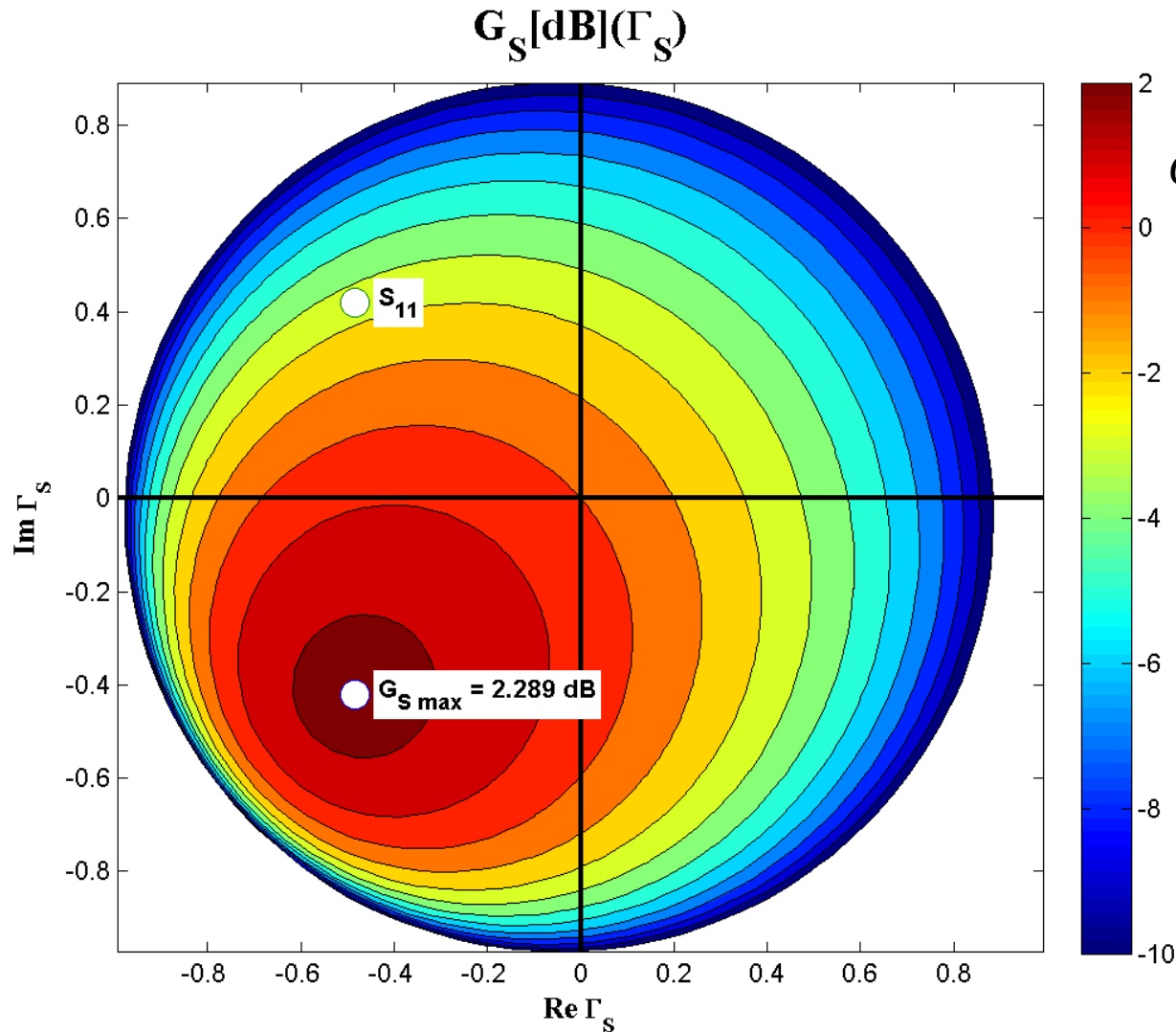
$$G_L = G_L(\Gamma_L)$$

$$G_S(\Gamma_S)$$



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

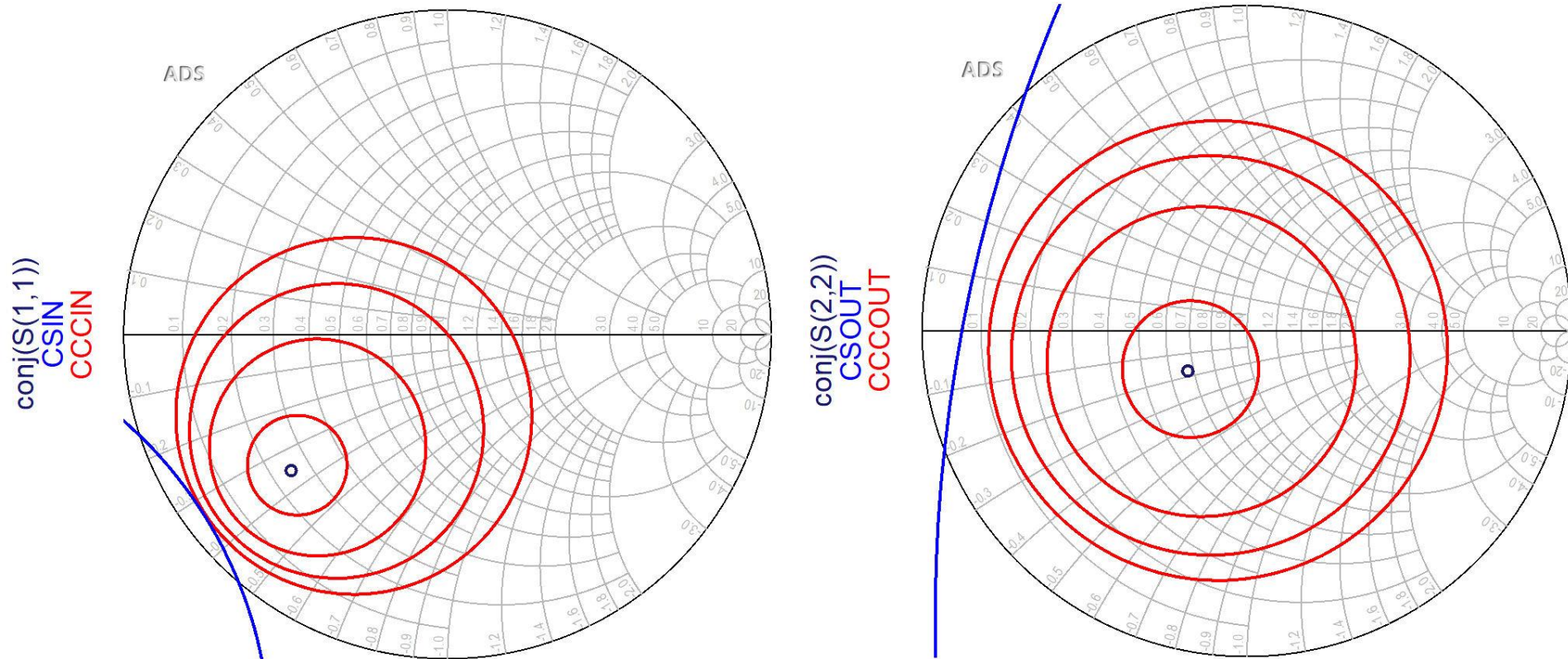
$G_S[\text{dB}](\Gamma_S)$, constant value contours



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

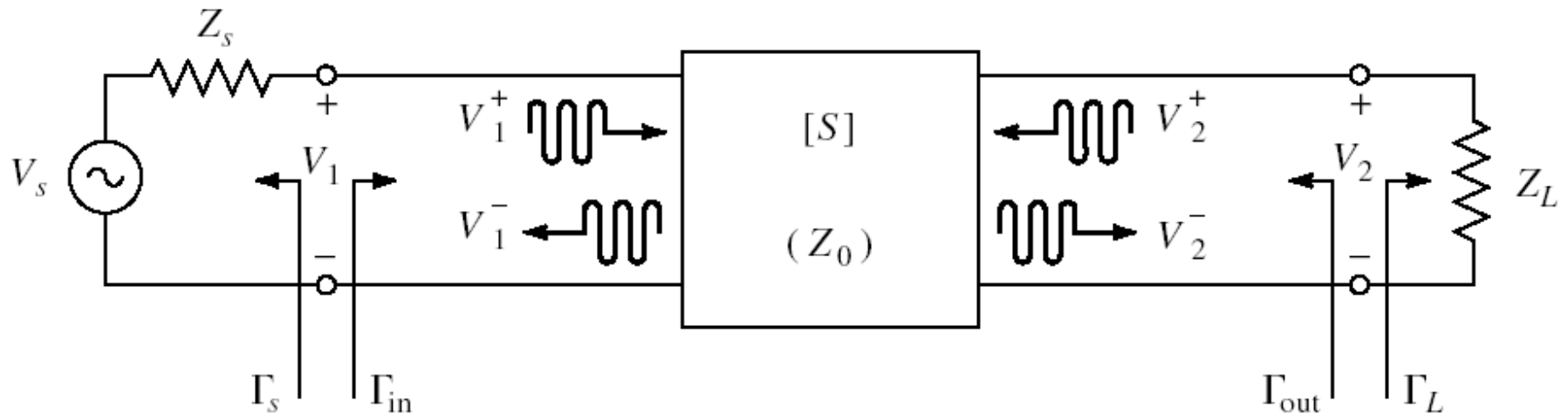
$$G_{S \max} = G_S|_{\Gamma_S = S_{11}^*}$$

ADS



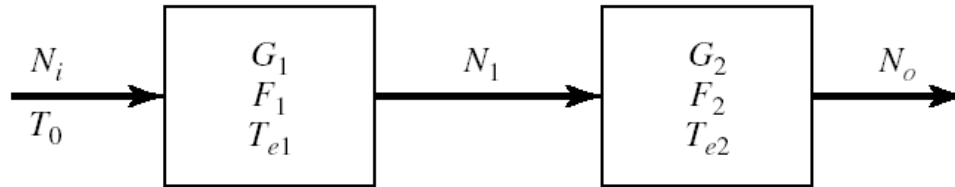
- Circles are plotted for requested values (**in dB!**)
- It is useful to compute $G_{S_{\max}}$ and $G_{L_{\max}}$ before
 - in order to request relevant circles

Amplifier as two-port

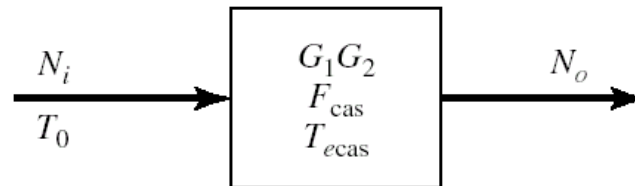


- For an amplifier two-port we are interested in:
 - stability
 - power gain
 - **noise** (sometimes – small signals)
 - linearity (sometimes – large signals)

Noise figure of a cascaded system



(a)



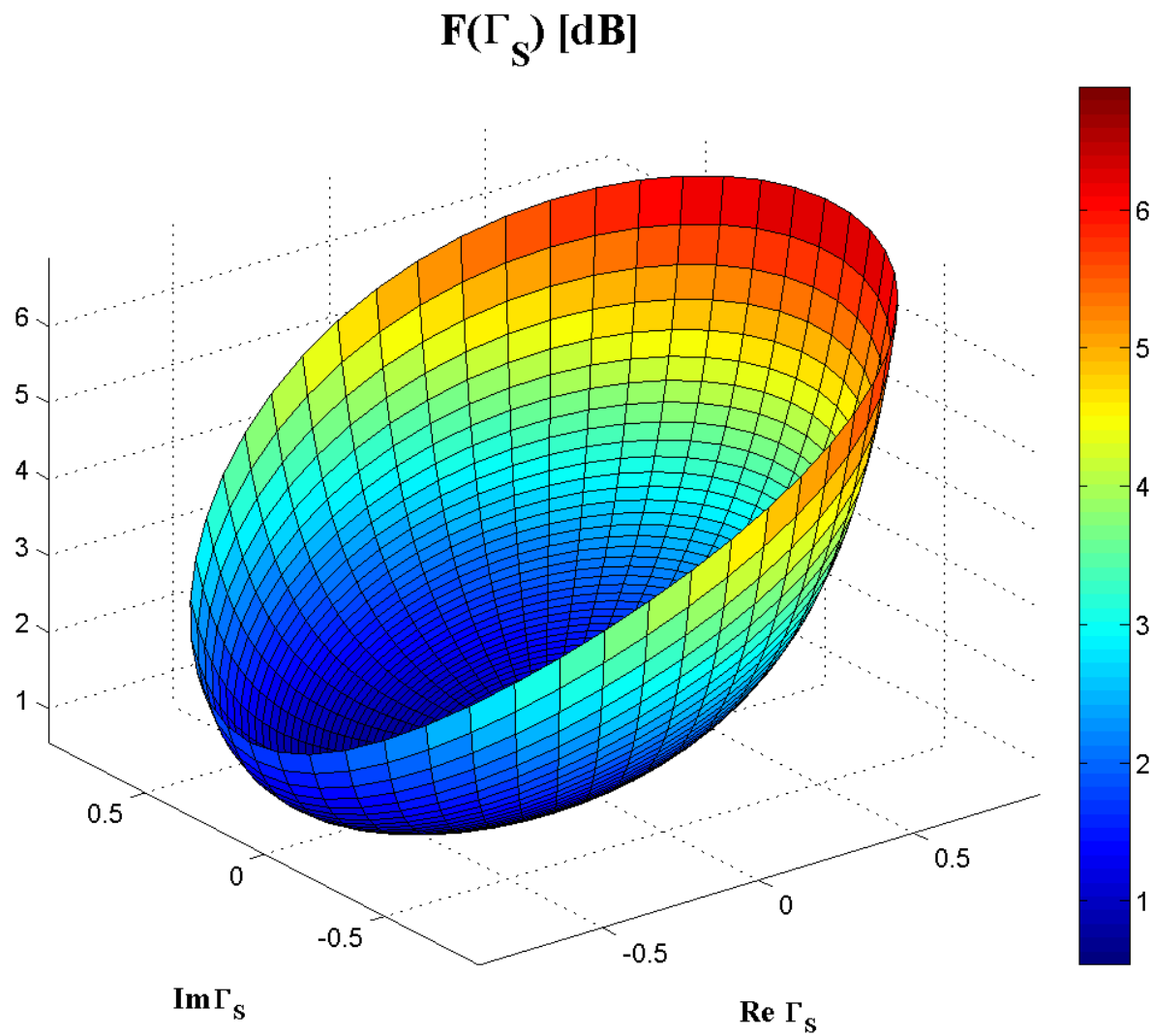
(b)

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

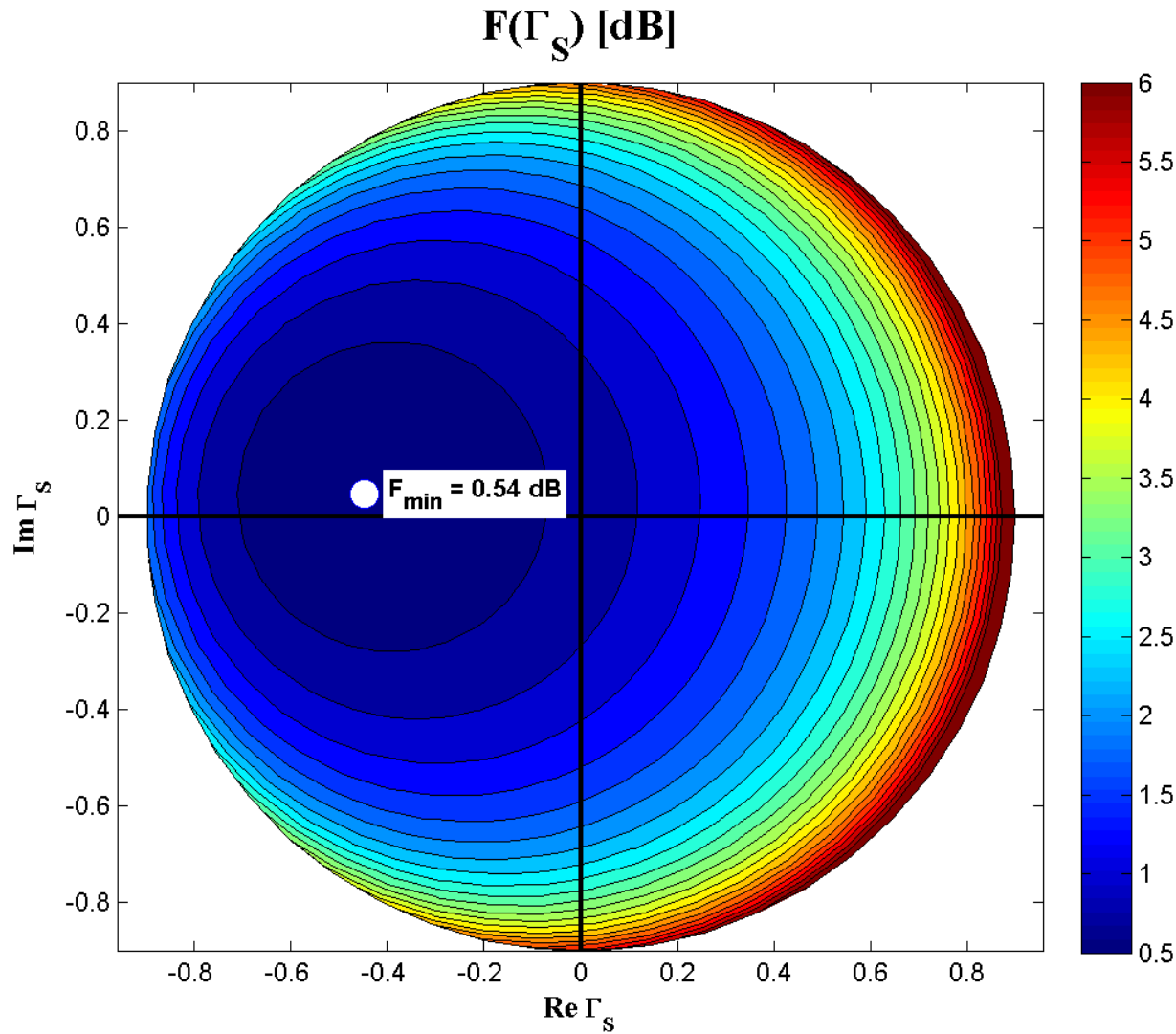
■ Friis Formula (!linear scale)

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

$F[\text{dB}](\Gamma_s)$

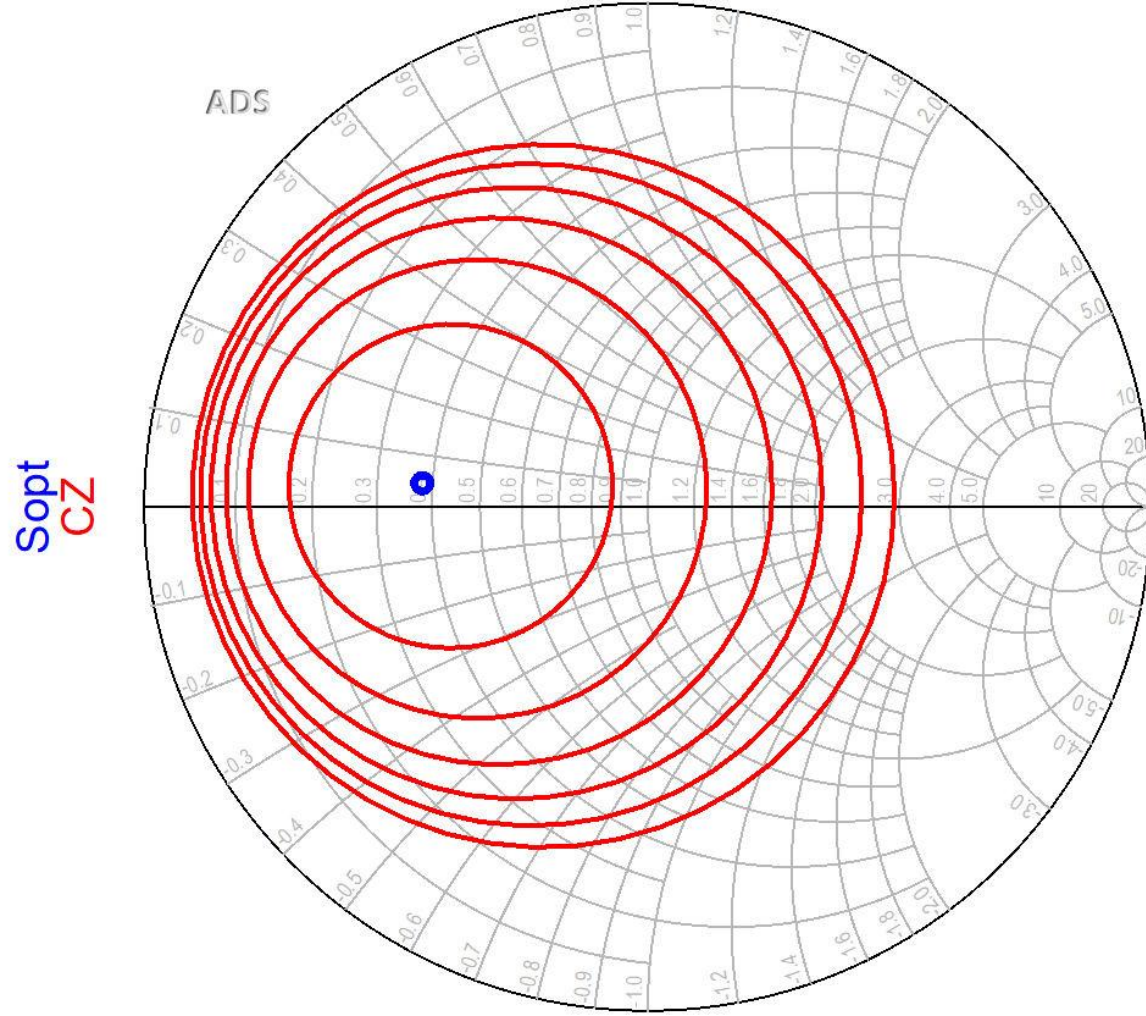


$G_s[\text{dB}](\Gamma_s)$, constant value contours



$$\Gamma_{\text{opt}} = 0.45 \angle 174^\circ$$

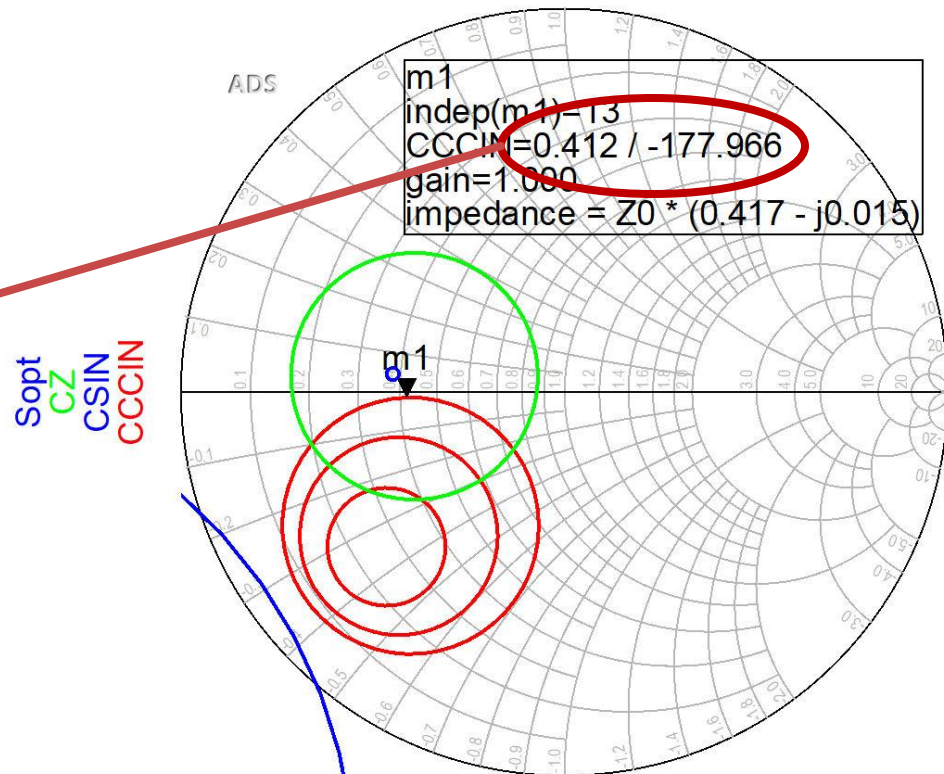
ADS



Matching – 2

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

$$\Gamma_s = 0.412 \angle -177.966^\circ$$

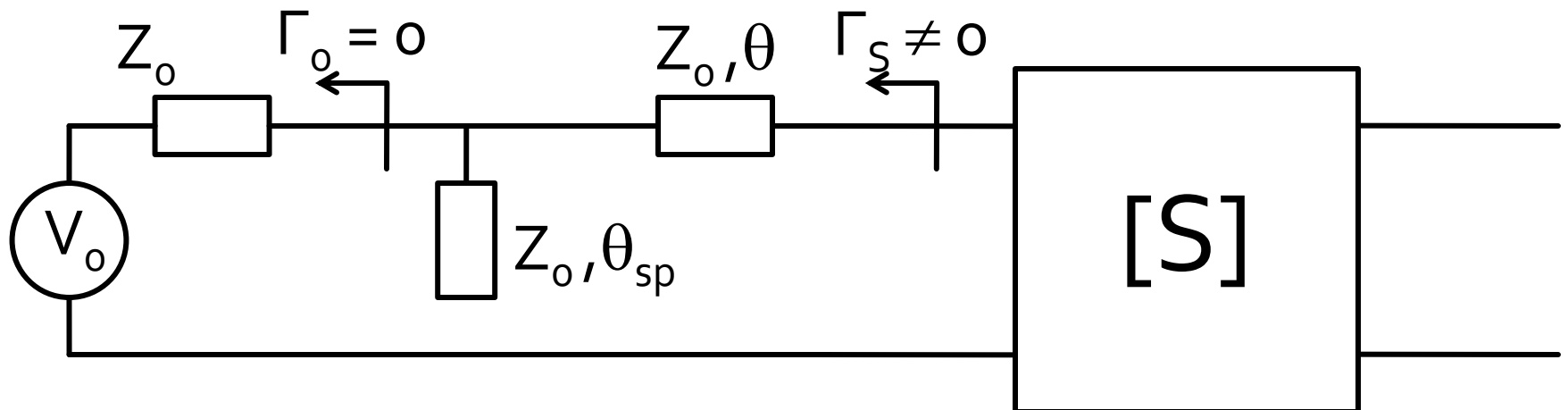


Matching – 5

- Computation depends solely on Γ_s (magnitude and phase)

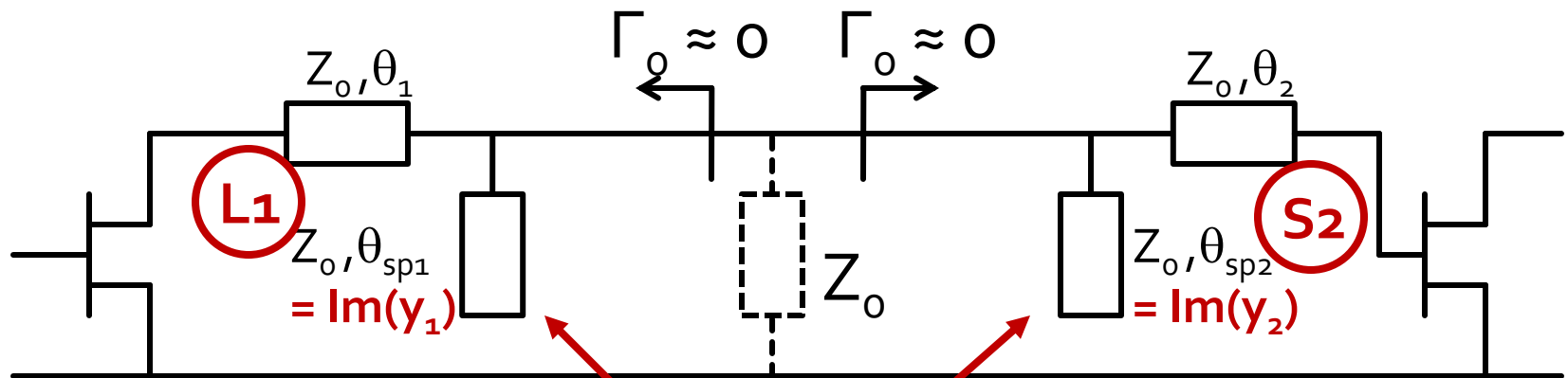
$$\cos(\varphi_s + 2\theta) = -|\Gamma_s| \quad \tan \theta_{sp} = \frac{\mp 2 \cdot |\Gamma_s|}{\sqrt{1 - |\Gamma_s|^2}}$$

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



Adaptare inter-etaje 2

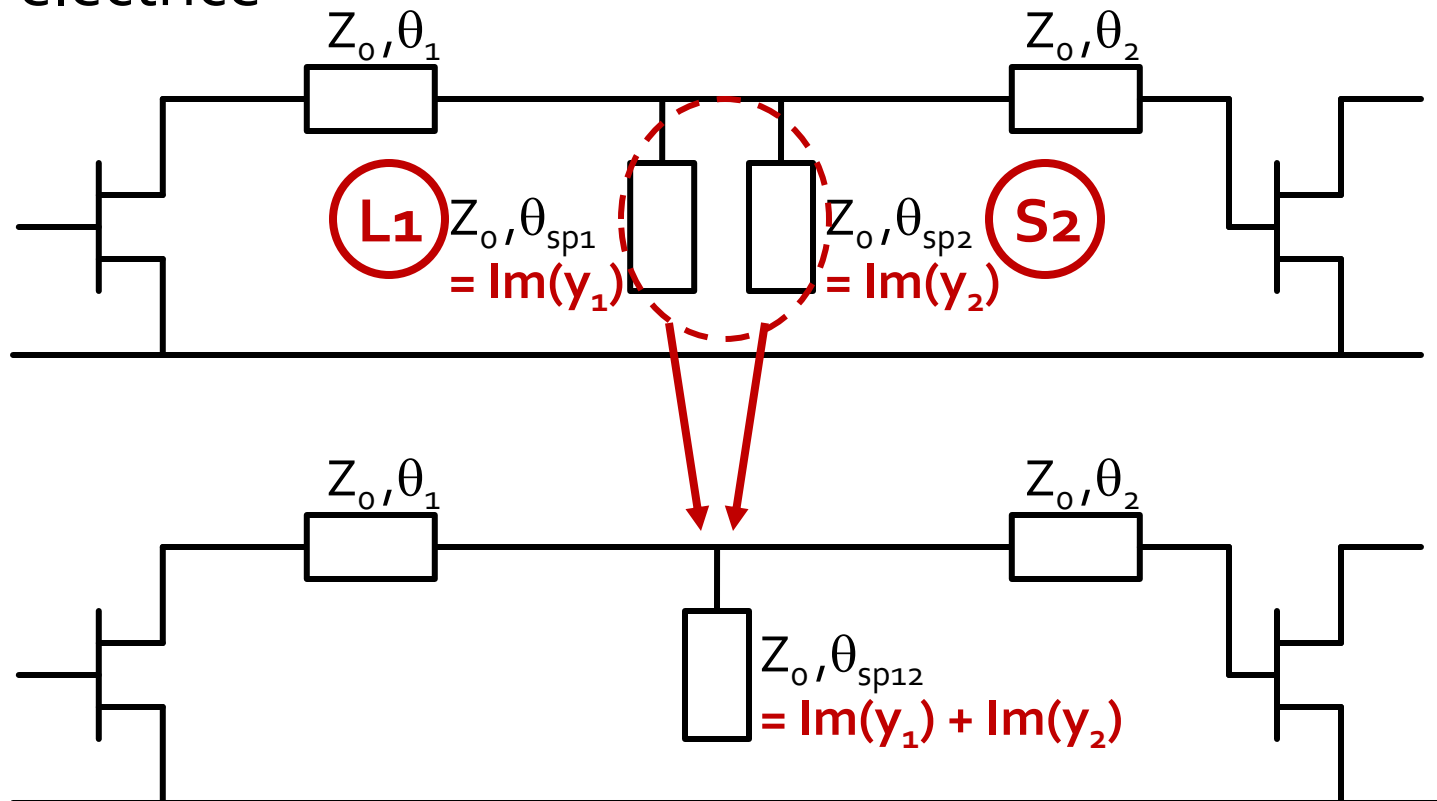
- Unul din etaje creaza prin reteaua sa de adaptare un coeficient de reflexie $\Gamma=0$ la care apoi se adapteaza celalalt etaj



Cele doua stub-uri in paralel
se combina intr-unul singur

Adaptare inter-etaje 2

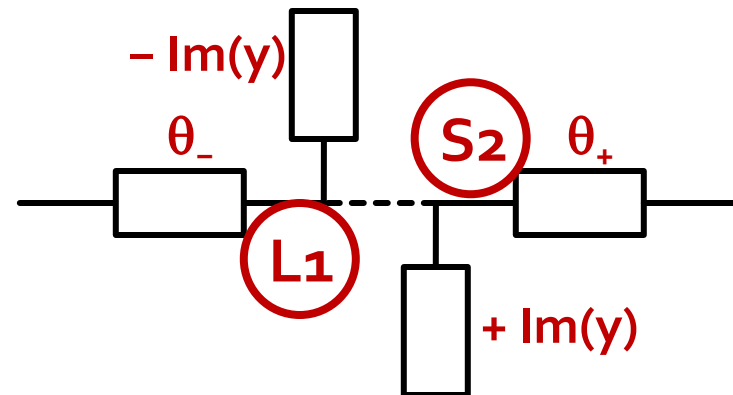
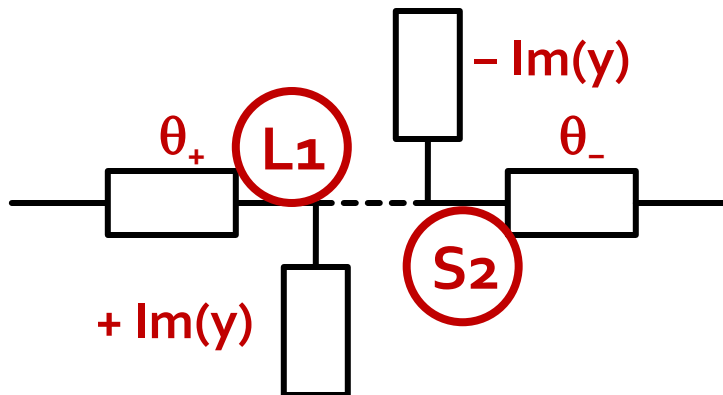
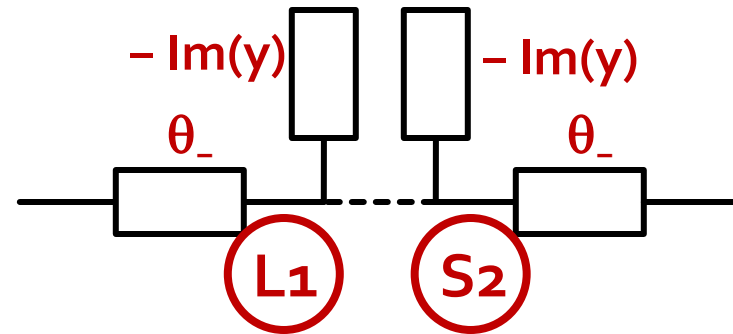
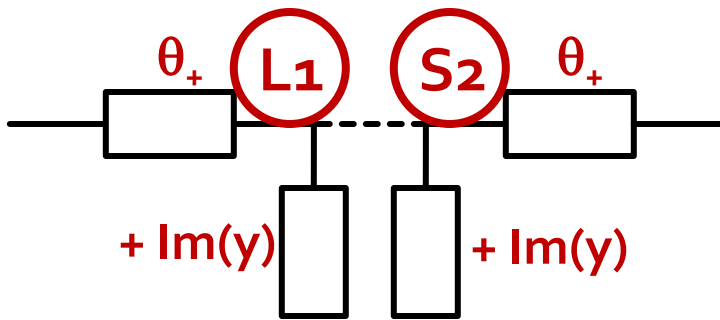
- Cele doua stub-uri in paralel se combina intr-unul singur
 - **admitantele** sunt in paralel si se aduna, **nu** lungimile electrice



Calcul analitic

- 4 combinatii posibile
 - admitantele** sunt in paralel si se aduna, **nu** lungimile electrice

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



Filtre pentru microunde

Metoda pierderilor de insertie

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

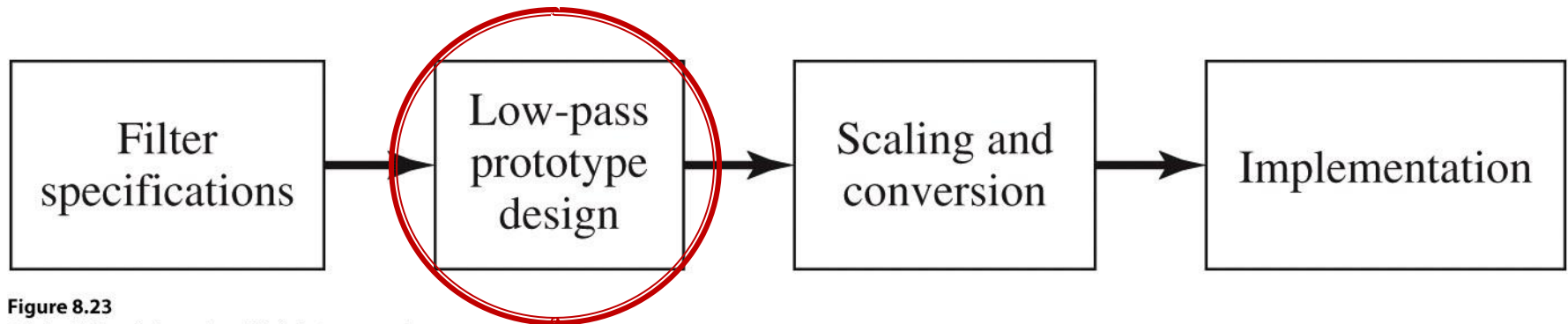


Figure 8.23

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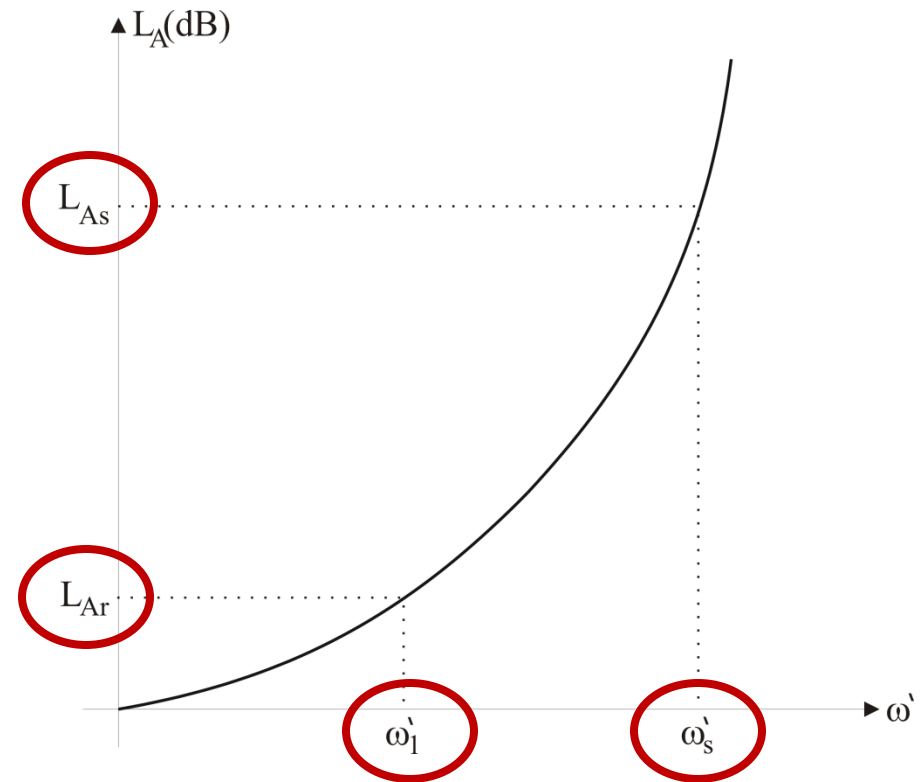
Caracteristici de filtru trece jos prototip

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

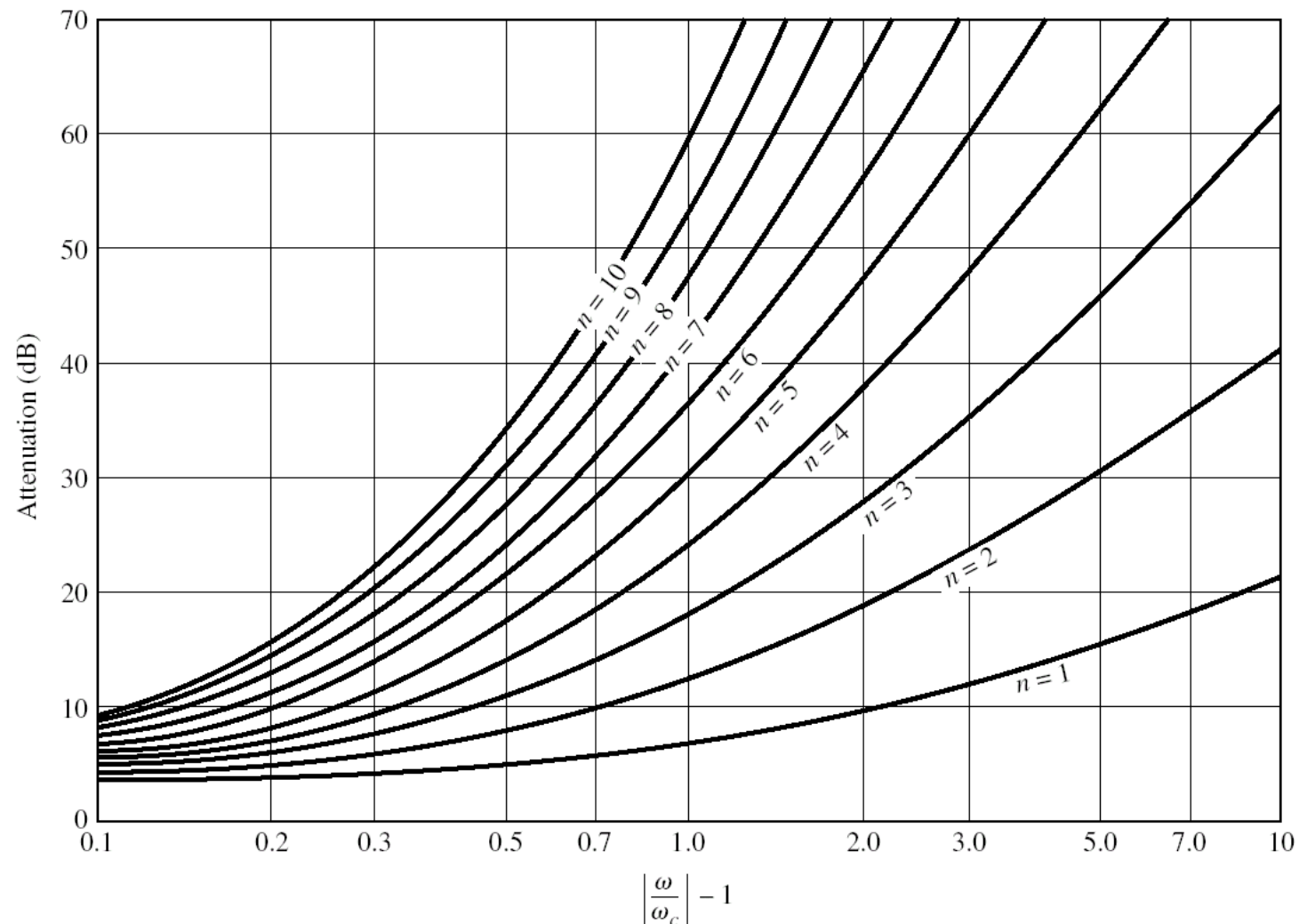
Calculul ordinului filtrului maxim plat

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

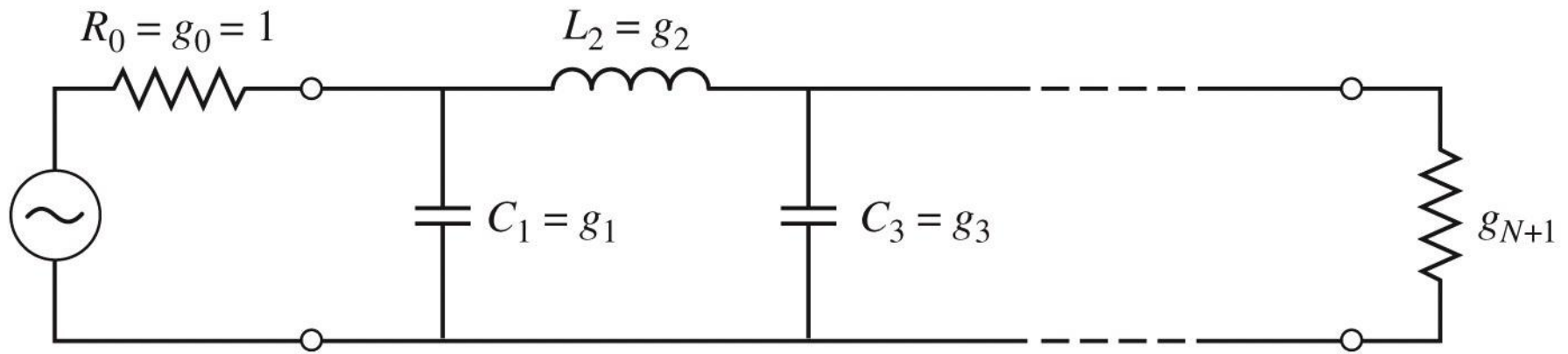
- cu atenuarile in **dB**



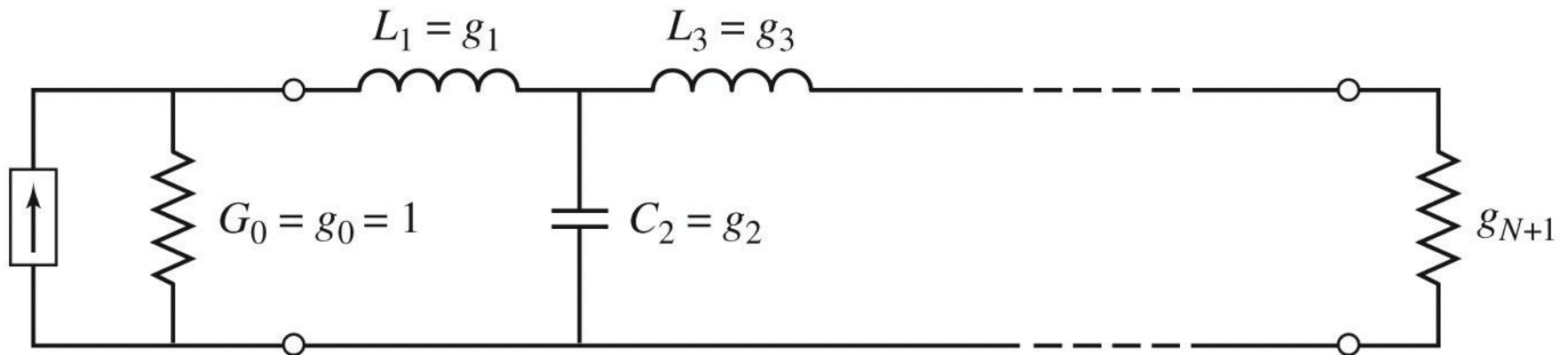
Raspunsul filtrului maxim plat



Filtre prototip



(a)



(b)

Filtru prototip maxim plat

- Calculul elementelor filtrului

$$g_0 = 1$$

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

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

Filtru prototip maxim plat

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 8.4 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.

Pentru ordin par al
filtrului ($N = 2, 4, 6, 8 \dots$)
filtrele echiriplu
trebuie sa fie inchise pe
o impedanta de
terminatie diferita de
cea standard $g_{N+1} \neq 1$
Daca aplicatia nu
suporta aceasta
comportare, e necesara
introducerea unei
adaptari de impedanta
suplimentare
(transformator in sfert
de lungime de unda,
binomial ...)

Scalare si conversie

- FTJ (prototip) are utilitate doar ca pas intermediar
 - filtru FTJ
 - frecventa de taiere $\omega_o = 1 \text{ rad/s}$ ($f_o = 0.159 \text{ Hz}$)
 - conectate la intrare la o rezistenta $R = 1$

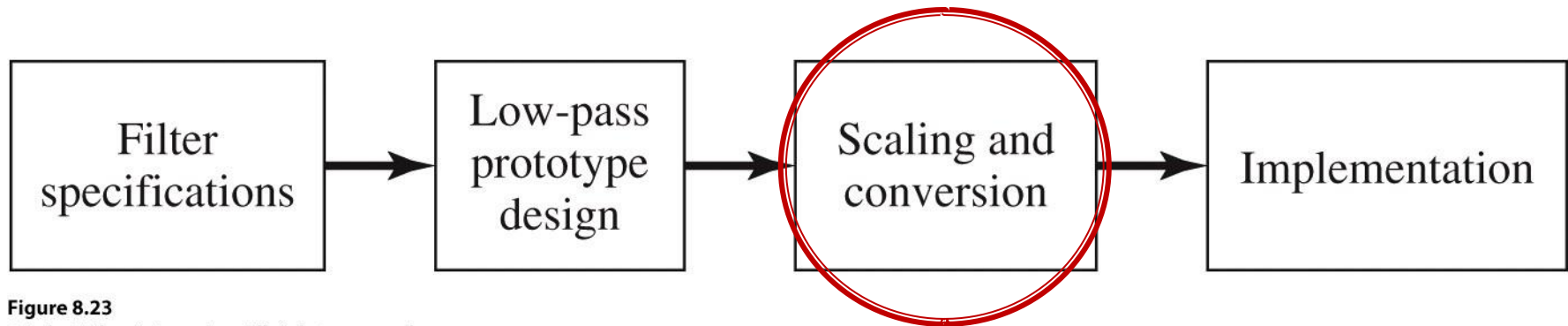


Figure 8.23

Scalarea in impedanta

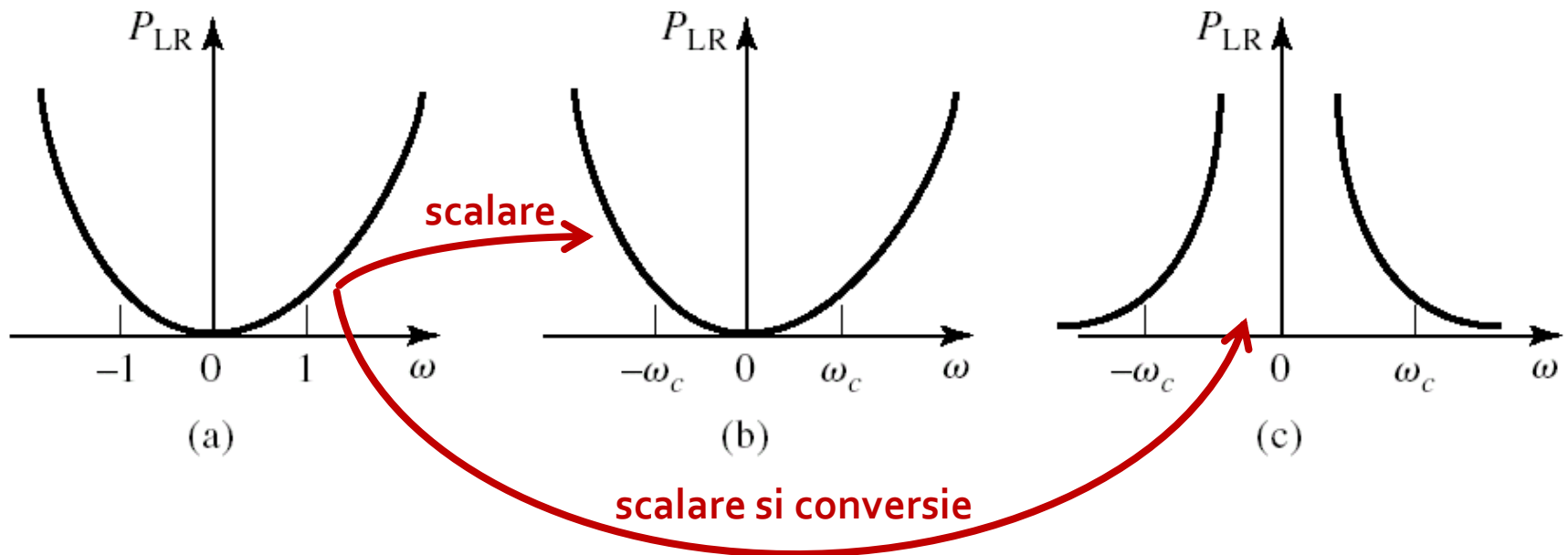
- Pentru a determina filtrul care lucreaza cu o alta rezistenta de intrare (R_0) se multiplica toate impedantele din circuit cu aceasta impedanta (cu " ' " se noteaza elementele noi)

$$R'_s = R_0 \quad R'_L = R_0 \cdot R_L$$

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

Scalarea in frecventa

- modificarea frecventei de taiere – (fig. b)
- modificarea caracteristicii de atenuare (de exemplu FTJ $\rightarrow$ FTS – fig. c)

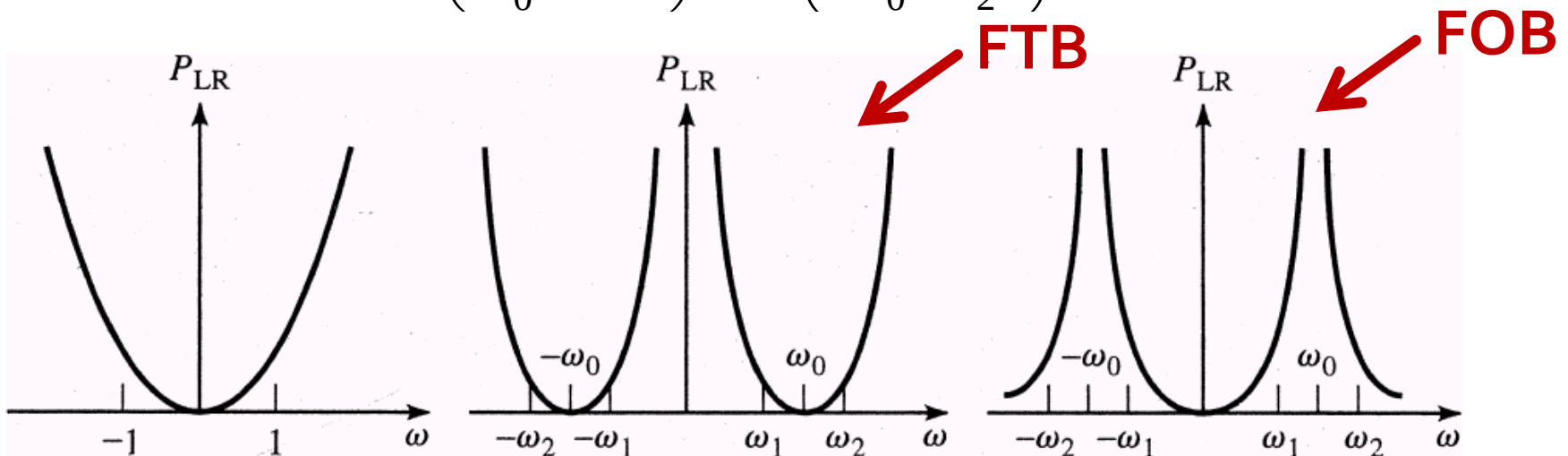


Conversie si scalare FTJ $\rightarrow$ FTB

$$\omega = \omega_0 \rightarrow \frac{1}{\Delta} \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right) = 0$$

$$\omega = \omega_1 \rightarrow \frac{1}{\Delta} \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right) = \frac{1}{\Delta} \left(\frac{\omega_1^2 - \omega_0^2}{\omega_0 \cdot \omega_1} \right) = -1$$

$$\omega = \omega_2 \rightarrow \frac{1}{\Delta} \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right) = \frac{1}{\Delta} \left(\frac{\omega_2^2 - \omega_0^2}{\omega_0 \cdot \omega_2} \right) = 1$$



Transformari ale filtrului prototip

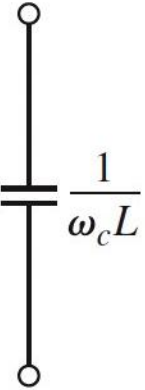
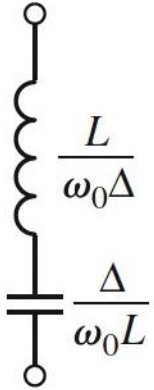
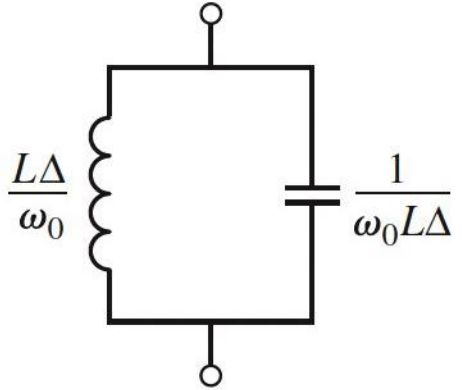
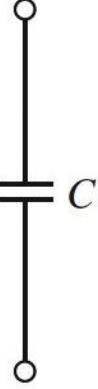
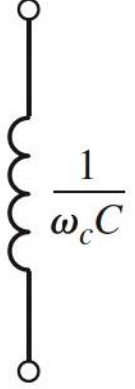
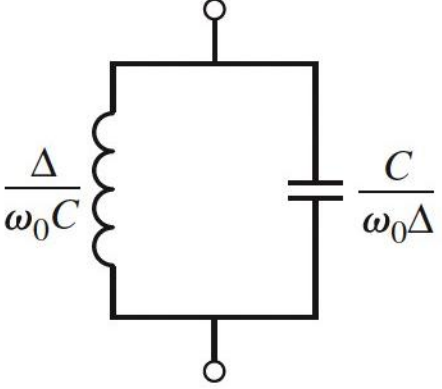
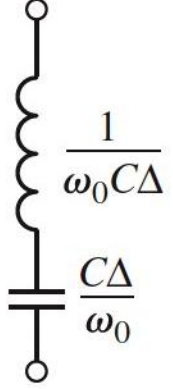
Low-pass	High-pass	Bandpass	Bandstop
			
			

Table 8.6

Exemplu

- Să se proiecteze un filtru trece-bandă de ordinul 3, avînd riplurile în bandă de 0.5 dB. Frecvența centrală a filtrului sa fie de 1 GHz. Banda să fie de 10%, și impedanța de 50 Ω .

Exemplu

- Să se proiecteze un filtru trece-bandă de ordinul 3, avînd riplurile în bandă de 0.5 dB. Frecvența centrală a filtrului sa fie de 1 GHz. Banda să fie de 10%, și impedanța de 50 Ω .

$$\omega_0 = 2 \cdot \pi \cdot 1GHz = 6.283 \cdot 10^9 \text{ rad} / s$$

$$\Delta = 0.1$$

Tabel echiriplu 0.5dB sau relatii de calcul:

$$g_1 = 1.5963 = L_1,$$

$$g_2 = 1.0967 = C_2,$$

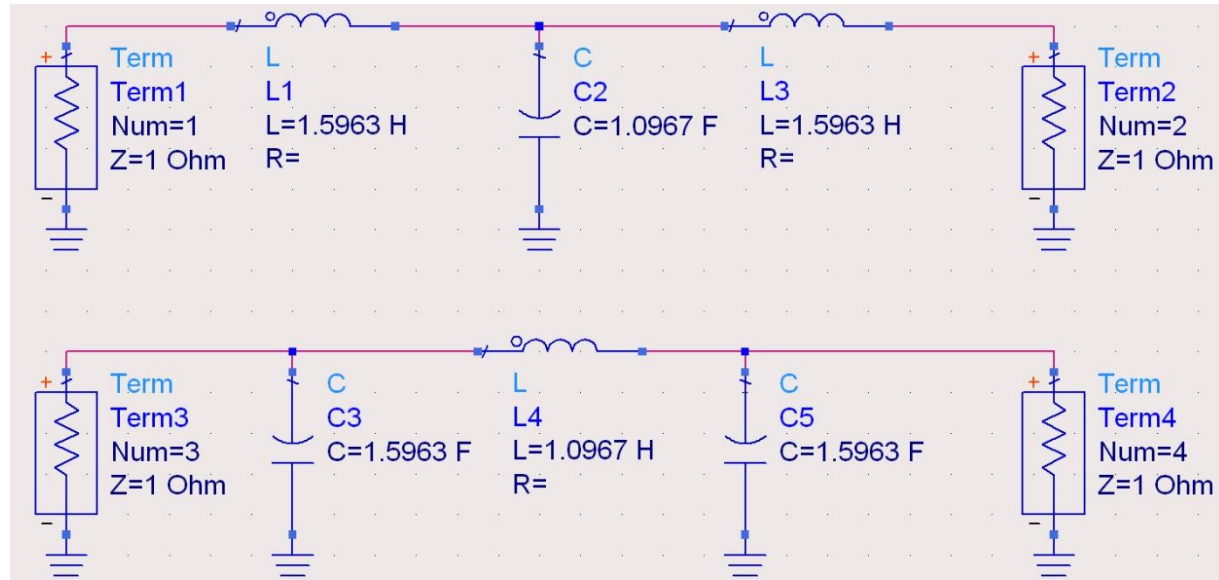
$$g_3 = 1.5963 = L_3,$$

$$g_4 = 1.000 = R_L$$

Filtru prototip

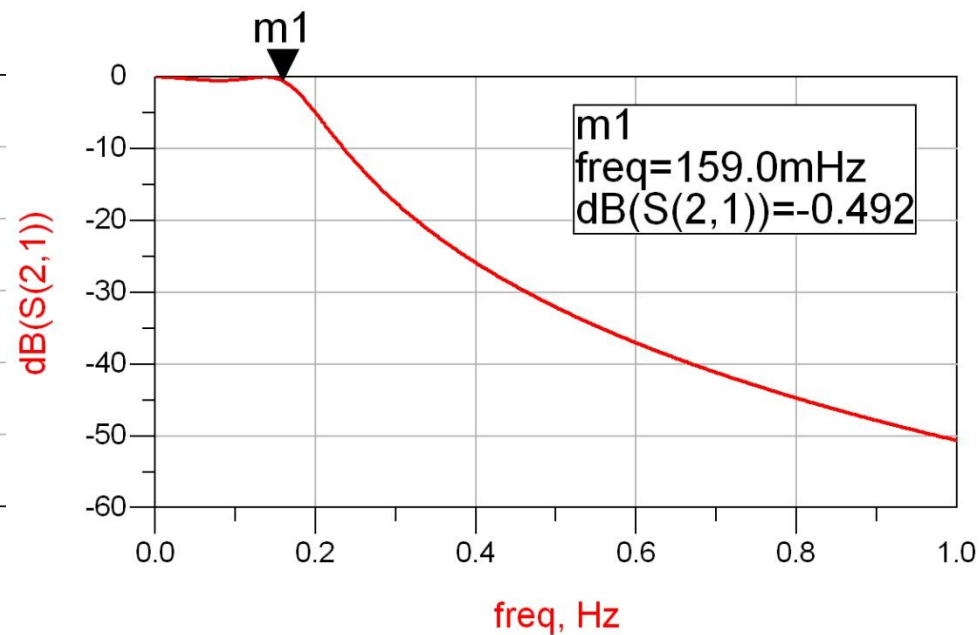
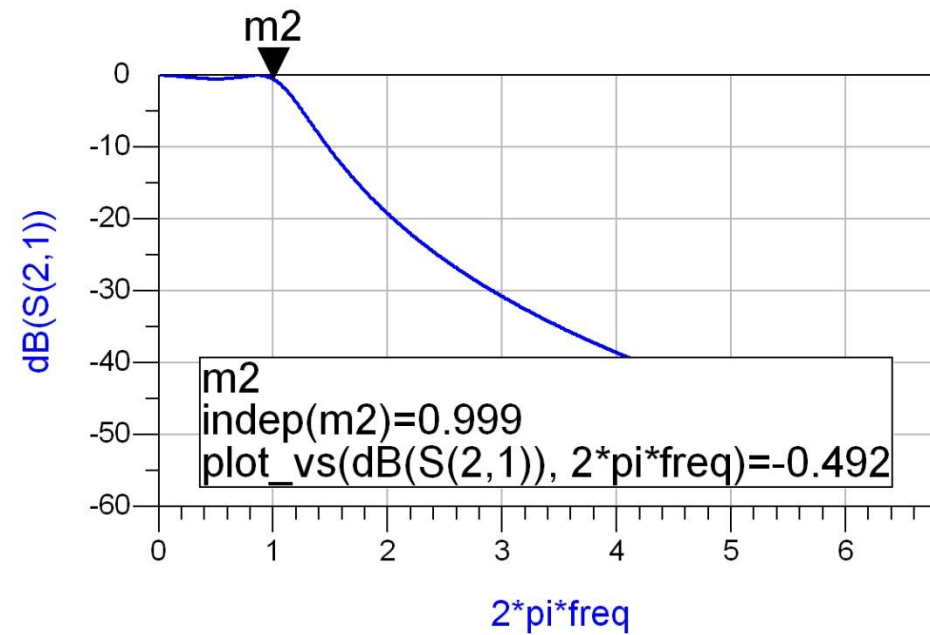
- Tabel echiriplu 0.5dB sau relatii de calcul:

- $g_1 = 1.5963 = L_1/C_3,$
- $g_2 = 1.0967 = C_2/L_4,$
- $g_3 = 1.5963 = L_3/C_5,$
- $g_4 = 1.000 = R_L$



Filtru prototip

- $\omega_o = 1 \text{ rad/s}$ ($f_o = \omega_o / 2\pi = 0.159 \text{ Hz}$)



Scalare frecventa / FTB

$$\omega_0 = 2 \cdot \pi \cdot 1 \text{GHz} = 6.283 \cdot 10^9 \text{rad/s} \quad \Delta = \frac{\Delta \omega}{\omega_0} = \frac{\Delta f}{f_0} = 0.1 \quad R_0 = 50 \Omega$$

$$g_1 = 1.5963 = L_1,$$

$$g_2 = 1.0967 = C_2,$$

$$g_3 = 1.5963 = L_3,$$

$$g_4 = 1.000 = R_L$$

$$L'_1 = \frac{L_1 \cdot R_0}{\Delta \cdot \omega_0} = 127.0 \text{nH}$$

$$C'_1 = \frac{\Delta}{\omega_0 \cdot L_1 \cdot R_0} = 0.199 \text{pF}$$

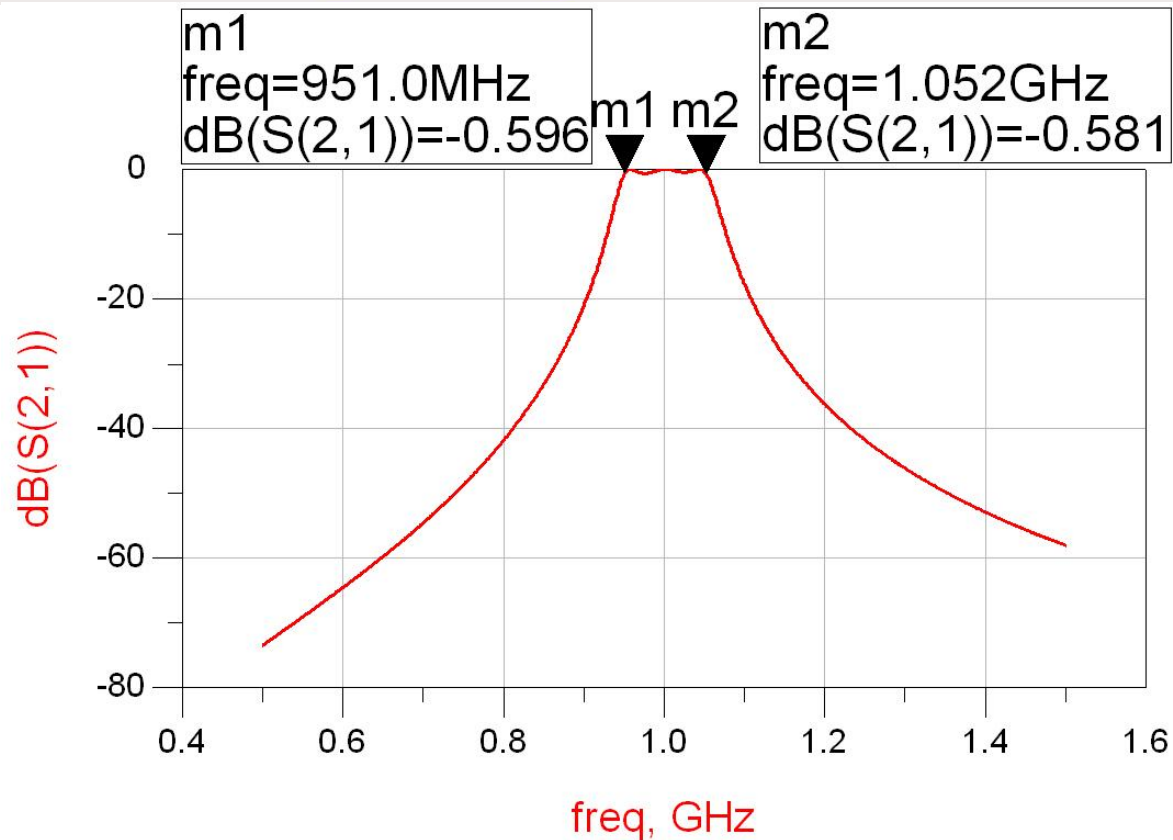
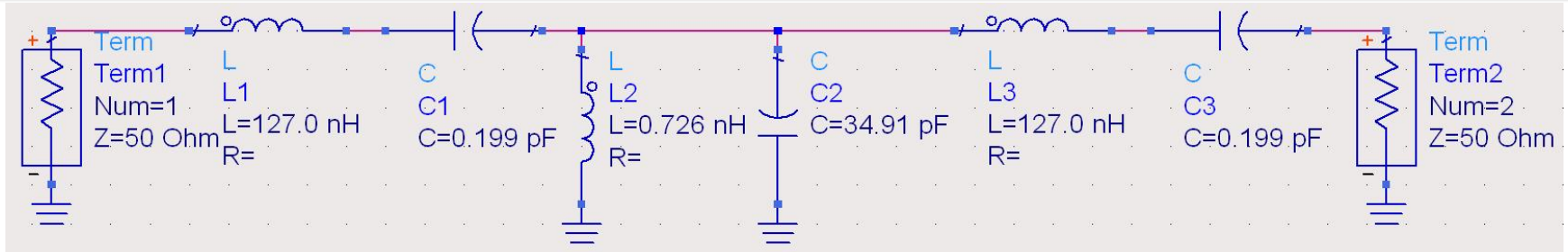
$$L'_2 = \frac{\Delta \cdot R_0}{\omega_0 \cdot C_2} = 0.726 \text{nH}$$

$$C'_2 = \frac{C_2}{\Delta \cdot \omega_0 \cdot R_0} = 34.91 \text{pF}$$

$$L'_3 = \frac{L_3 \cdot R_0}{\Delta \cdot \omega_0} = 127.0 \text{nH}$$

$$C'_3 = \frac{\Delta}{\omega_0 \cdot L_3 \cdot R_0} = 0.199 \text{pF}$$

ADS



Implementarea filtrelor pentru microunde

Implementare filtre pentru microunde

- Implementarea cu elemente concentrate (L, C) este utilizabila mai ales in zona frecventelor ceva mai reduse (RF) datorita:
 - dificultatii de implementare a valorilor rezultante pentru componente
 - dificultatii de a asigura toleranta (foarte mica) necesara pentru componente

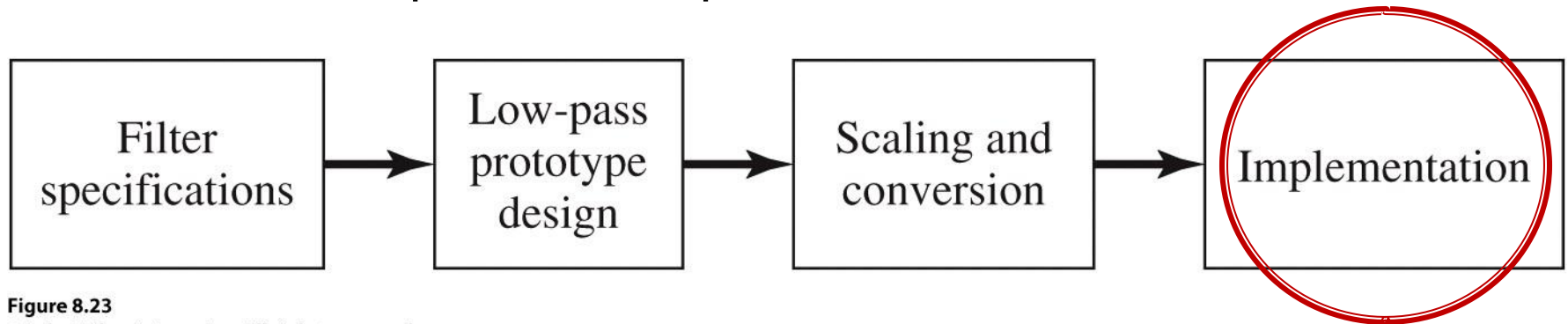
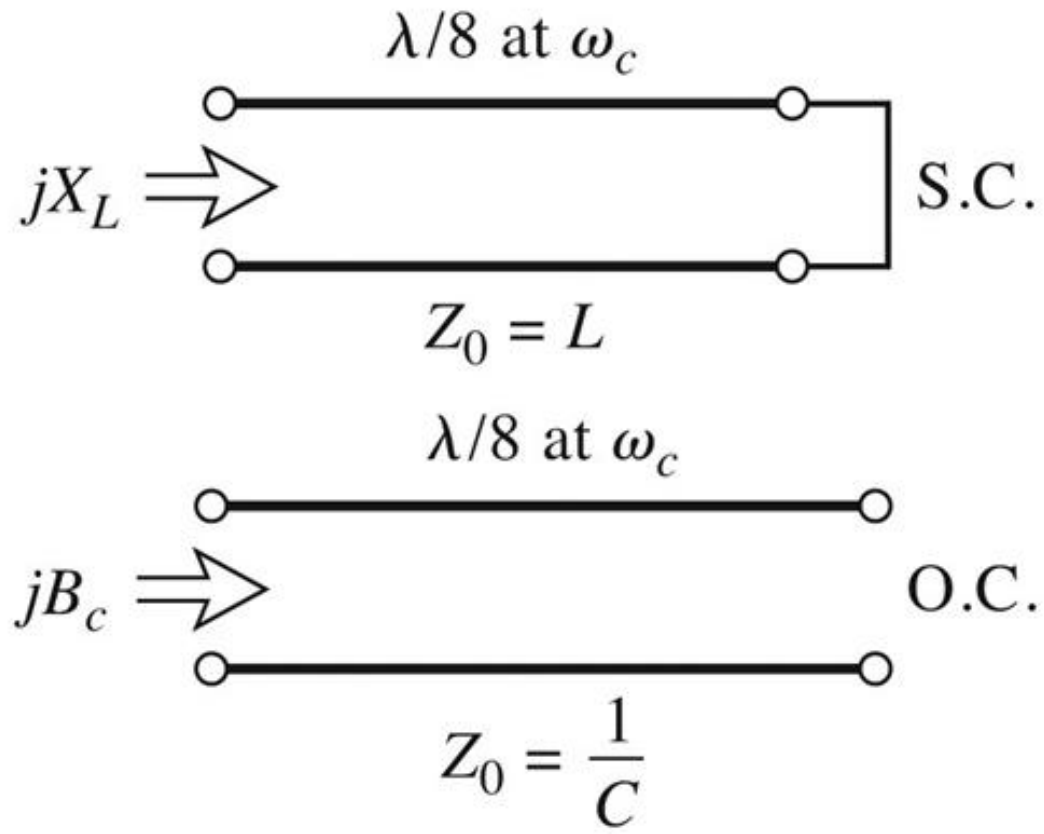
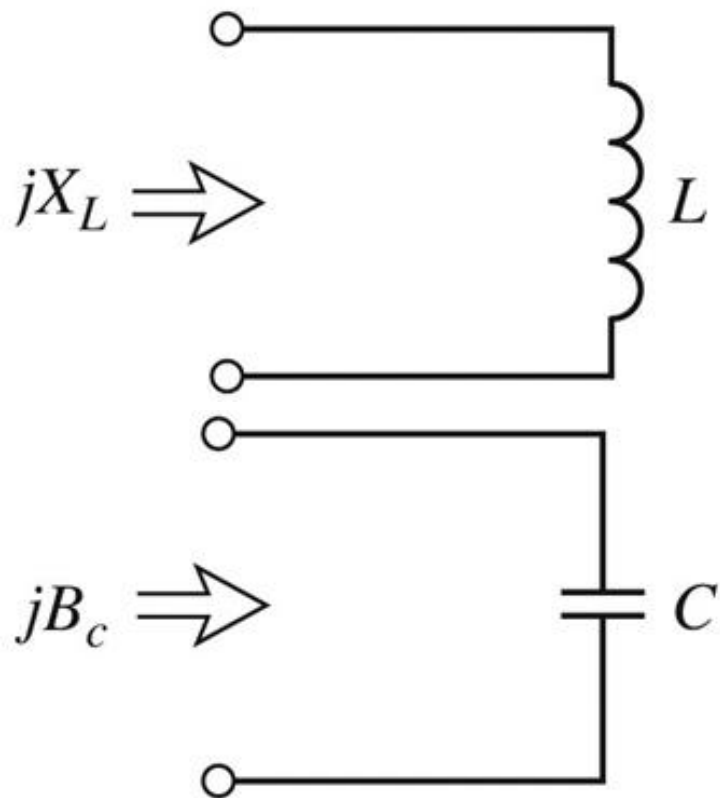


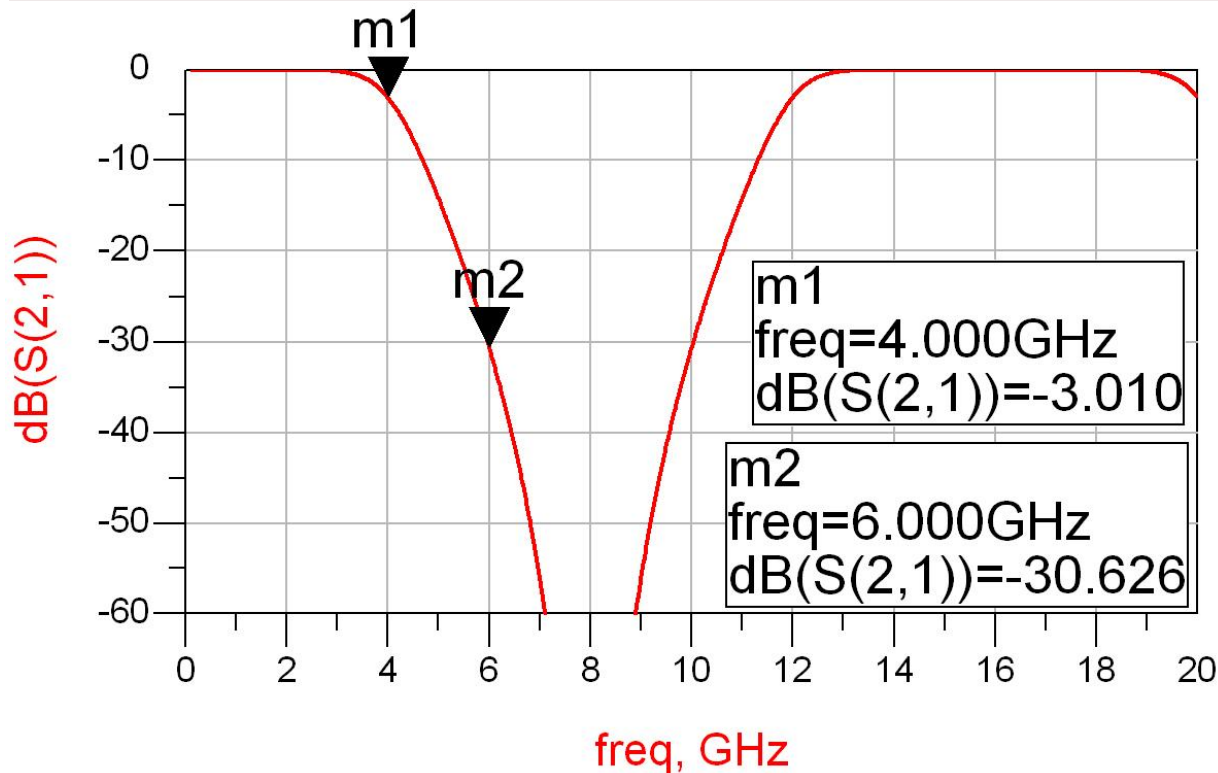
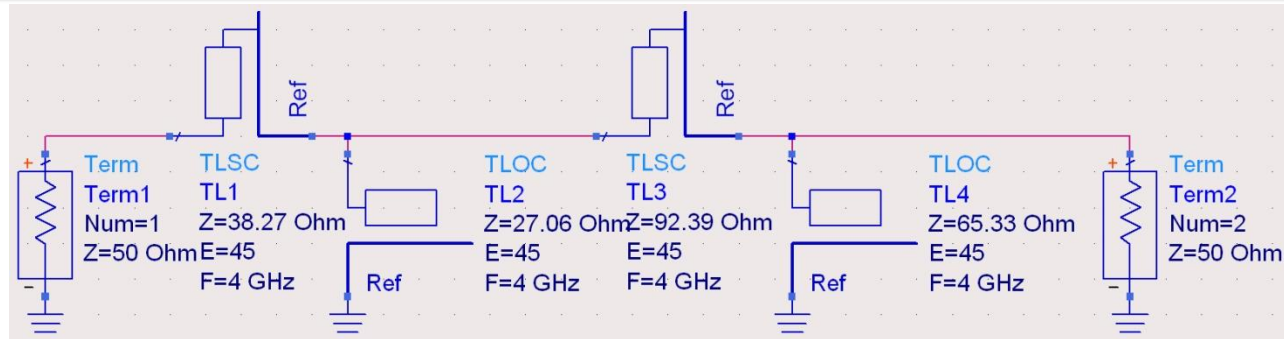
Figure 8.23

Transformarea Richards

- permite obtinerea cu sectiuni de linii a inductantelor si capacitatilor **dupa** scalarea prototipului pentru functia corespunzatoare (FTJ/FTS/FTB /FOB)

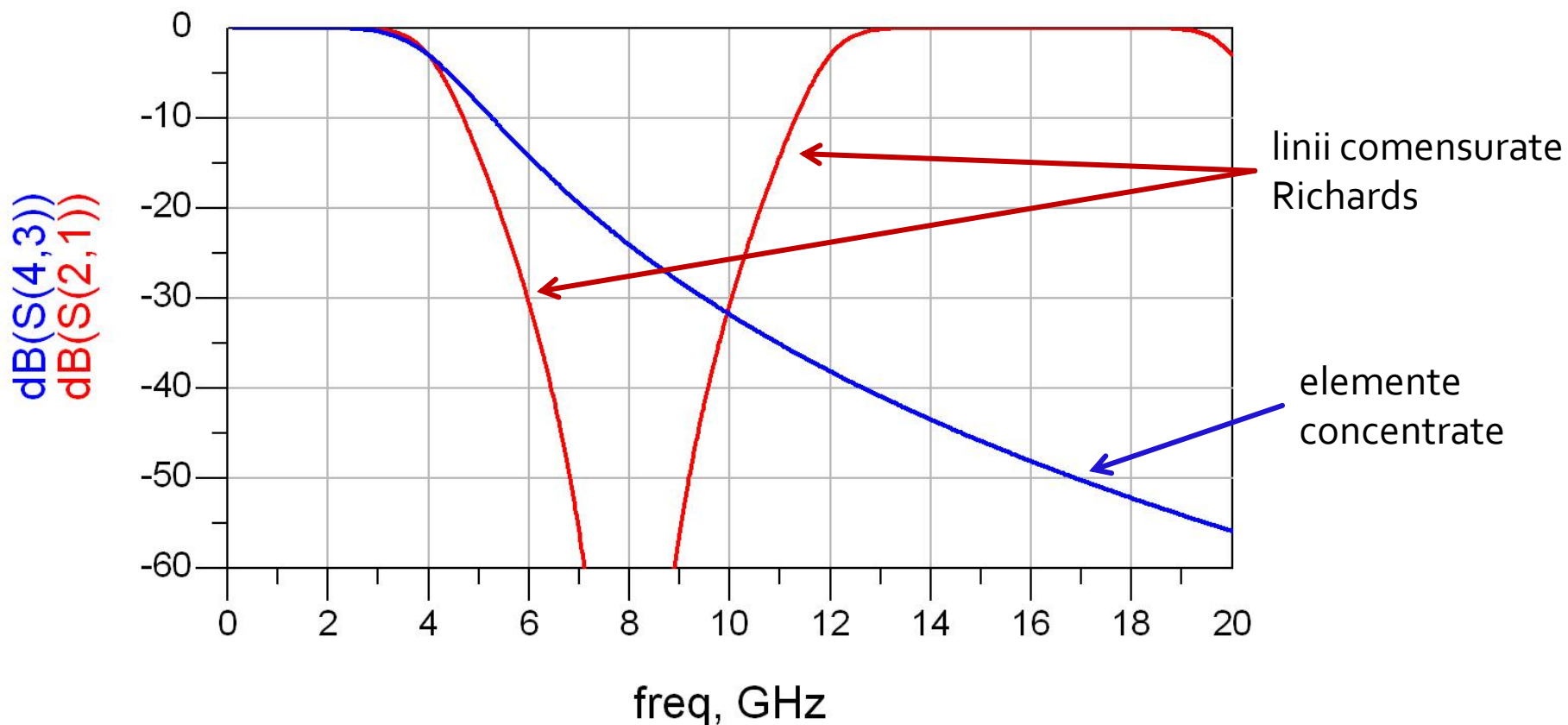


Transformarea Richards – ADS



Transformarea Richards

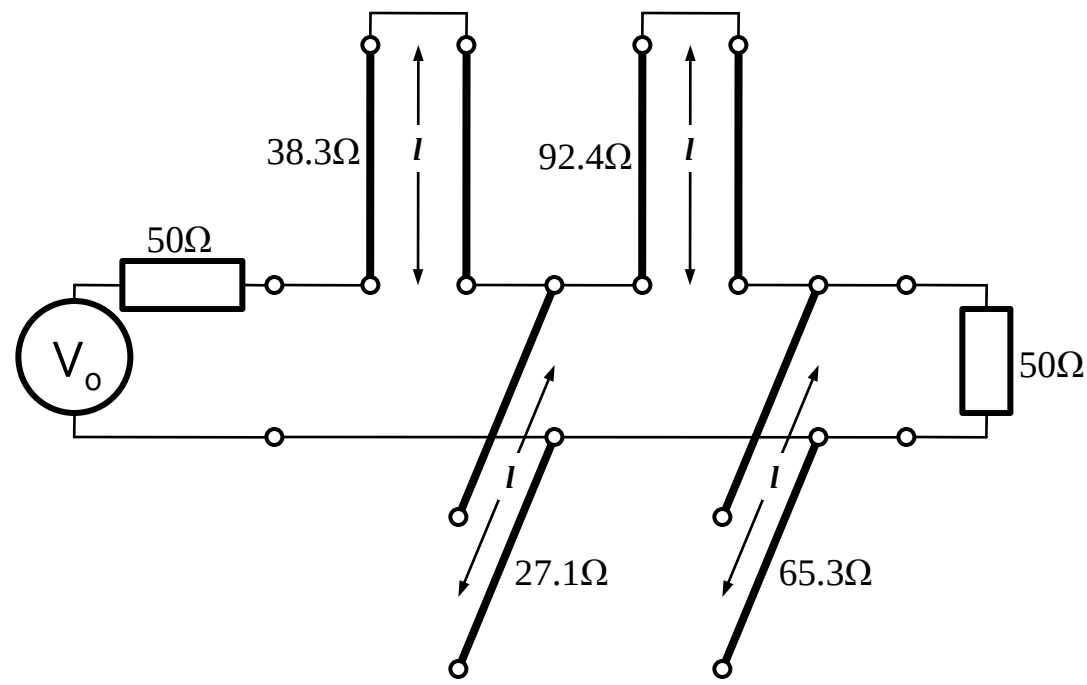
- Filtre realizate cu transformarea Richards
 - beneficiaza de polul suplimentar de atenuare $2 \cdot \omega_c$
 - au dezavantajul periodicitatii in frecventa, de obicei se prevede un filtru trece jos suplimentar neperiodic daca e necesar



Identitatile Kuroda

- Filtre implementate cu transformarea Richards au anumite dezavantaje in ceea ce priveste implementarea practica
- Identitatile/Transformarile Kuroda pot fi utilizate pentru a elimina o parte din aceste dezavantaje

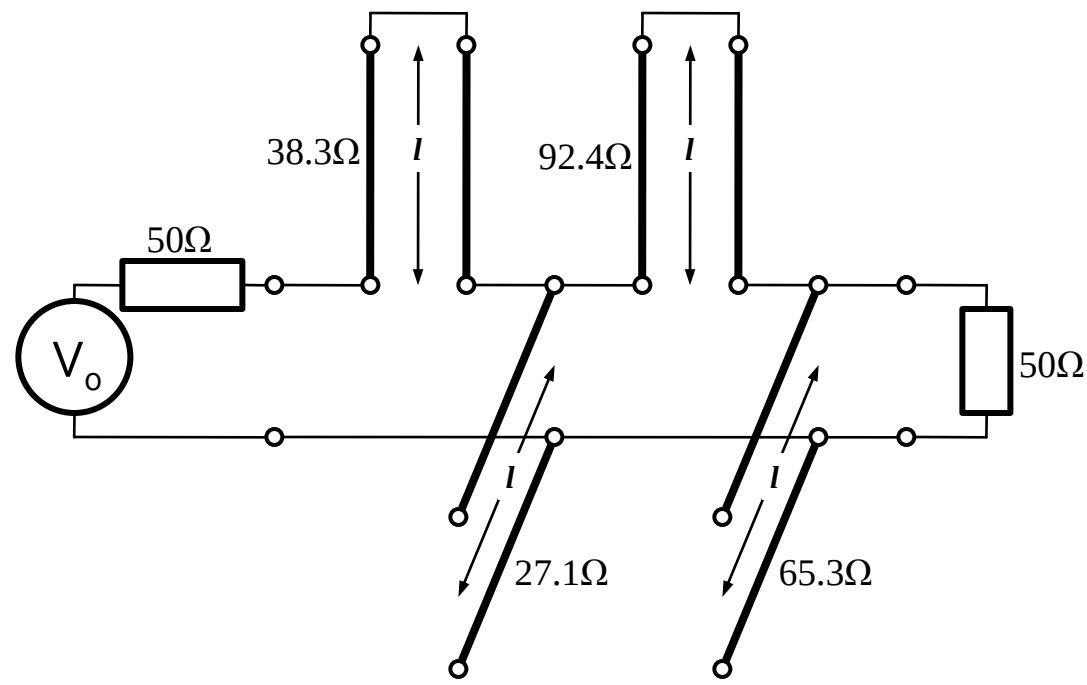
- Se utilizeaza sectiuni de linie suplimentare pentru a obtine sisteme mai simplu de implementat in practica
- Liniile suplimentare se numesc elemente unitare si au lungimi de $\lambda/8$ la frecventa de taiere dorita (ωc) fiind comensurate cu celelalte sectiuni de linie



Identitatile Kuroda

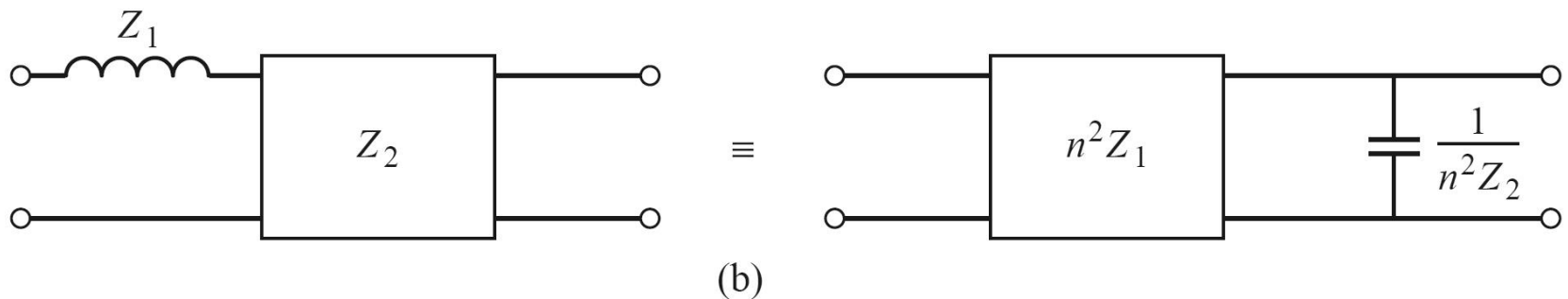
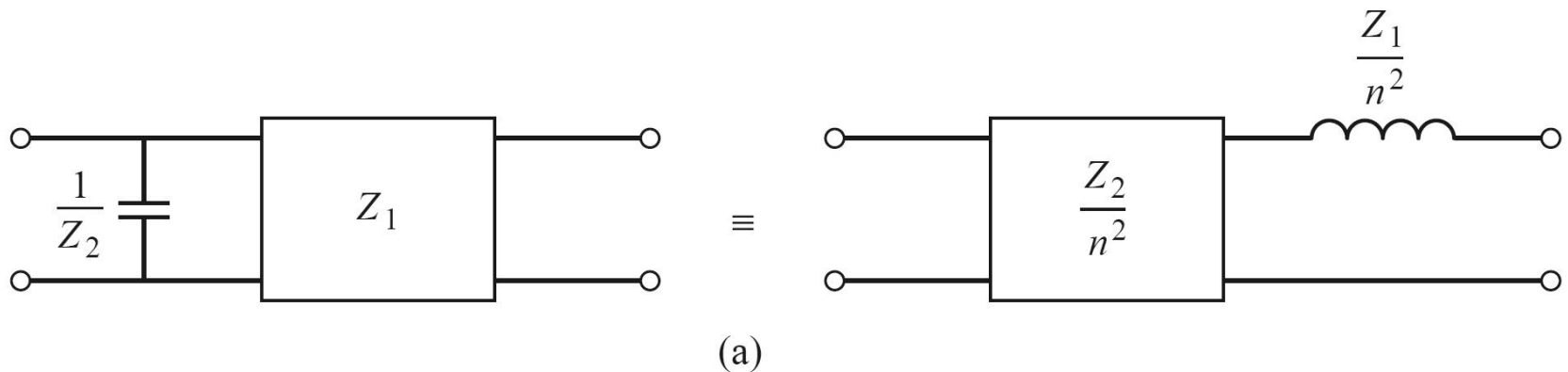
- Identitatile Kuroda pot fi utilizate pentru a realiza urmatoarele operatii:

- Separarea fizica a diferitelor stub-uri
- Transformarea stub-urilor serie in stub-uri paralel sau invers
- Obtinerea unor impedante caracteristice mai "realizabile" pentru linii ($\sim 50\Omega$)



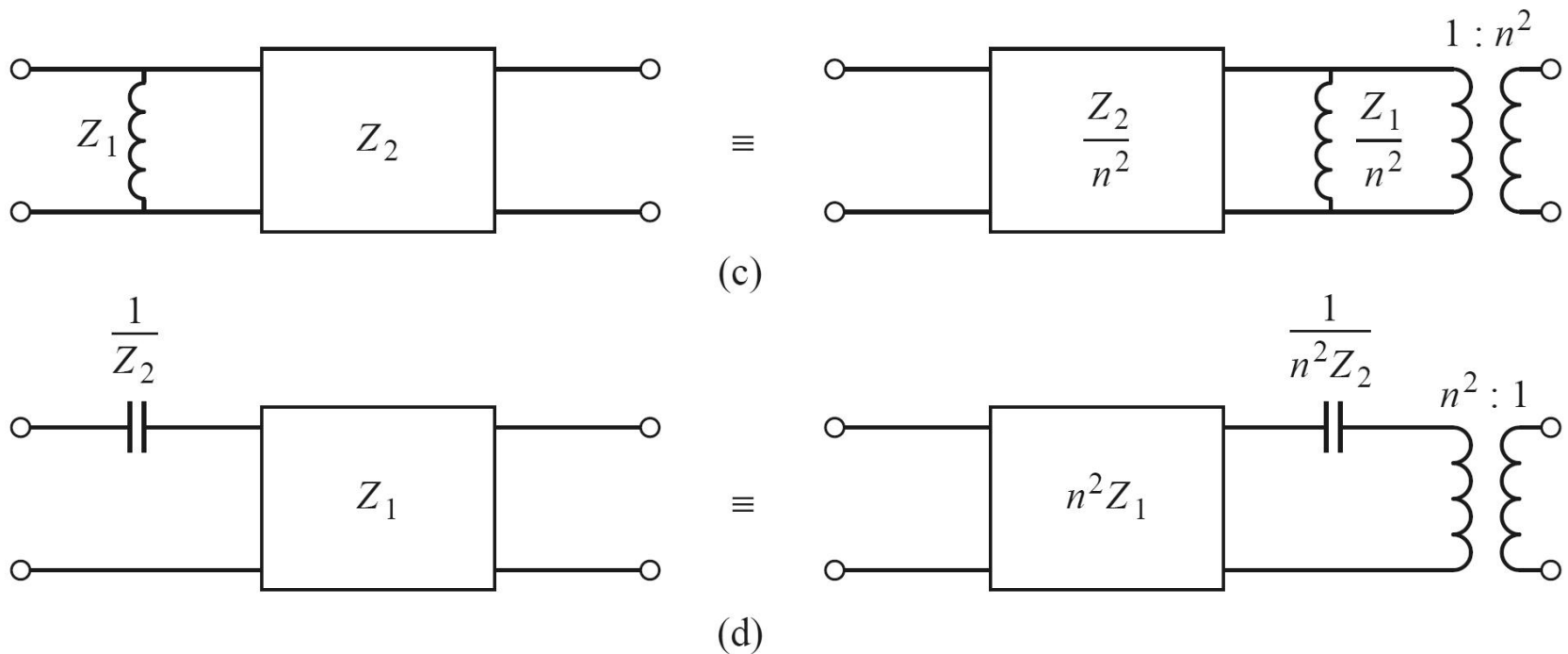
Identitățile Kuroda

- 4 echivalente de circuit



Identitatile Kuroda

- 4 echivalente de circuit



Identitatile Kuroda

- In toate echivalentele de scheme Kuroda:
 - n:
$$n^2 = 1 + \frac{Z_2}{Z_1}$$
 - inductantele si capacitatile reprezinta stub-uri scurtcircuitate sau in gol (obtinute prin transformarea Richards, de lungime $\lambda/8$)
 - blocurile reprezinta elemente unitare (linii de transmisie de impedanta caracteristica indicata si lungime $\lambda/8$)

Prima Identitate Kuroda

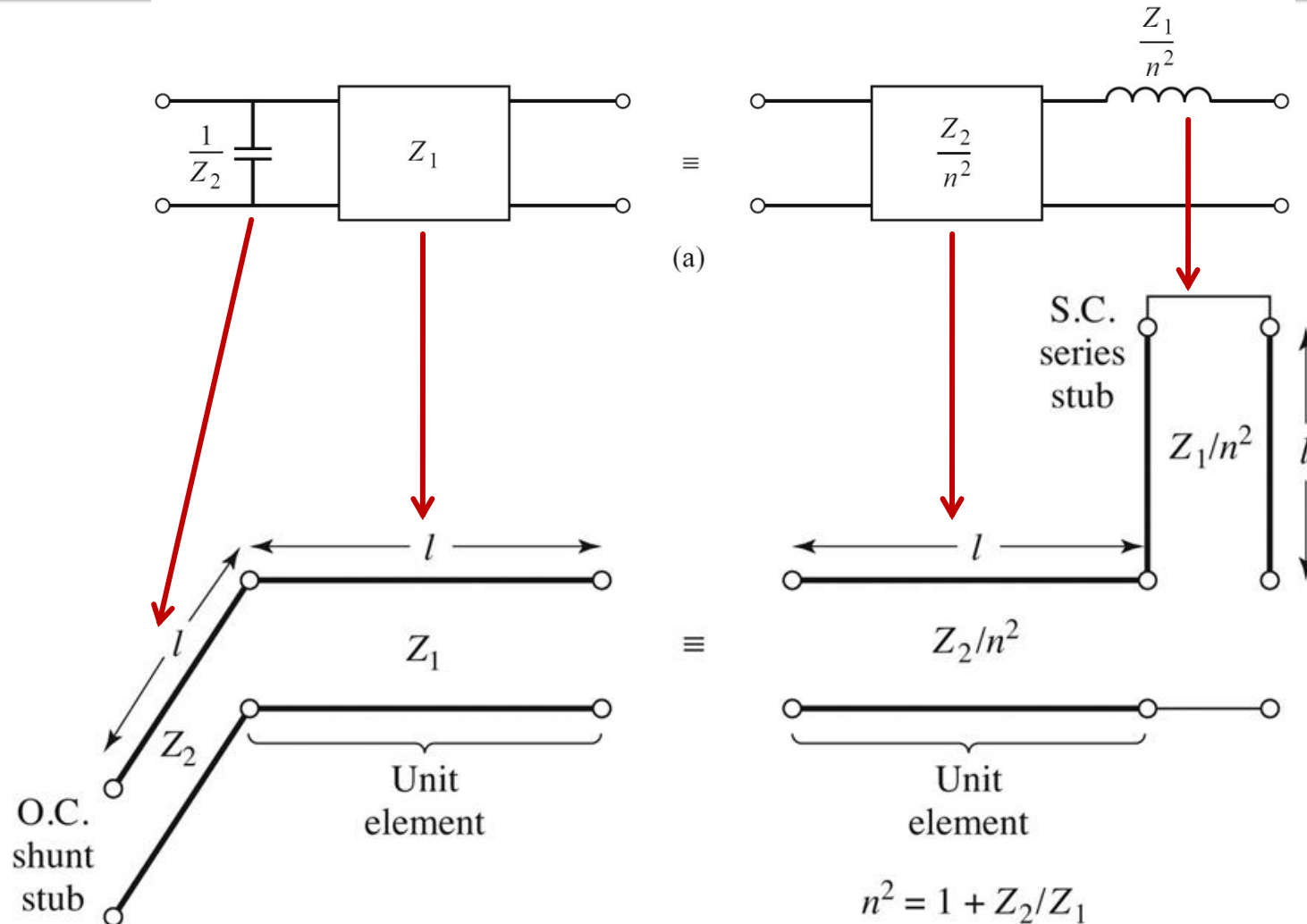


Figure 8.35
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Prima Identitate Kuroda – Demonstratie

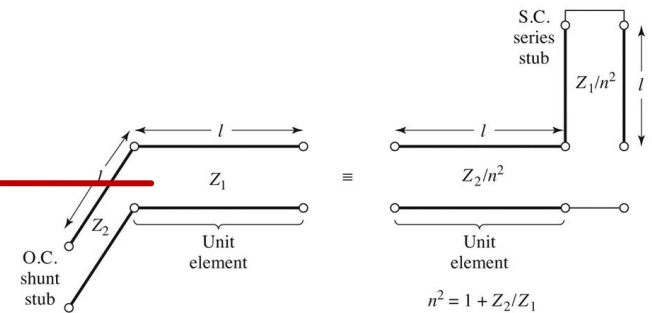
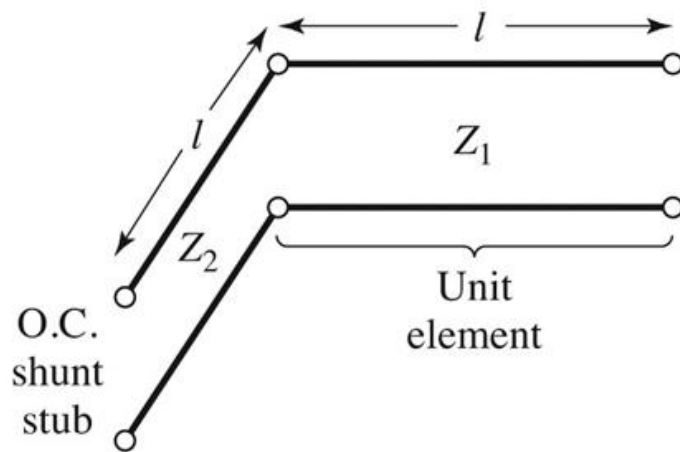
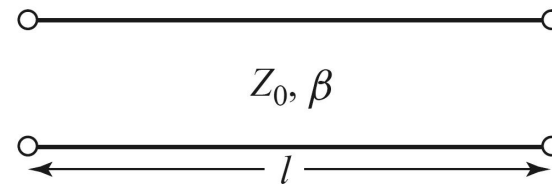
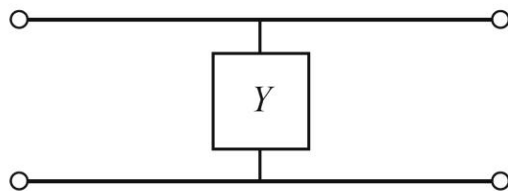


Figure 8.35
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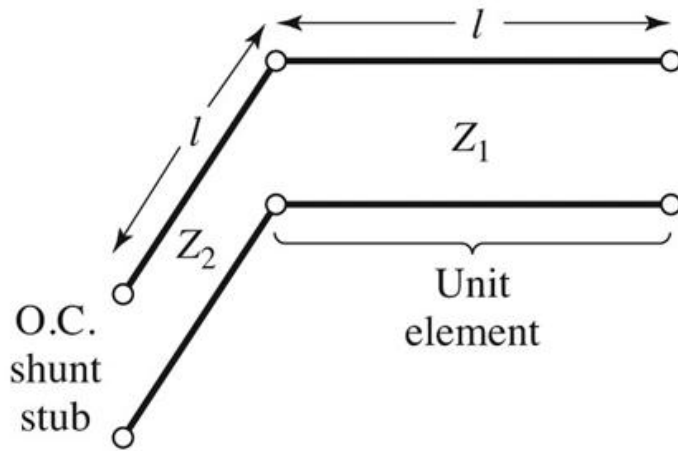
■ Matrici ABCD, C₄



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

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

Prima Identitate Kuroda – Demonstratie



$$\Omega = \tan \beta \cdot l$$

$$\cos \beta \cdot l = \frac{1}{\sqrt{1 + \Omega^2}} \quad \sin \beta \cdot l = \frac{\Omega}{\sqrt{1 + \Omega^2}}$$

$$Z_{in,g} = -j \cdot Z_2 \cdot \cot \beta \cdot l = -j \cdot \frac{Z_2}{\Omega}$$

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ \frac{j \cdot \Omega}{Z_2} & 1 \end{bmatrix} \cdot \begin{bmatrix} \frac{1}{\sqrt{1 + \Omega^2}} & j \cdot Z_1 \cdot \frac{\Omega}{\sqrt{1 + \Omega^2}} \\ j \cdot \frac{1}{Z_1} \cdot \frac{\Omega}{\sqrt{1 + \Omega^2}} & \frac{1}{\sqrt{1 + \Omega^2}} \end{bmatrix}$$

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \frac{1}{\sqrt{1 + \Omega^2}} \cdot \begin{bmatrix} 1 & 0 \\ \frac{j \cdot \Omega}{Z_2} & 1 \end{bmatrix} \cdot \begin{bmatrix} 1 & j \cdot \Omega \cdot Z_1 \\ \frac{j \cdot \Omega}{Z_1} & 1 \end{bmatrix} = \frac{1}{\sqrt{1 + \Omega^2}} \cdot \begin{bmatrix} 1 & j \cdot \Omega \cdot Z_1 \\ j \cdot \Omega \cdot \left(\frac{1}{Z_1} + \frac{1}{Z_2} \right) & 1 - \Omega^2 \cdot \frac{Z_1}{Z_2} \end{bmatrix}$$

Prima Identitate Kuroda – Demonstratie

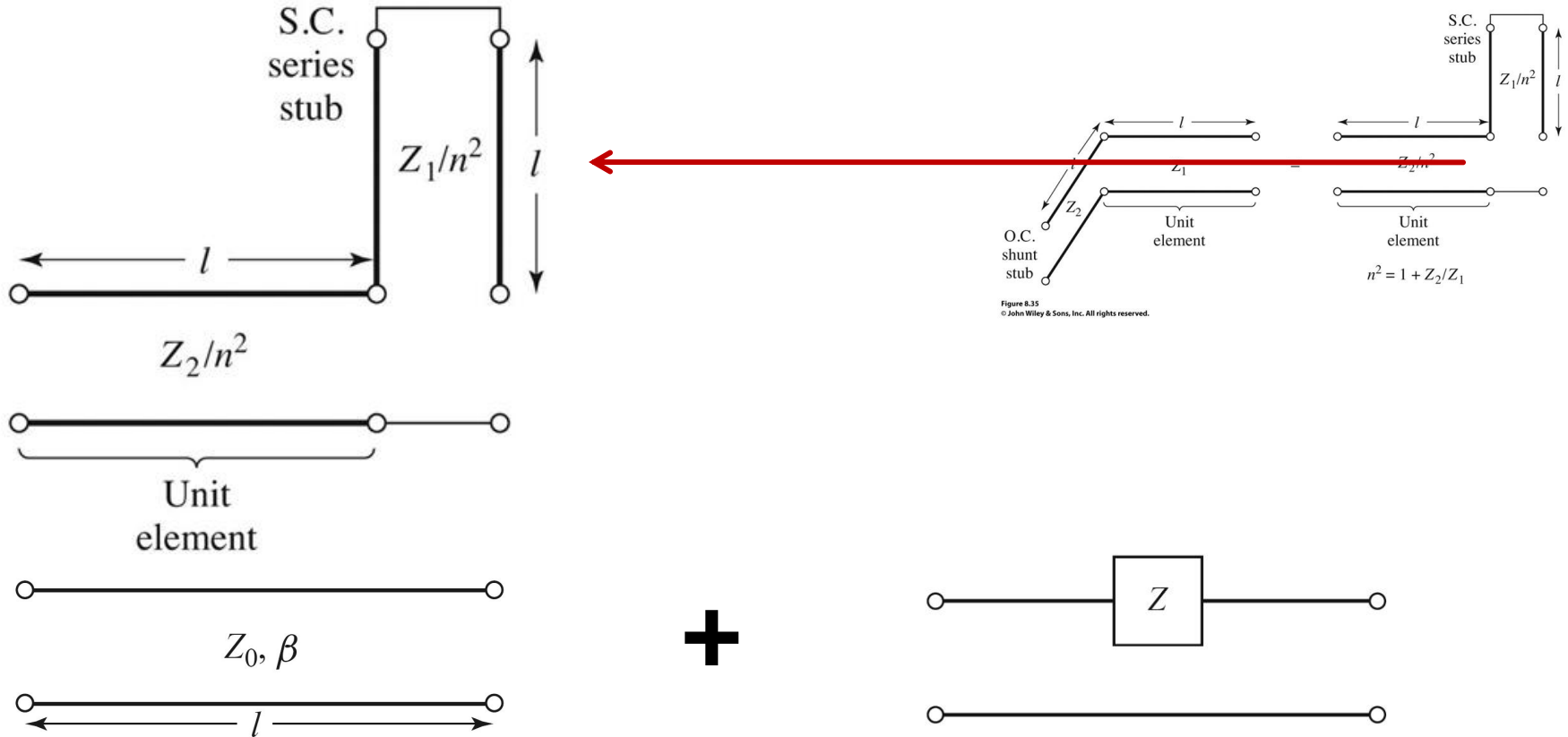
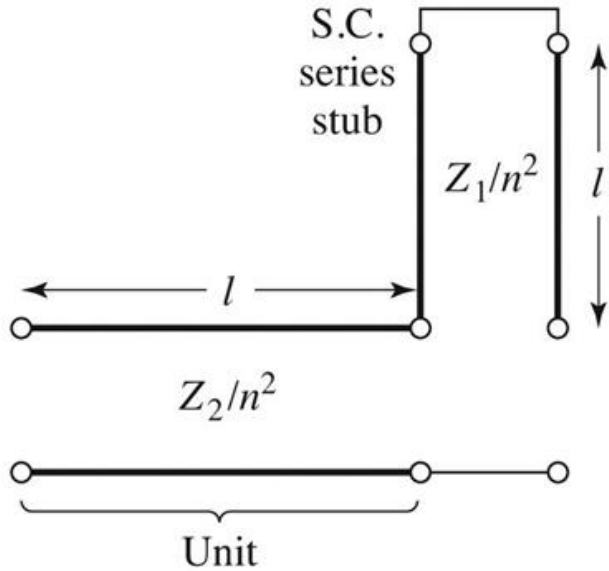


Figure 8.35
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$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \cos \beta \cdot l & j \cdot Z_0 \cdot \sin \beta \cdot l \\ j \cdot Y_0 \cdot \sin \beta \cdot l & \cos \beta \cdot l \end{bmatrix}$$

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

Prima Identitate Kuroda – Demonstratie



$$\Omega = \tan \beta \cdot l$$

$$\cos \beta \cdot l = \frac{1}{\sqrt{1 + \Omega^2}} \quad \sin \beta \cdot l = \frac{\Omega}{\sqrt{1 + \Omega^2}}$$

$$Z_{in,sc} = j \cdot \left(\frac{Z_1}{n^2} \right) \cdot \tan \beta \cdot l = \frac{j \cdot \Omega \cdot Z_1}{n^2}$$

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} 1 & j \cdot \frac{Z_2}{n^2} \cdot \frac{\Omega}{\sqrt{1 + \Omega^2}} \\ j \cdot \frac{n^2}{Z_2} \cdot \frac{\Omega}{\sqrt{1 + \Omega^2}} & 1 \end{bmatrix} \cdot \begin{bmatrix} 1 & \frac{j \cdot \Omega \cdot Z_1}{n^2} \\ 0 & 1 \end{bmatrix}$$

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \frac{1}{\sqrt{1 + \Omega^2}} \cdot \begin{bmatrix} 1 & j \cdot \Omega \cdot \frac{Z_2}{n^2} \\ \frac{j \cdot \Omega \cdot n^2}{Z_2} & 1 \end{bmatrix} \cdot \begin{bmatrix} 1 & j \cdot \Omega \cdot \frac{Z_1}{n^2} \\ 0 & 1 \end{bmatrix} = \frac{1}{\sqrt{1 + \Omega^2}} \cdot \begin{bmatrix} 1 & j \cdot \frac{\Omega}{n^2} \cdot (Z_1 + Z_2) \\ \frac{j \cdot \Omega \cdot n^2}{Z_2} & 1 - \Omega^2 \cdot \frac{Z_1}{Z_2} \end{bmatrix}$$

Prima Identitate Kuroda – Demonstratie

- Prima schema

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \frac{1}{\sqrt{1+\Omega^2}} \cdot \begin{bmatrix} 1 & j \cdot \Omega \cdot Z_1 \\ j \cdot \Omega \cdot \left(\frac{1}{Z_1} + \frac{1}{Z_2} \right) & 1 - \Omega^2 \cdot \frac{Z_1}{Z_2} \end{bmatrix}$$

- A doua schema

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \frac{1}{\sqrt{1+\Omega^2}} \cdot \begin{bmatrix} 1 & j \cdot \frac{\Omega}{n^2} \cdot (Z_1 + Z_2) \\ \frac{j \cdot \Omega \cdot n^2}{Z_2} & 1 - \Omega^2 \cdot \frac{Z_1}{Z_2} \end{bmatrix}$$

- Rezultatele sunt identice daca alegem

$$n^2 = 1 + \frac{Z_2}{Z_1}$$

- Similar se pot demonstra si celelalte trei identitati

(Acelasi) Exemplu

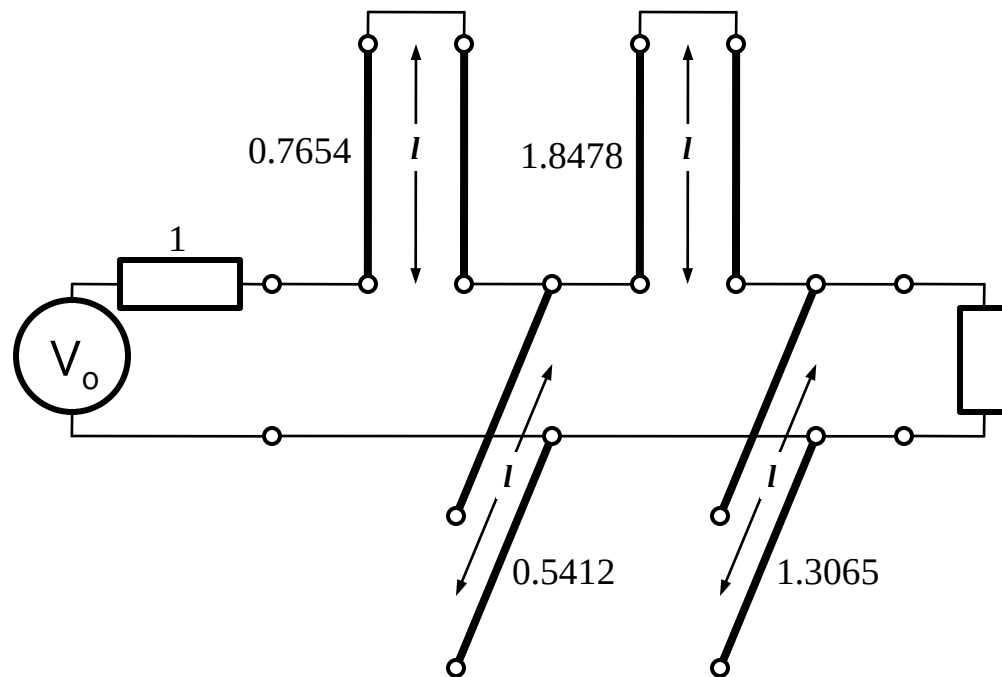
- Filtru trece jos de ordinul 4, cu frecventa de taiere de 4 GHz, de tip maxim plat (care sa functioneze pe 50Ω la intrare si iesire)
- Tabel maxim plat sau relatii de calcul:
 - $g_1 = 0.7654 = L_1$
 - $g_2 = 1.8478 = C_2$
 - $g_3 = 1.8478 = L_3$
 - $g_4 = 0.7654 = C_4$
 - $g_5 = 1$ (**nu** are nevoie de adaptare suplimentara la iesire – apare la filtrele de ordin par echiriplu)

Exemplu

■ Se aplica transformarea Richards

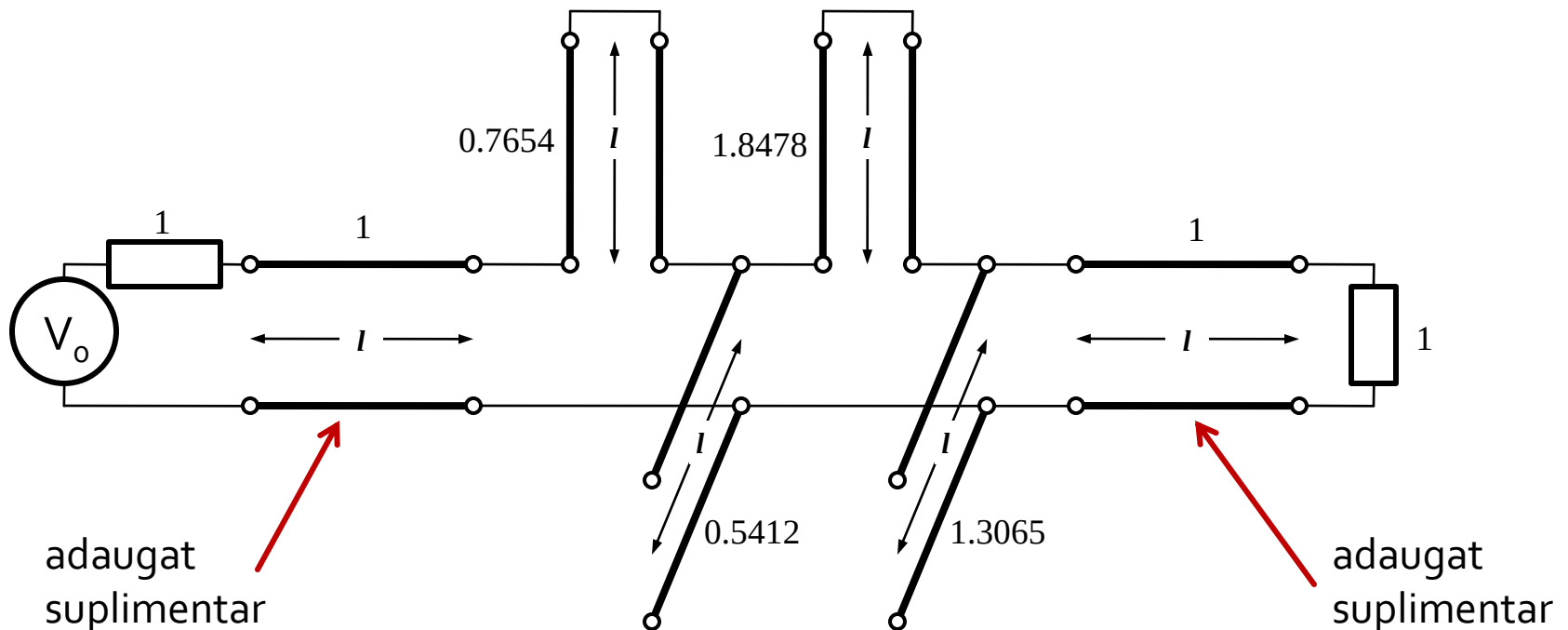
■ Probleme:

- stub-urile in serie sunt **extrem de dificil** de implementat in tehnologie microstrip
- cu tehnologia microstrip e preferabil sa avem stub-uri in gol (scurtcircuit necesita un **via-hole** spre planul de masa)
- cele 4 stub-uri sunt conectate in acelasi punct, o implementare care sa elimine/micsoreze **cuplajul** intre aceste linii e imposibila
- nu e cazul aici, dar pot aparea situatii cand impedantele raportate sunt mult diferite de 1. Majoritatea tehnologiilor sunt concepute pentru linii cu impedante caracteristice in jur de 50Ω



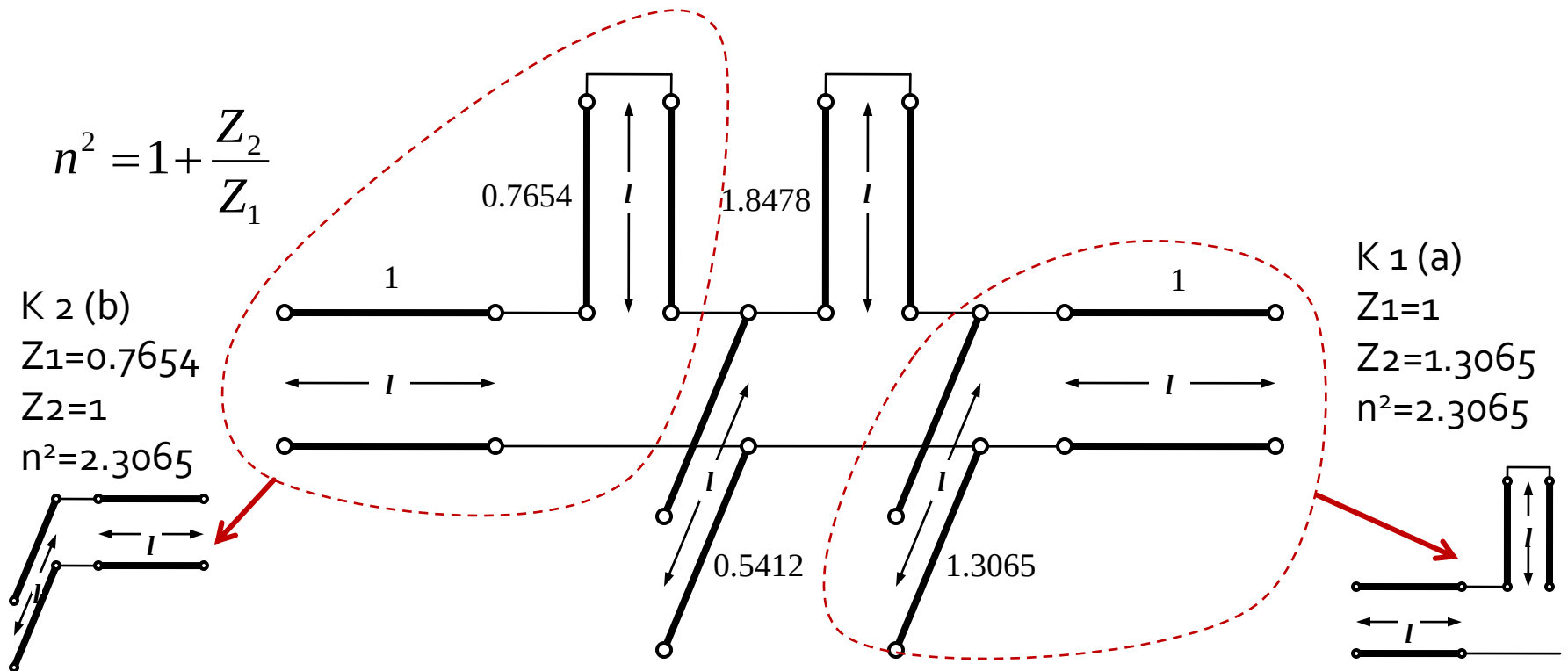
Exemplu

- Identitățile Kuroda se referă **intotdeauna** la o schema cu o secțiune de linie în serie:
 - se **adauga** elementele unitare ($z = 1, l = \lambda/8$) la fiecare capăt al circuitului (adaugarea nu modifică proprietățile filtrului acesta fiind adaptat la $z = 1$ la fiecare capăt)
 - se **aplică** una din identitățile Kuroda la fiecare capăt și se **continuă**
 - un indicator al **opririi** procedurii este apariția unei secțiuni de linie între toate stub-urile obținute cu transformarea Richards



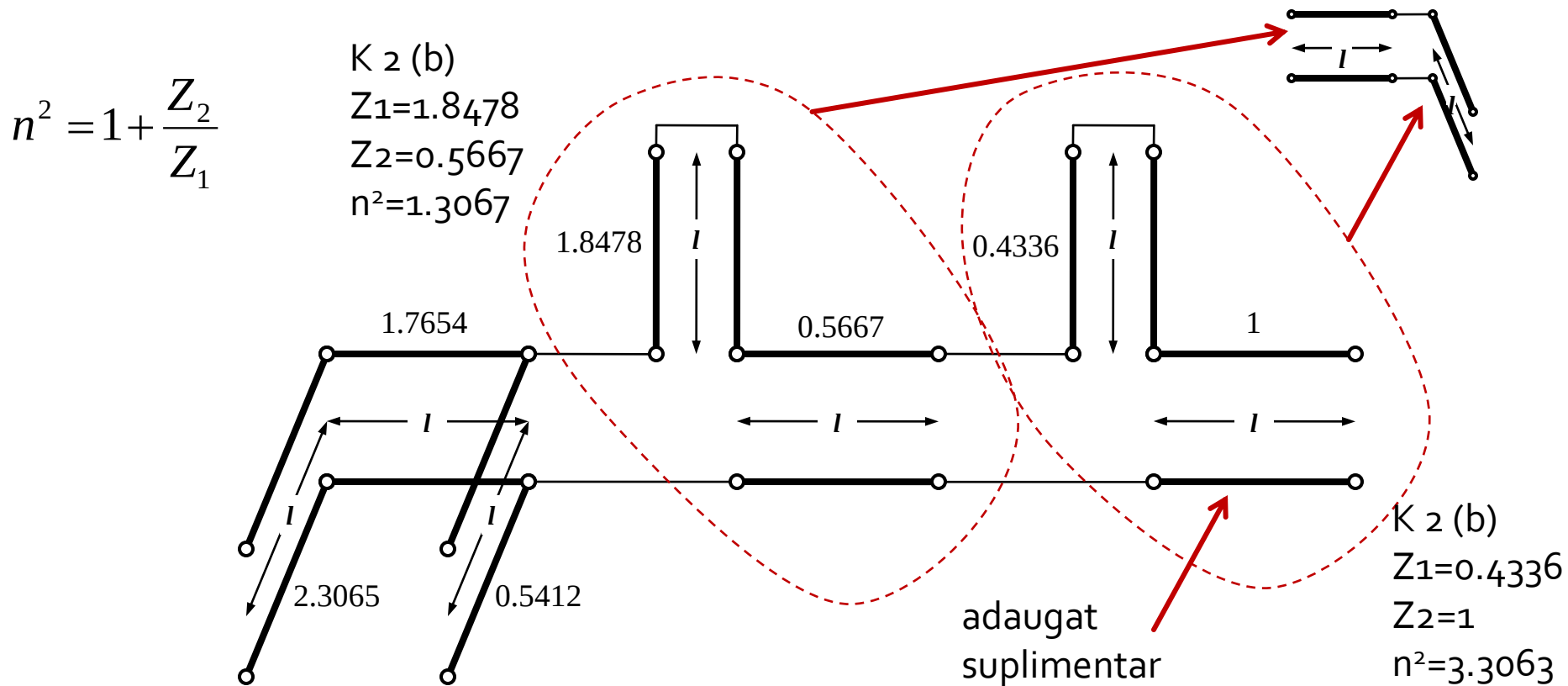
Exemplu

- Se aplica :
 - Kuroda 2 (L, Z cunoscut $\rightarrow C, Z$) in partea stanga
 - Kuroda 1 (C, Z cunoscut $\rightarrow L, Z$) in partea dreapta

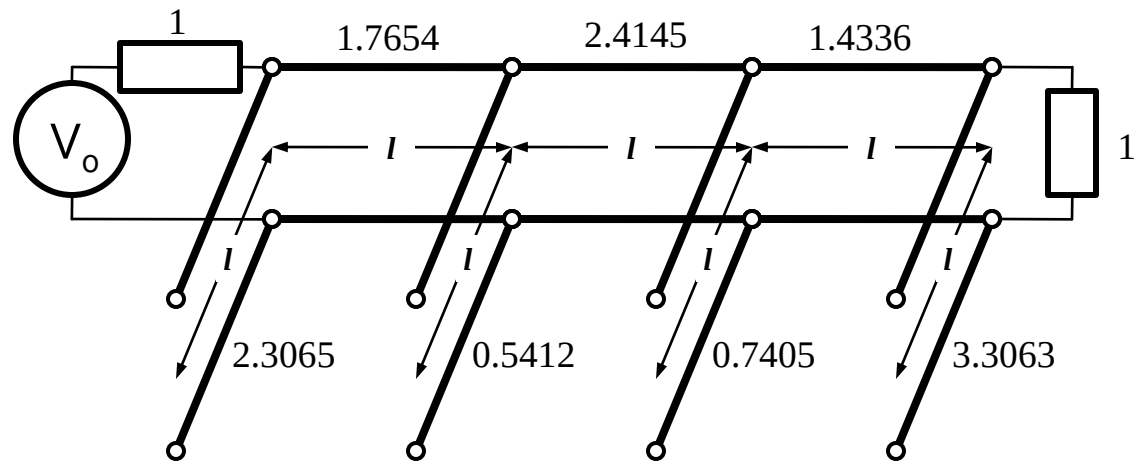


Exemplu

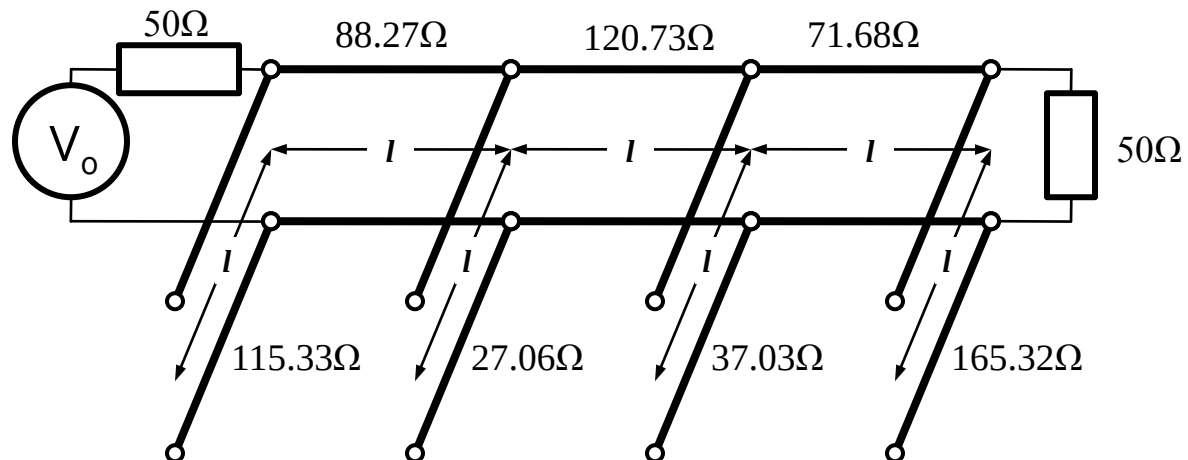
- Se mai adauga un element unitar in partea dreapta si se aplica Kuroda 2 de doua ori



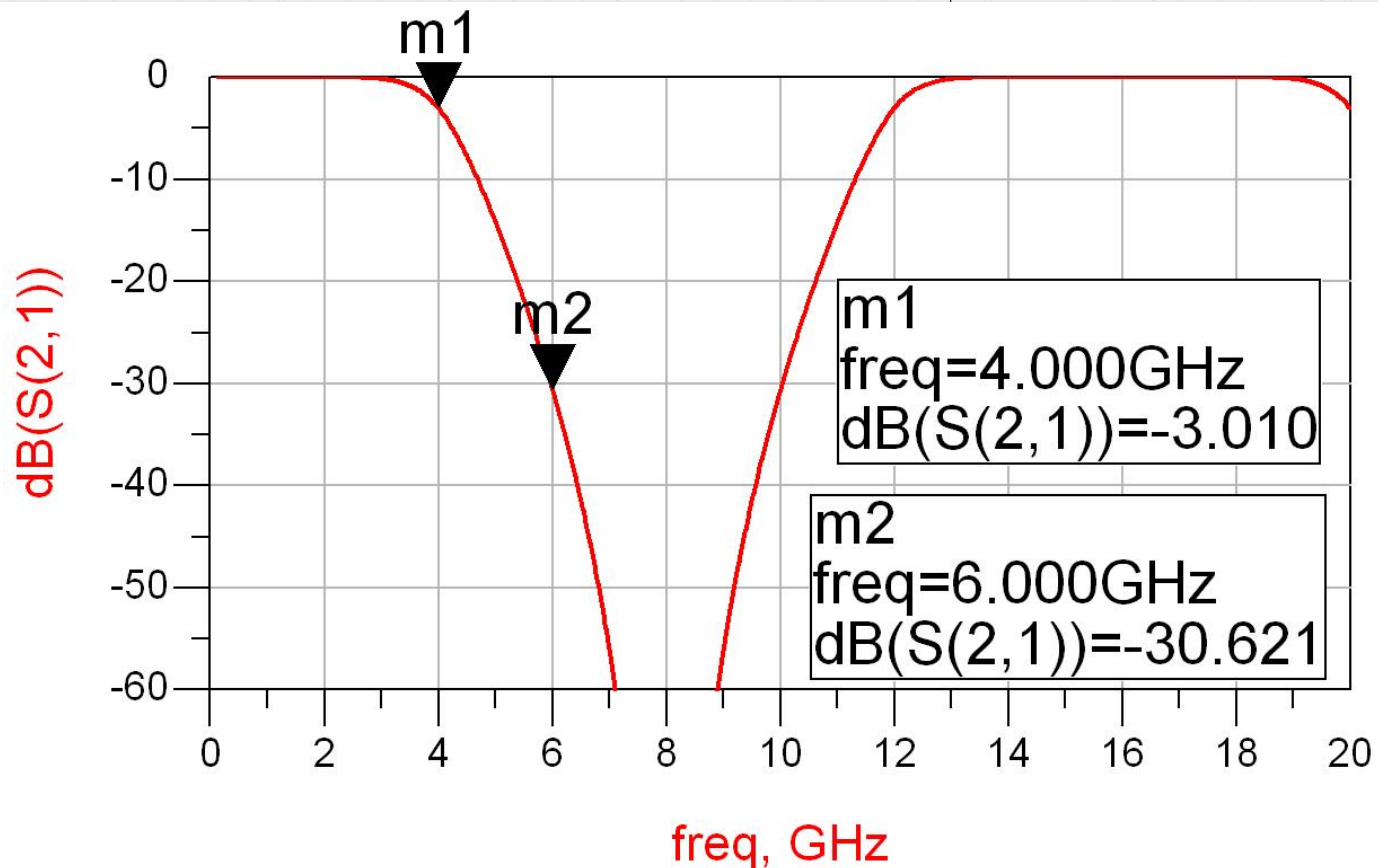
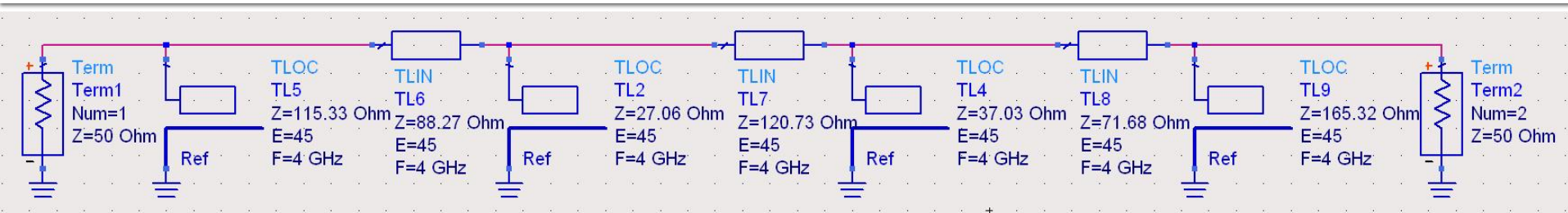
Exemplu



- Scalare la 50Ω



Identitatile Kuroda – ADS



Example

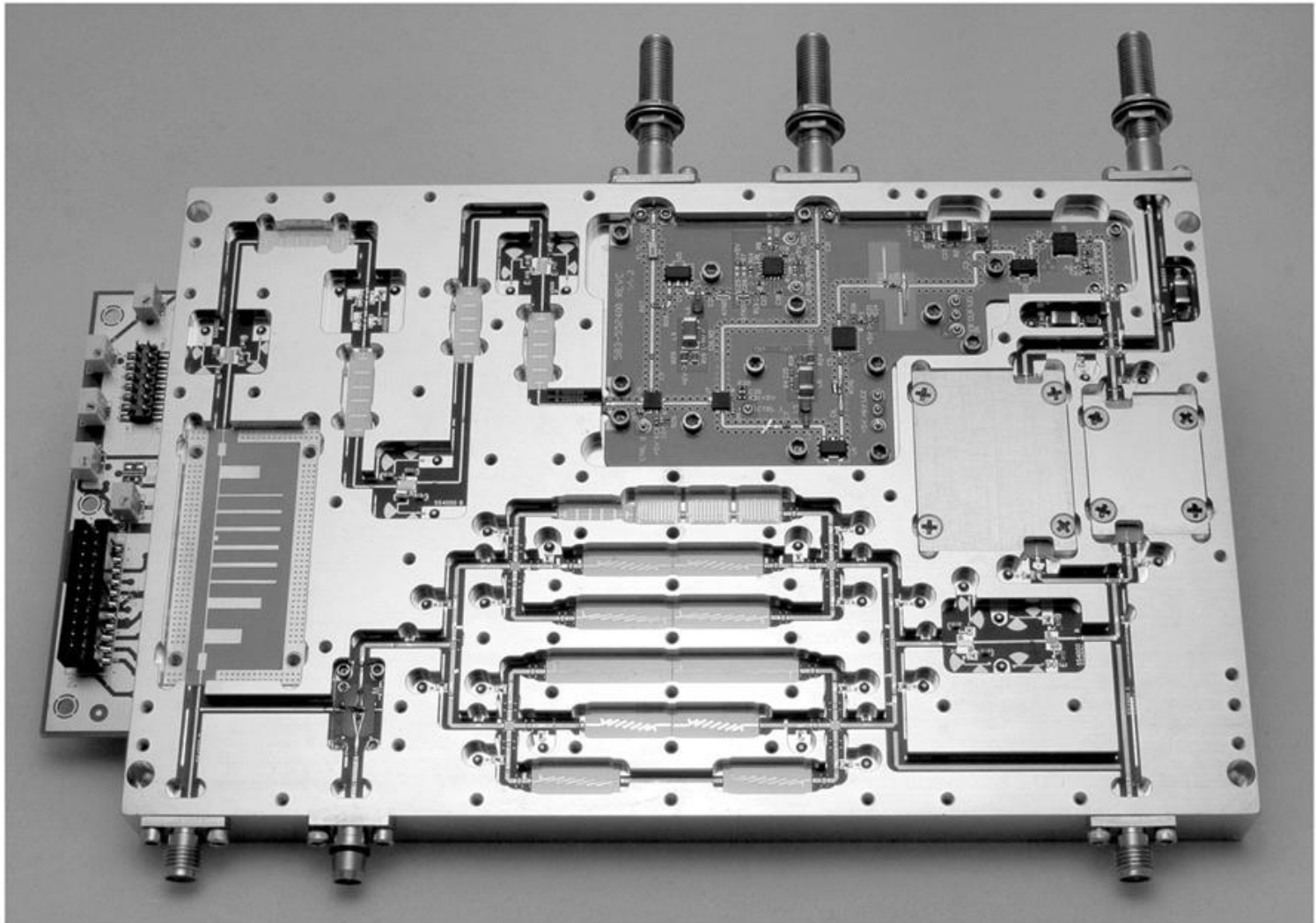


Figure 8.55
Courtesy of LNX Corporation, Salem, N.H.

Example

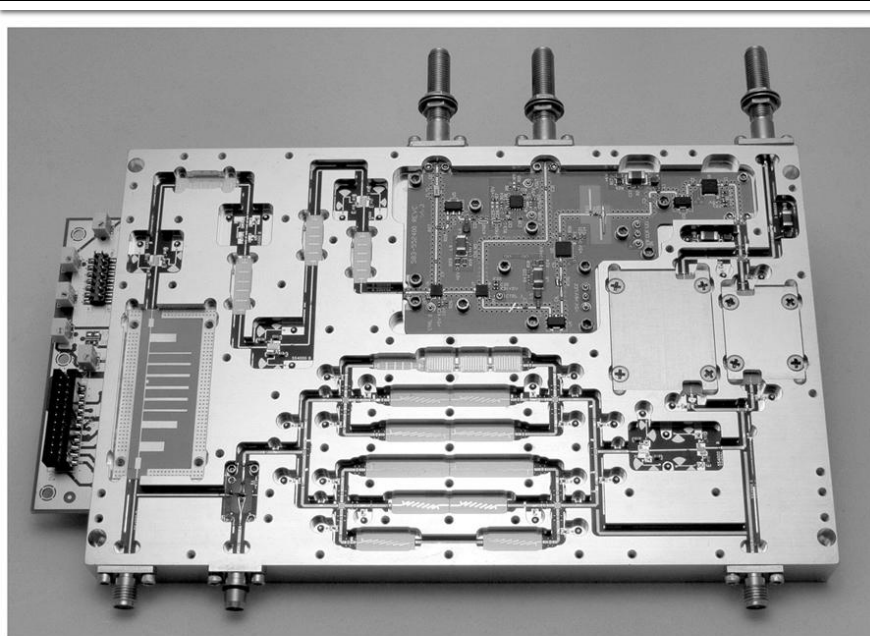
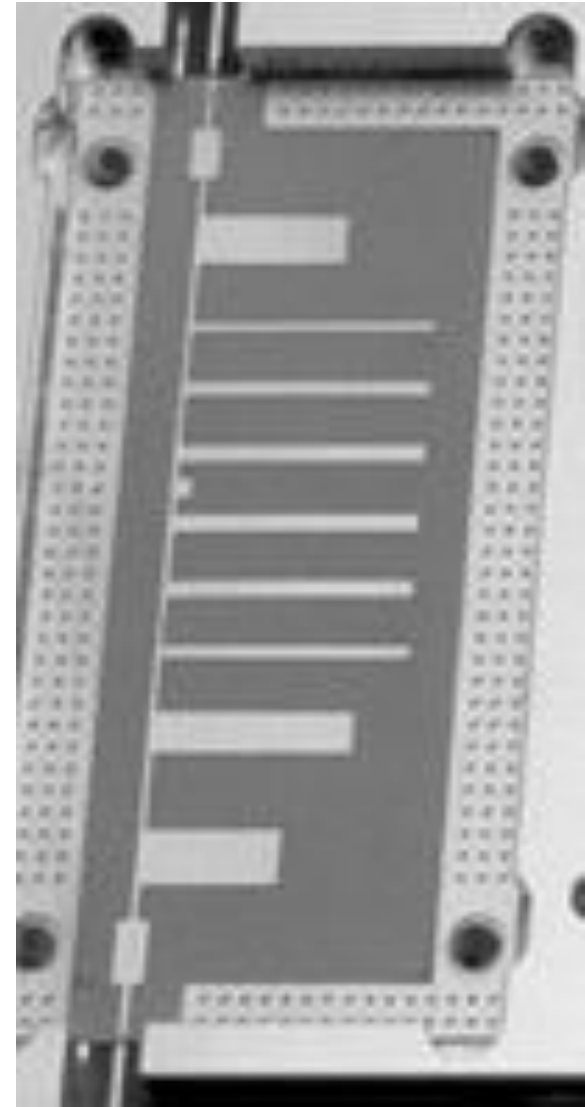
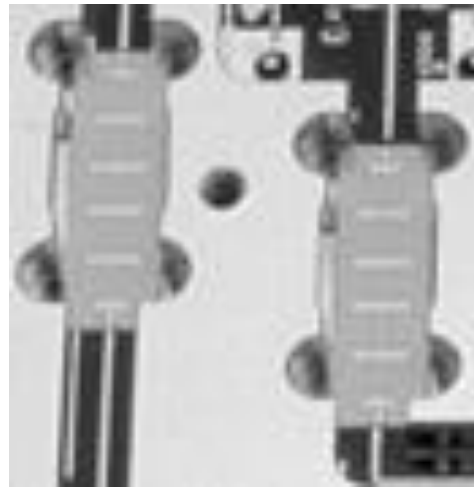
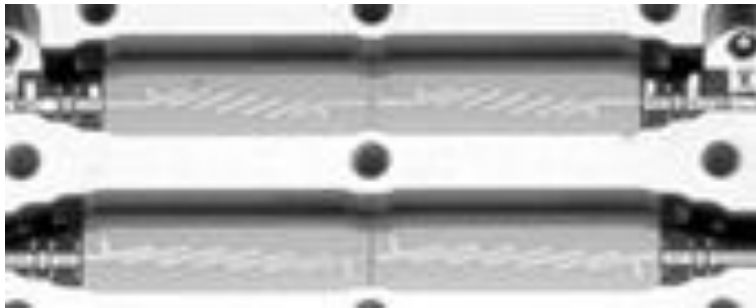


Figure 8.55
Courtesy of LNX Corporation, Salem, N.H.



Inversoare de impedanta si admitanta

- Transformarea Richard si identitatile Kuroda sunt utile mai ales pentru filtrele trece jos in tehnologiile in care stub-urile serie sunt dificil/imposibil de realizat (microstrip)
- De exemplu in cazul filtrului trece banda de ordinul 3:
 - se poate implementa inductanta serie utilizand K1-K2
 - capacitatea serie in schimb nu poate fi echivalata cu un stub paralel

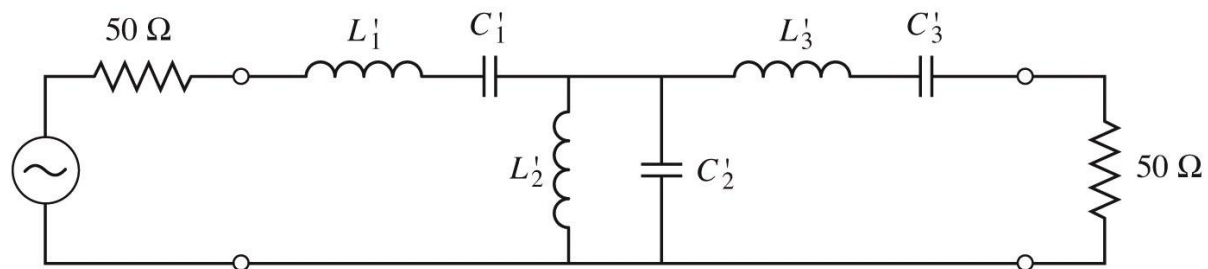


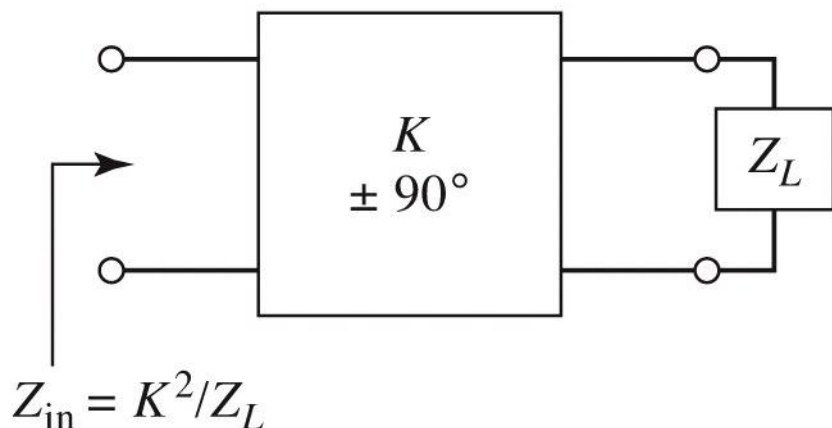
Figure 8.32
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Inversoare de impedanta si admitanta

- Pentru situatiile in care implementarea cu Richards + Kuroda nu ofera solutii practice se folosesc structuri de circuit numite **inversoare de impedanta si admitanta**

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

Impedance inverters



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

Admittance inverters

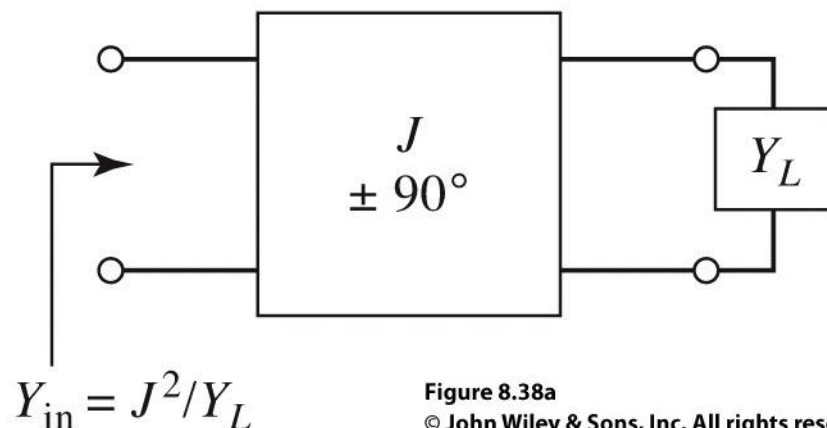


Figure 8.38a

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Inversoare de impedanta si admitanta

- Cel mai simplu exemplu de inversor de impedanta/admitanta este **transformatorul in sfert de lungime de unda** (C2)

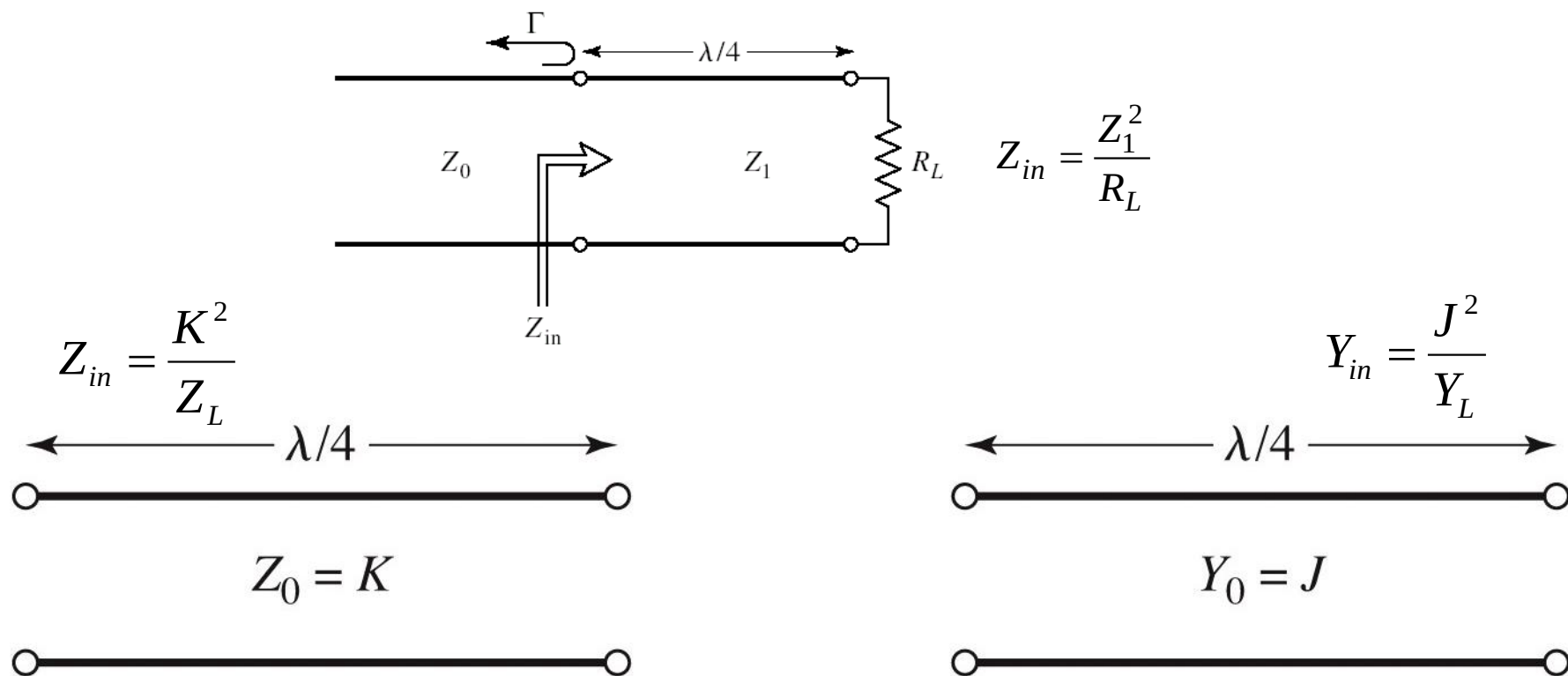
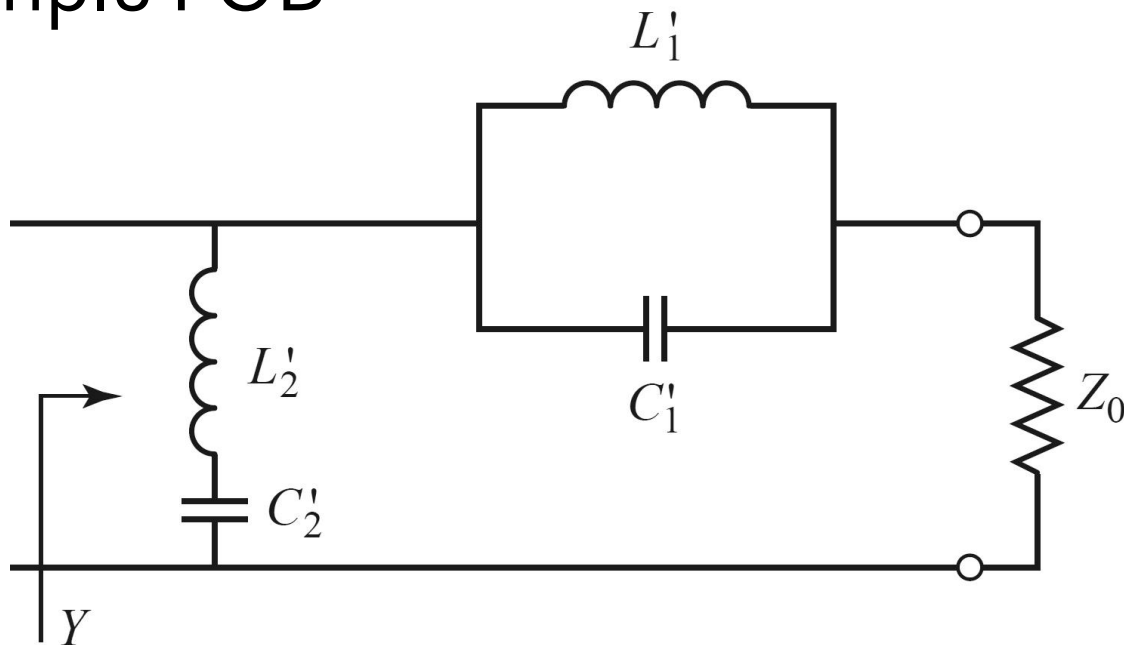


Figure 8.38b

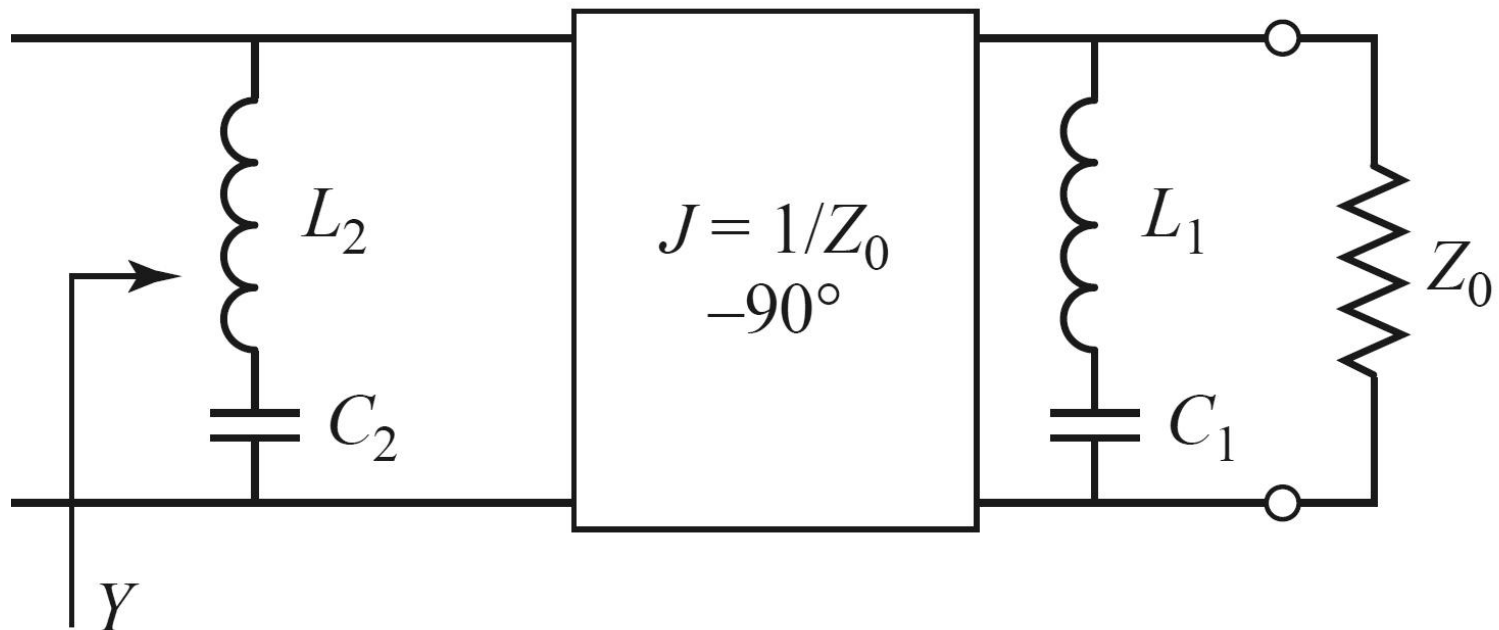
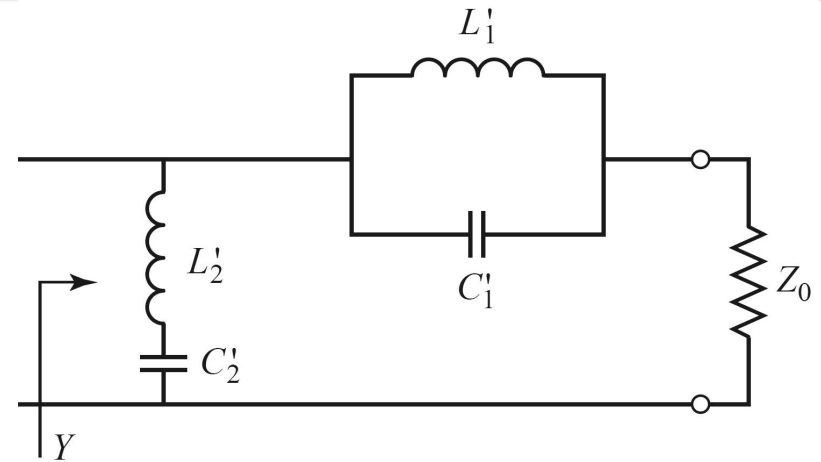
Inversoare de impedanta si admitanta

- Inversoarele de impedanta/admitanta pot fi utilizate pentru a schimba structura filtrelor in forme realizabile
- Exemplu FOB



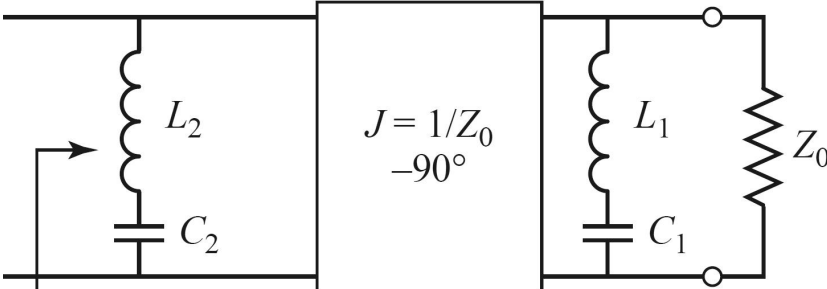
Inversoare de impedanta si admitanta

- Elementele serie pot fi eliminate prin introducerea unui inversor de admitanta



Inversoare de impedanta si admitanta

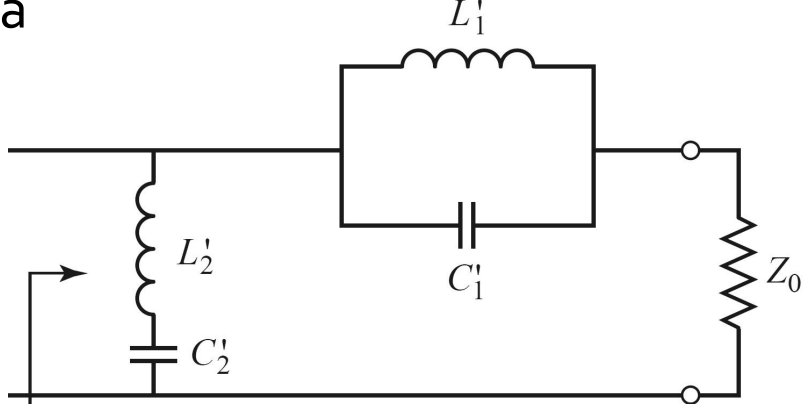
- Echivalenta celor doua scheme se demonstreaza prin obtinerea aceleiasi admitante de intrare
- Echivalenta completa se obtine prin incadrarea grupului simulat intre doua invertoare de admitanta



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

$$L_n \cdot C_n = L'_n \cdot C'_n = \frac{1}{\omega_0^2} \Rightarrow \frac{1}{Z_0^2} \cdot \sqrt{\frac{L_1}{C_1}} = \sqrt{\frac{C'_1}{L'_1}} \Rightarrow Y = Y'$$

$$\sqrt{\frac{L_2}{C_2}} = \sqrt{\frac{L'_2}{C'_2}}$$

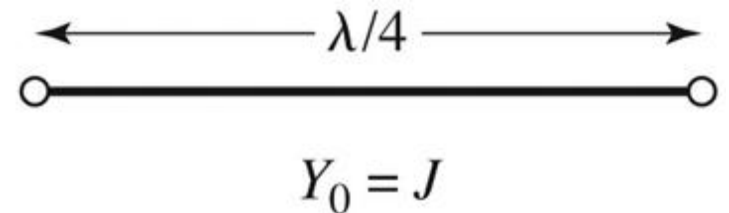
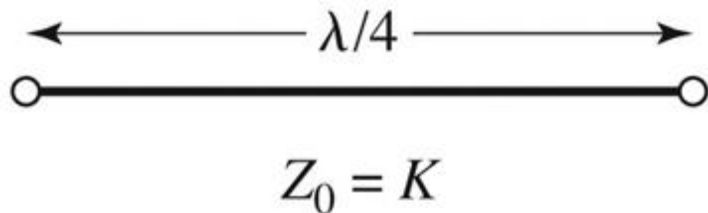


$$Y' = \frac{1}{j \cdot \omega \cdot L'_2 + \frac{1}{j \cdot \omega \cdot C'_2}} + \left(\frac{1}{j \cdot \omega \cdot C'_1 + \frac{1}{j \cdot \omega \cdot L'_1}} + Z_0 \right)^{-1}$$

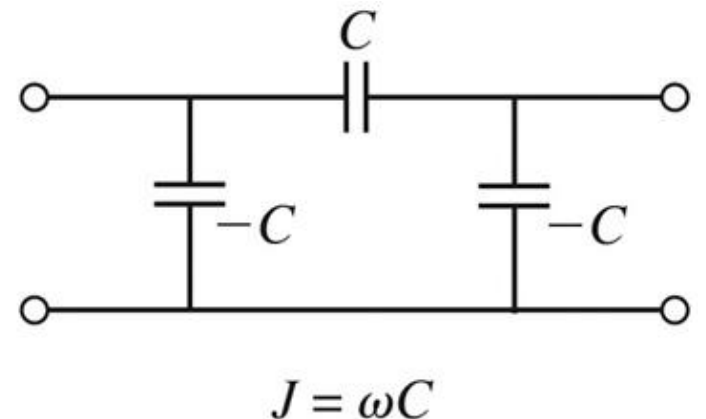
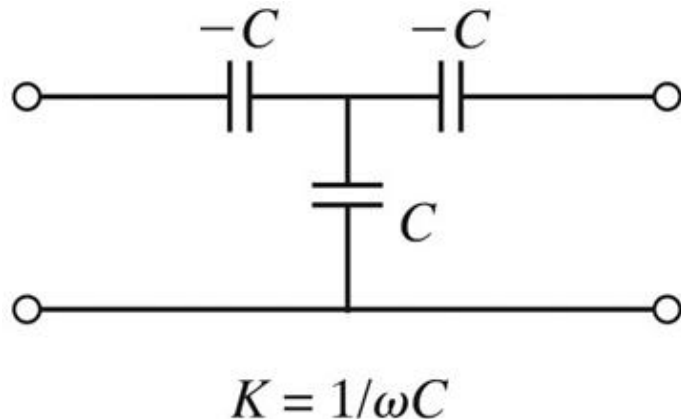
- Rezultat similar se obtine si pentru filtrul trece banda
- Un grup **LC serie** introdus in serie se poate inlocui cu un grup **LC paralel** introdus in paralel incadrat de doua invertoare de admitanta

Realizari practice de inversoare de impedanta/admitanta

- Cel mai uzual se foloseste transformatorul in sfert de lungime de unda

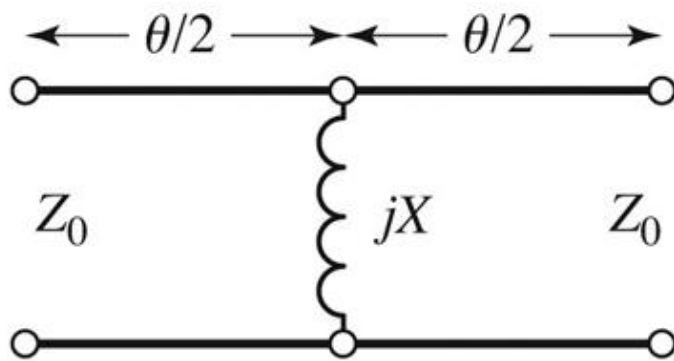


- Realizare cu elemente concentrate



Realizari practice de inversoare de impedanta/admitanta

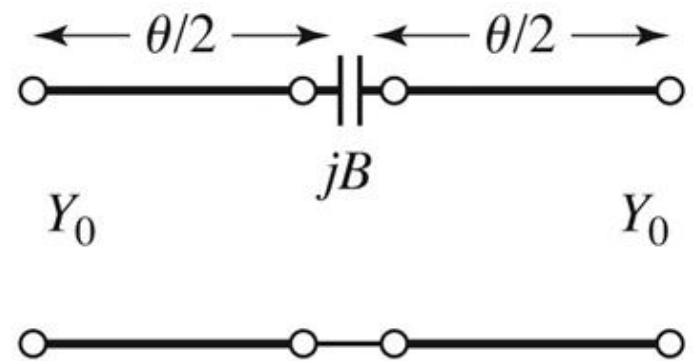
■ Realizare cu linii



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

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

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



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

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

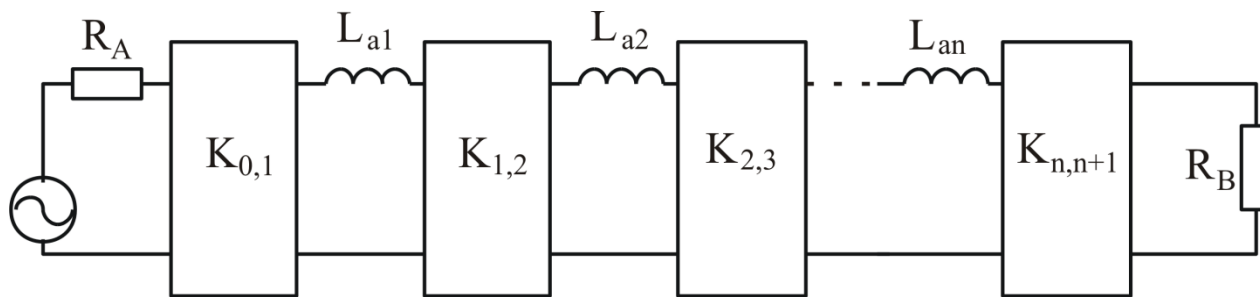
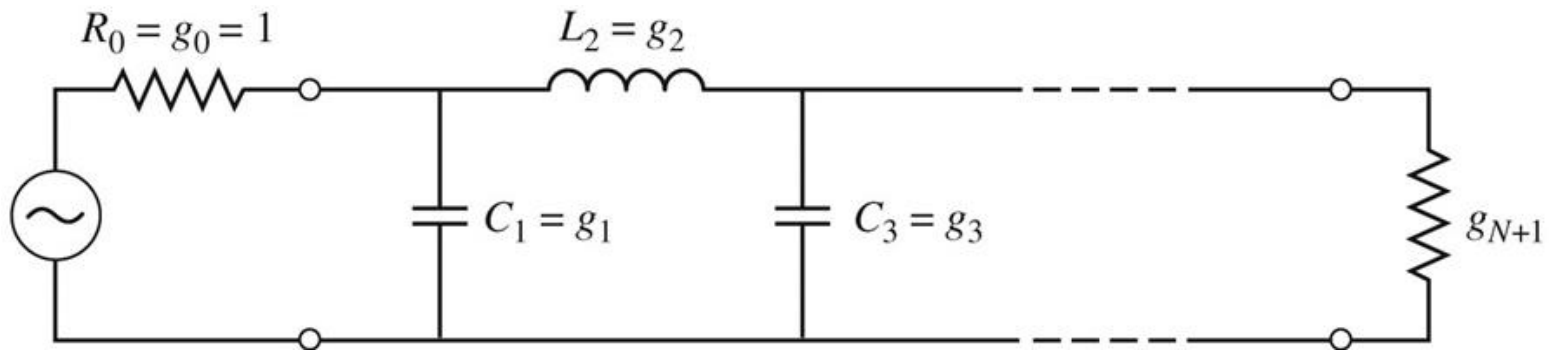
$$\theta = -\tan^{-1} \frac{2 \cdot B}{Y_0}$$

Red arrows from the theta terms in the equations above point to this result:

$$\theta < 0$$

Filtre prototip cu inversoare

- Utilizand inversoare de admitanta se pot implementa filtrele prototip utilizand un singur tip de element



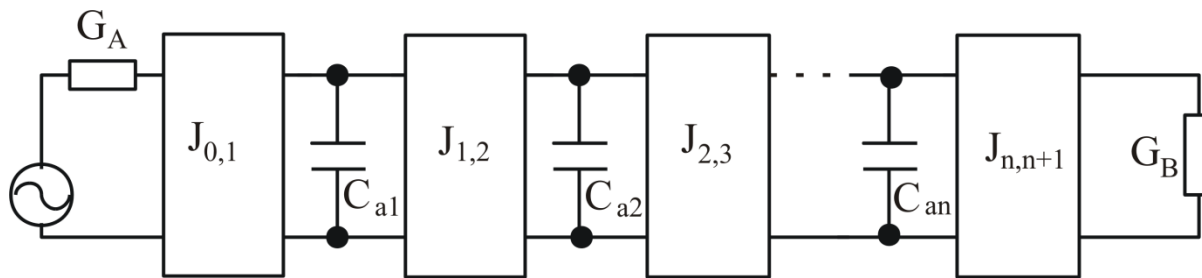
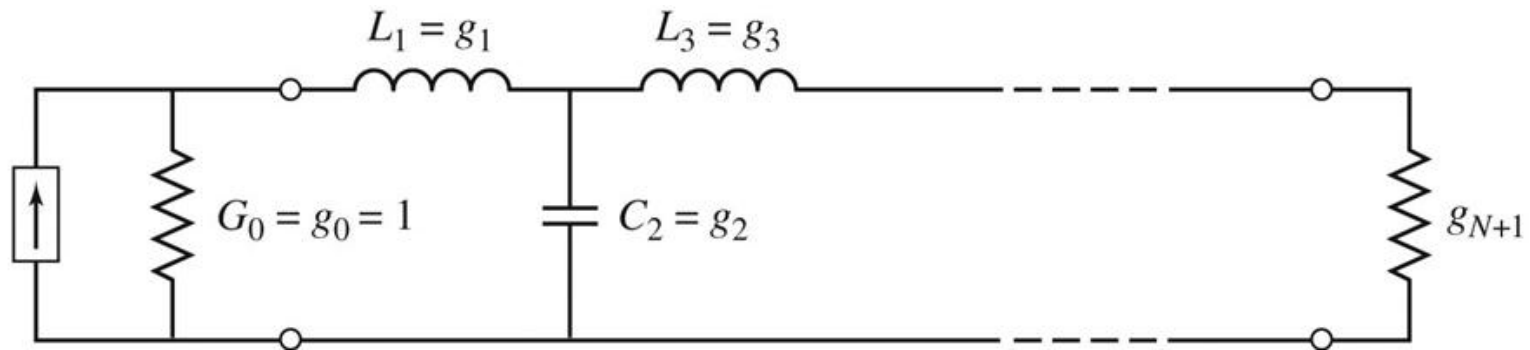
$$K_{0,1} = \sqrt{\frac{R_A \cdot L_{a,1}}{g_0 \cdot g_1}}$$

$$K_{k,k+1} \Big|_{k=1, n-1} = \sqrt{\frac{L_{a,k} \cdot L_{a,k+1}}{g_k \cdot g_{k+1}}}$$

$$K_{n,n+1} = \sqrt{\frac{L_{a,n} \cdot R_B}{g_n \cdot g_{n+1}}}$$

Filtre prototip cu inversoare

- Utilizand inversoare de admitanta se pot implementa filtrele prototip utilizand un singur tip de element



$$J_{0,1} = \sqrt{\frac{G_A \cdot C_{a,1}}{g_0 \cdot g_1}}$$

$$J_{k,k+1} \Big|_{k=1, n-1} = \sqrt{\frac{C_{a,k} \cdot C_{a,k+1}}{g_k \cdot g_{k+1}}}$$

$$J_{n,n+1} = \sqrt{\frac{C_{a,n} \cdot g_B}{g_n \cdot g_{n+1}}}$$

Filtre prototip cu inversoare

- Pentru filtrele prototip cu inversoare exista $2 \cdot N + 1$ parametri si $N + 1$ ecuatii care asigura echivalenta raspunsului deci N parametri pot fi alesi din considerente oarecare
 - se pot alege valorile reactantelor, urmand ca parametrii inversoarelor sa rezulte din calcul
 - se pot alege convenabil inversoarele, urmand ca reactantele sa rezulte din ecuatiile de echivalare

Filtre prototip cu inversoare

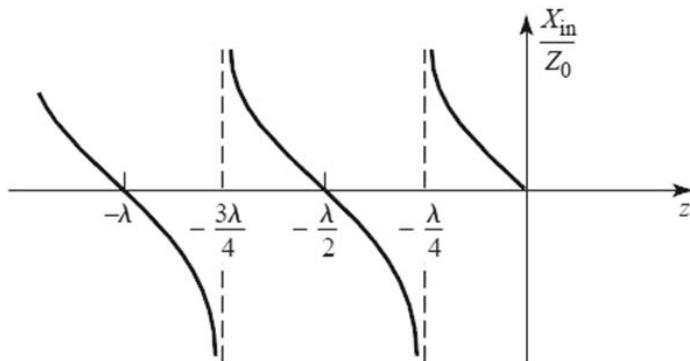
- Principiul se poate aplica si pentru filtrele trece banda/opreste banda, acestea putand fi realizate din $N+1$ inversoare si N rezonatori (grupuri LC serie sau paralel cu frecventa de rezonanta ω_0) conectate fie in serie fie in paralel intre inversoare
 - FTB se realizeaza cu
 - grup LC serie conectat in serie intre inversoare
 - grup LC paralel conectat in paralel intre inversoare
 - FOB se realizeaza cu
 - grup LC paralel conectat in serie intre inversoare
 - grup LC serie conectat in paralel intre inversoare

Linii ca rezonatoare

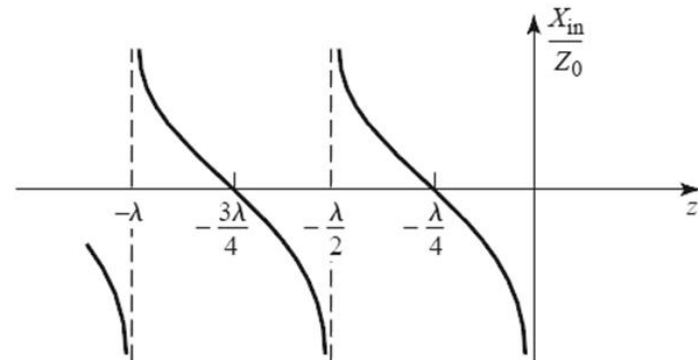
- Impedanta de intrare intr-o linie (stub) scurtcircuitata sau lasata in gol la capat manifesta comportament rezonant care poate fi utilizat pentru implementarea rezonatoarelor

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

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

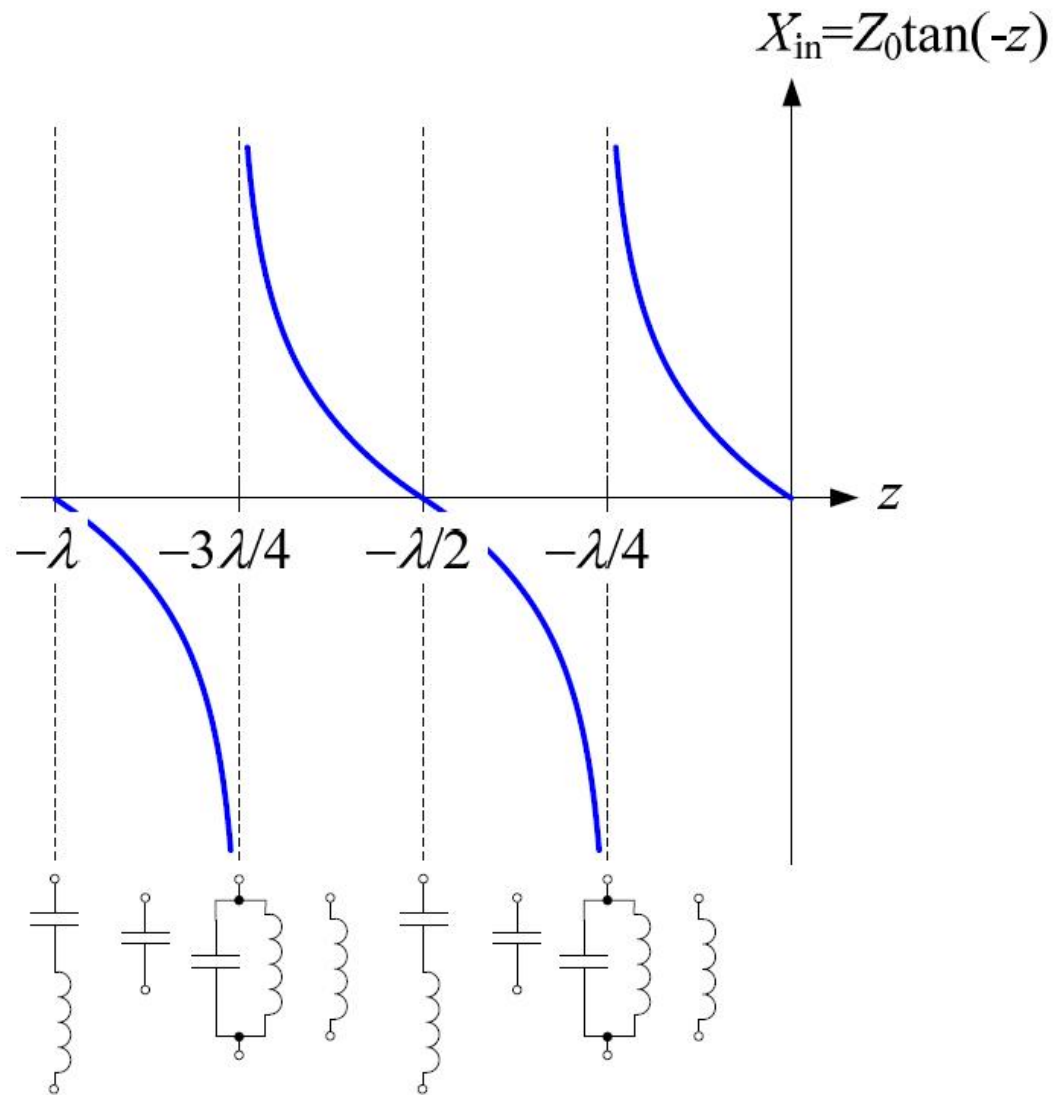


$$Z_{in,g} = -j \cdot Z_0 \cdot \cot \beta \cdot l$$



Linii ca rezonatoare

- Linie in scurtcircuit
- Pentru frecventa (ω_0) la care $l = \lambda/4$ se obtine un circuit rezonant LC paralel
 - linia are comportament capacitiv pentru frecvente mai mici ($l > \lambda/4$)
 - linia are comportament inductiv pentru frecvente mai mari ($l < \lambda/4$)
- Discutie similara pentru linia in gol (LC serie la frecventa la care $l = \lambda/4$)



FTB/FOB relatii de proiectare

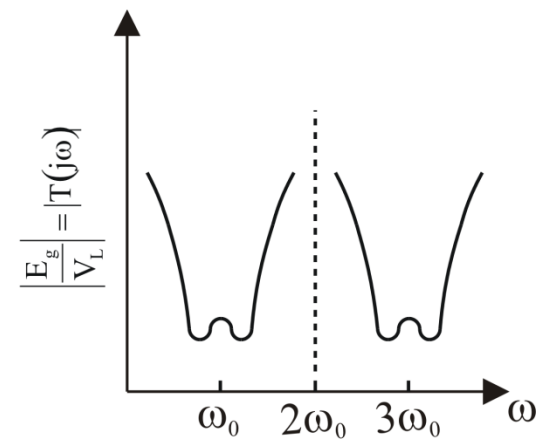
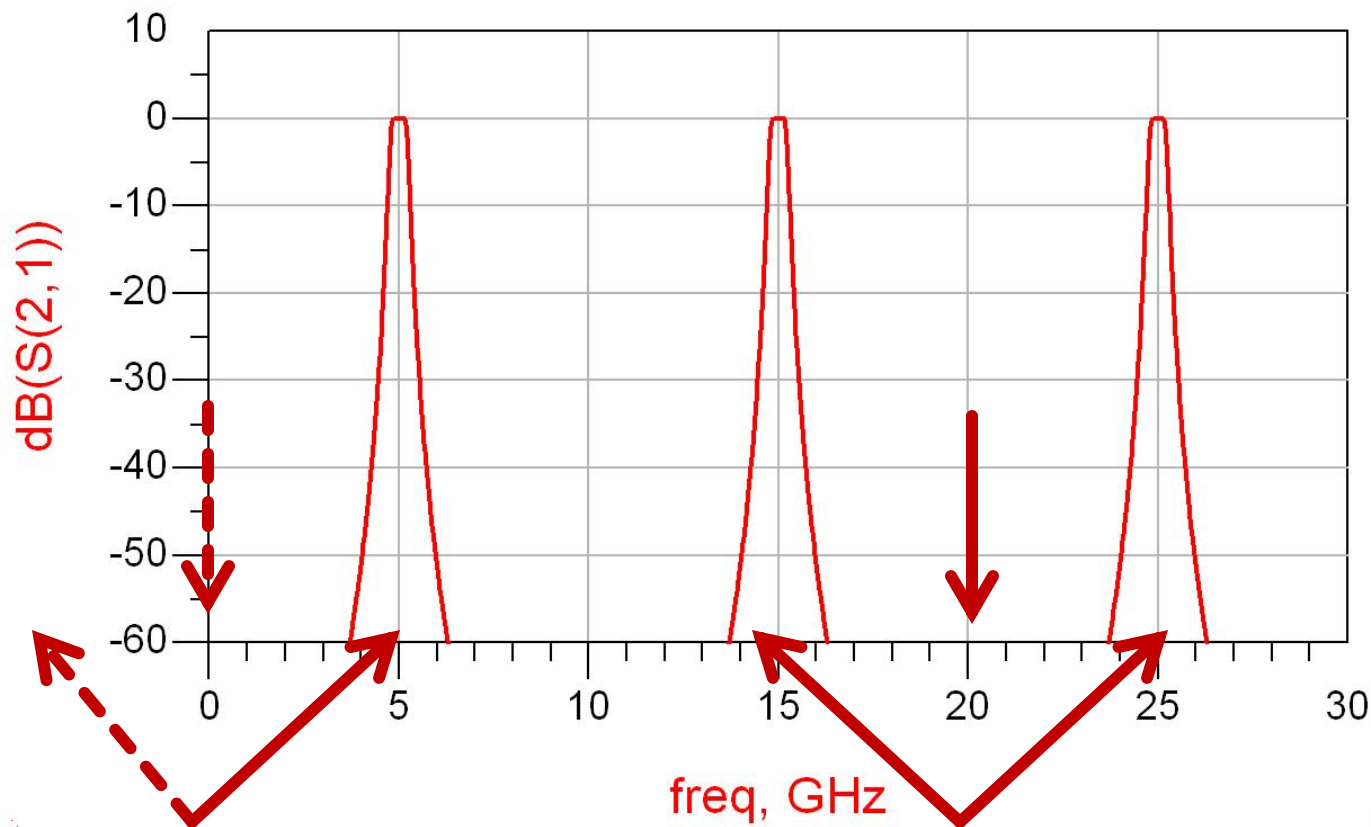
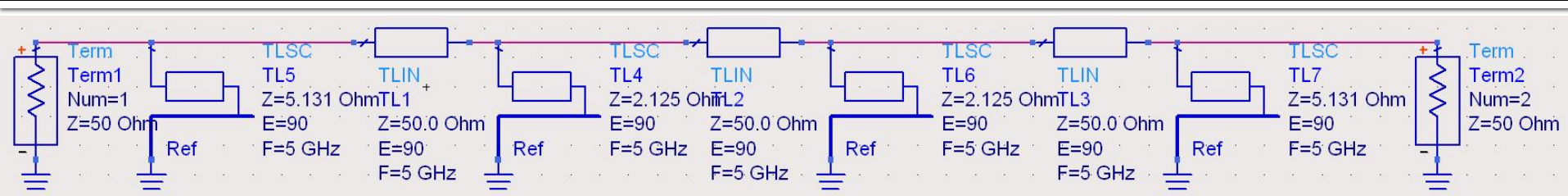
- Pentru cazul particular in care se implementeaza inversoarele de admitanta cu transformatoare in sfert de lungime de unda si impedanta caracteristica Z_0
 - FTB – stub-uri paralel scurtcircuitate la $l = \lambda/4$
$$Z_{0n} \approx \frac{\pi \cdot Z_0 \cdot \Delta}{4 \cdot g_n}$$
 - FOB – stub-uri paralel in gol de lungime $l = \lambda/4$
$$Z_{0n} \approx \frac{4 \cdot Z_0}{\pi \cdot g_n \cdot \Delta}$$
 - ! necesitatea Z_0 la ambele capete → **nu** se pot utiliza relatiile pentru filtru **echiriplu** de ordin **par**

Exemplu

- Similar cu o tema de proiect
- Continuarea amplificatorului C11
- Filtru trece banda de ordinul 4, $f_0 = 5\text{GHz}$, banda 10 %
- Tabel maxim plat sau relatii de calcul:

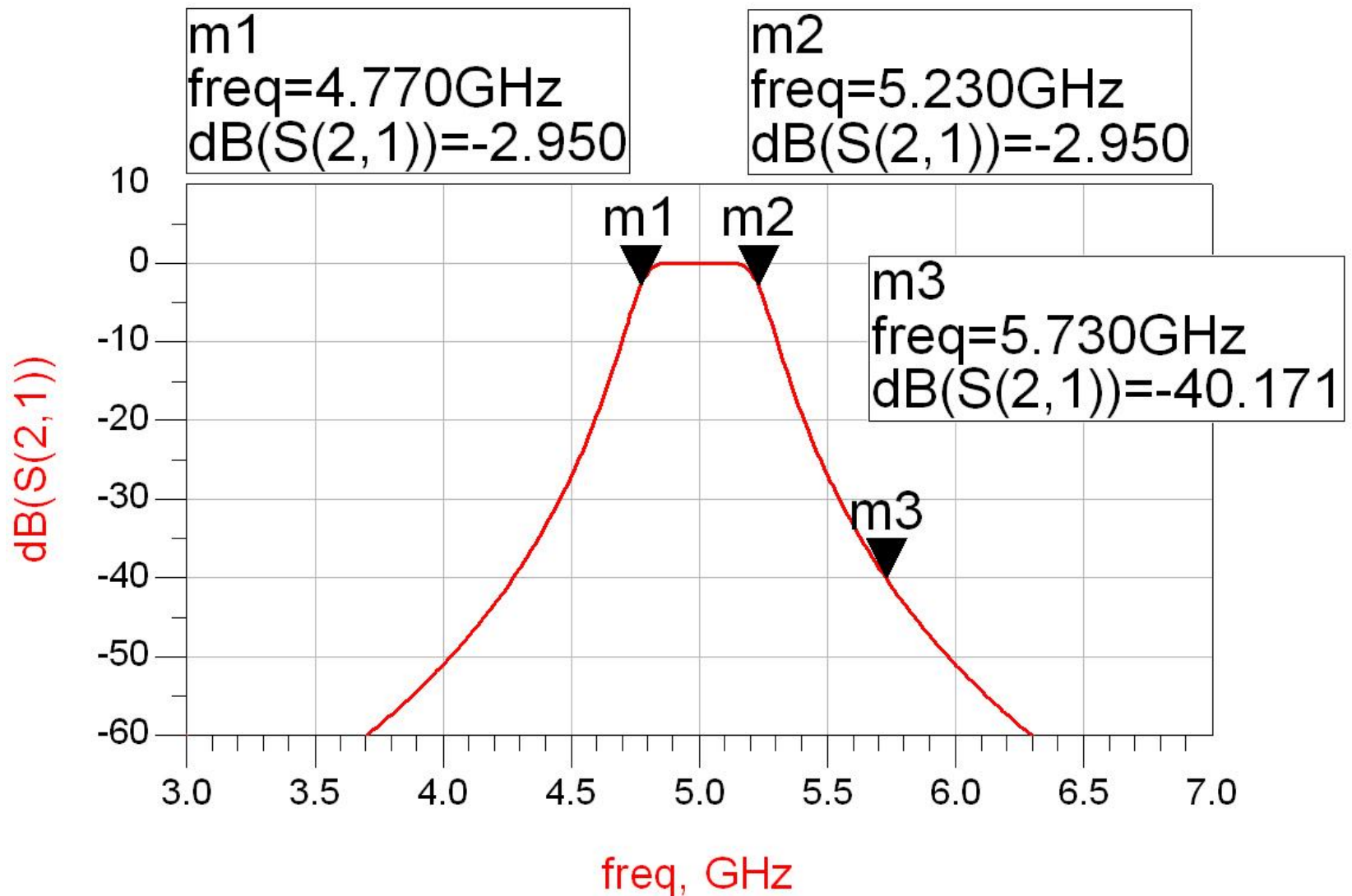
n	g_n	$Z_{on}(\Omega)$
1	0.7654	5.131
2	1.8478	2.125
3	1.8478	2.125
4	0.7654	5.131

ADS – FTB

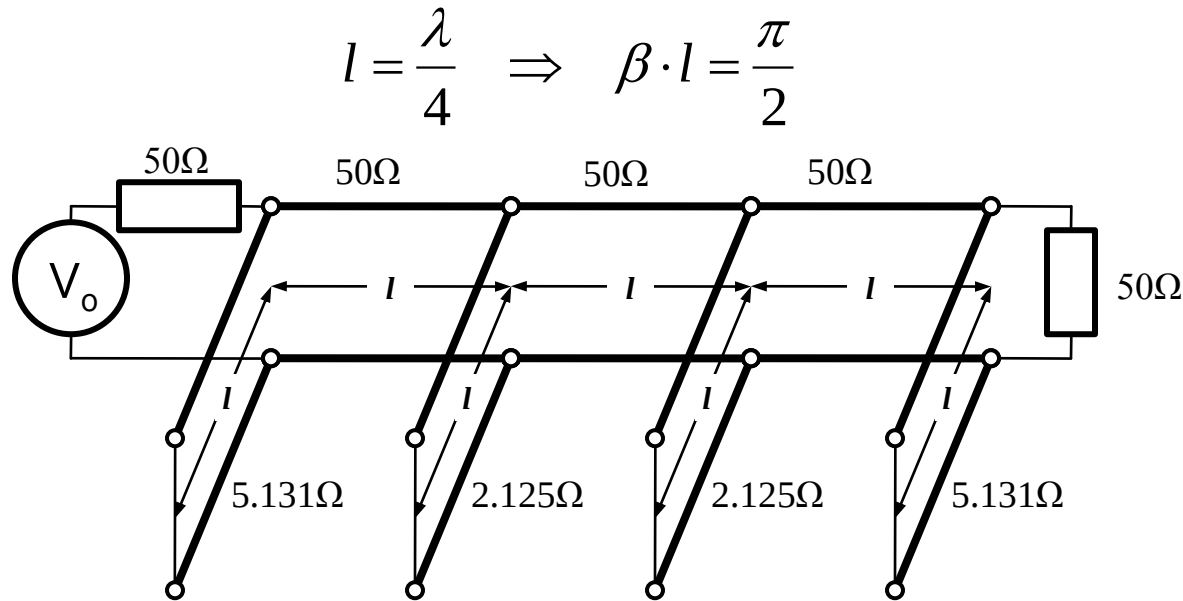


- raspunsul filtrului se repeta la fiecare $4 \cdot \omega_c$

ADS – FTB



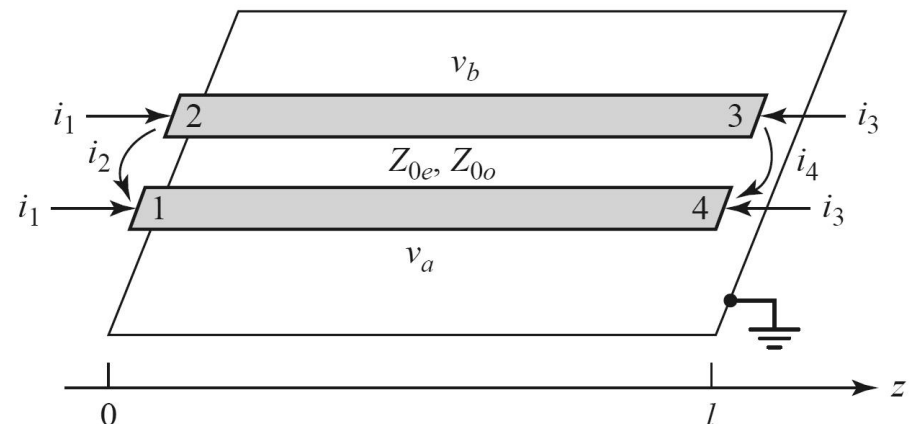
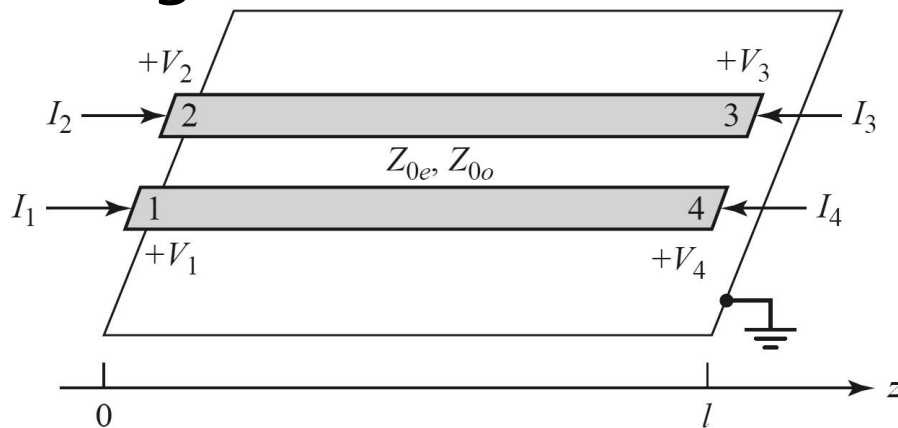
Exemplu



- Problemele filtrelor realizate cu linii ca rezonatoare si invertoare de impedanta
 - stub-uri in scurtcircuit (via-hole) pentru FTB
 - deseori impedantele caracteristice pentru stub-uri rezulta de **valori** dificil de realizat in practica (2.125Ω)

Filtre cu linii cuplate

- Analiza sectiunilor de linii cuplate se face punand in evidenta comportarea pe modul par si pe modul impar
- Aceste moduri sunt caracterizate de impedantele caracteristice de mod par/impar a caror valoare va impune in functie de tehnologia utilizata geometria liniilor (latime/distanta dintre linii)



Linii cuplate

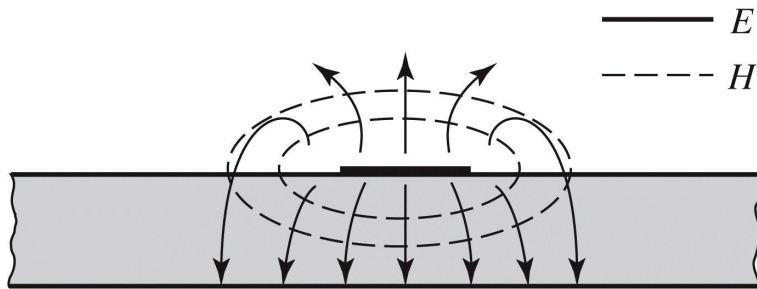
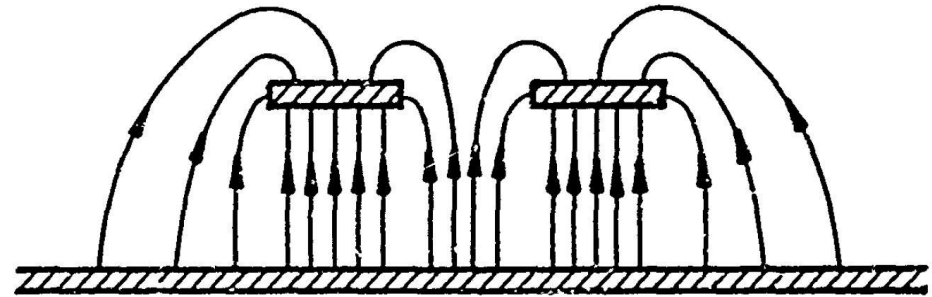
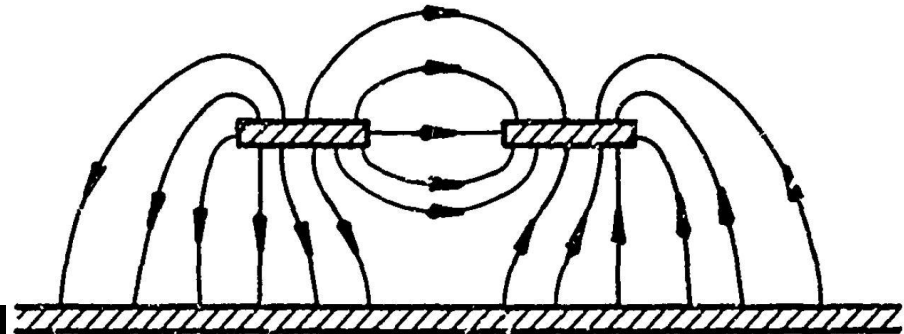


Figure 3.25b
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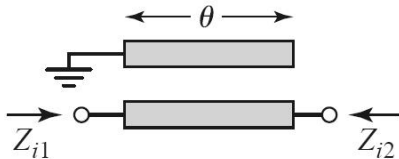
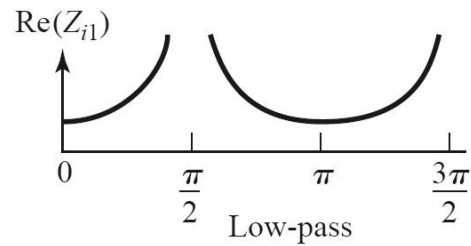
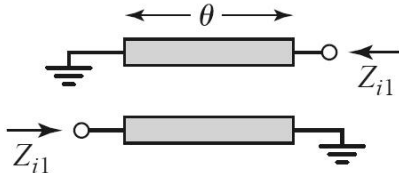
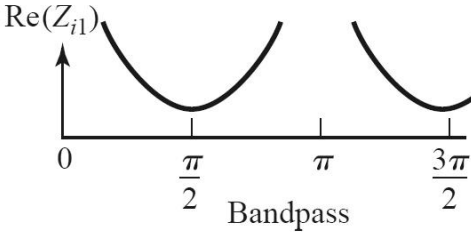
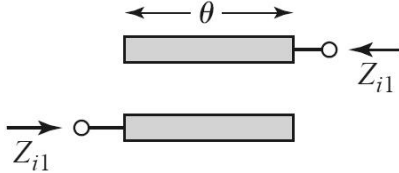
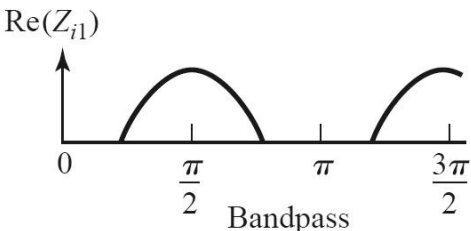
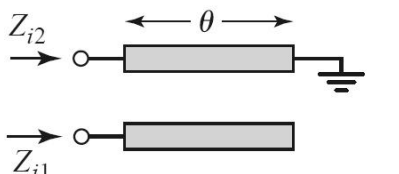
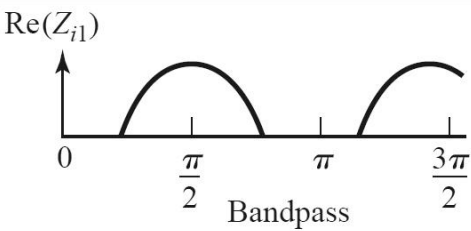
b) EVEN MODE ELECTRIC FIELD PATTERN (SCHEMATIC)



c) ODD MODE ELECTRIC FIELD PATTERN (SCHEMATIC)

- Mod par – caracterizeaza semnalul de mod comun de pe cele doua linii
- Mod impar – caracterizeaza semnalul de mod diferential dintre cele doua linii

Filtre cu linii cuplate

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

Filtre cu linii cuplate

- Filtru trece banda cu rezonanta la $\theta = \pi/2$ ($l = \lambda/4$)

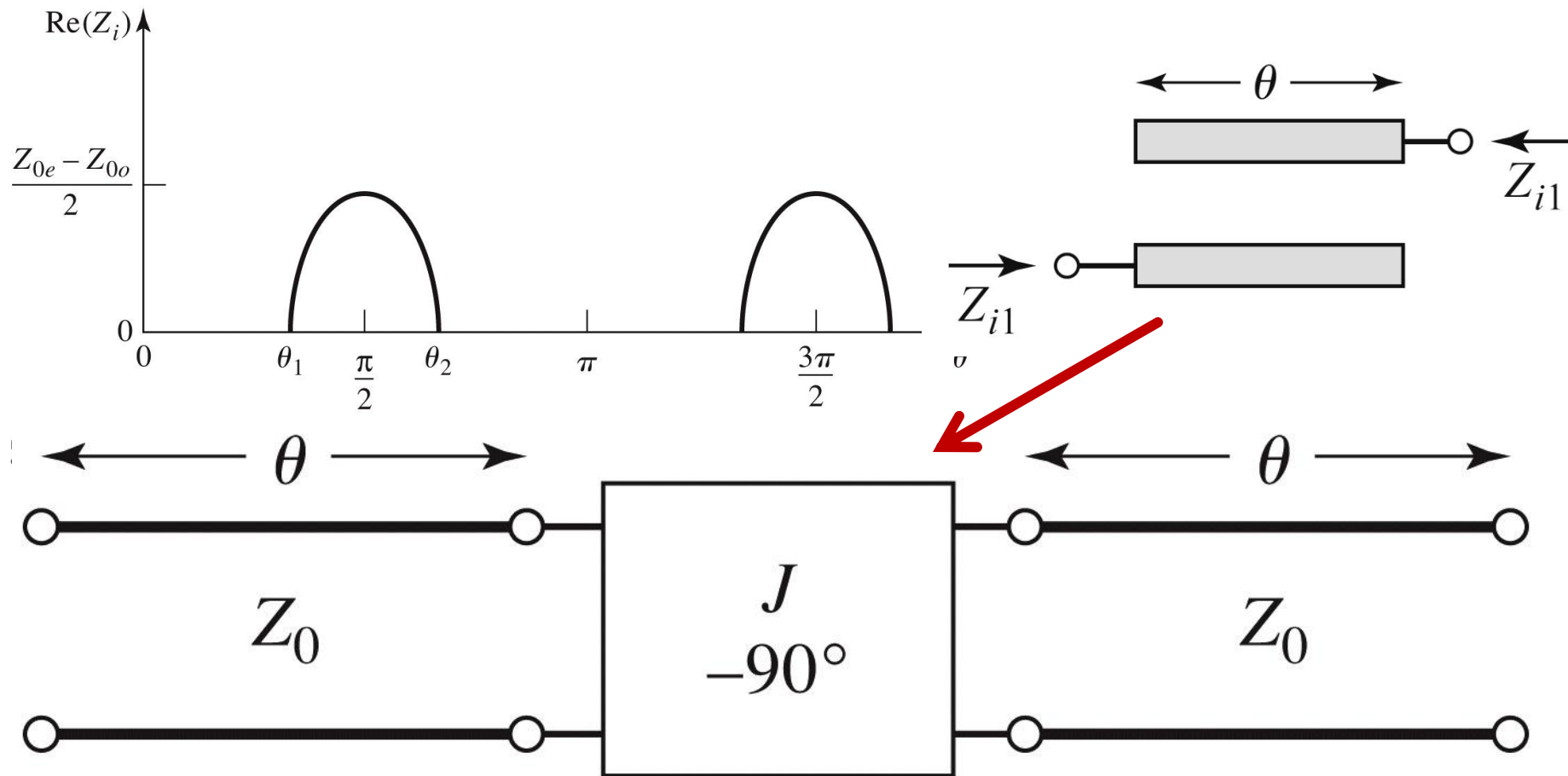
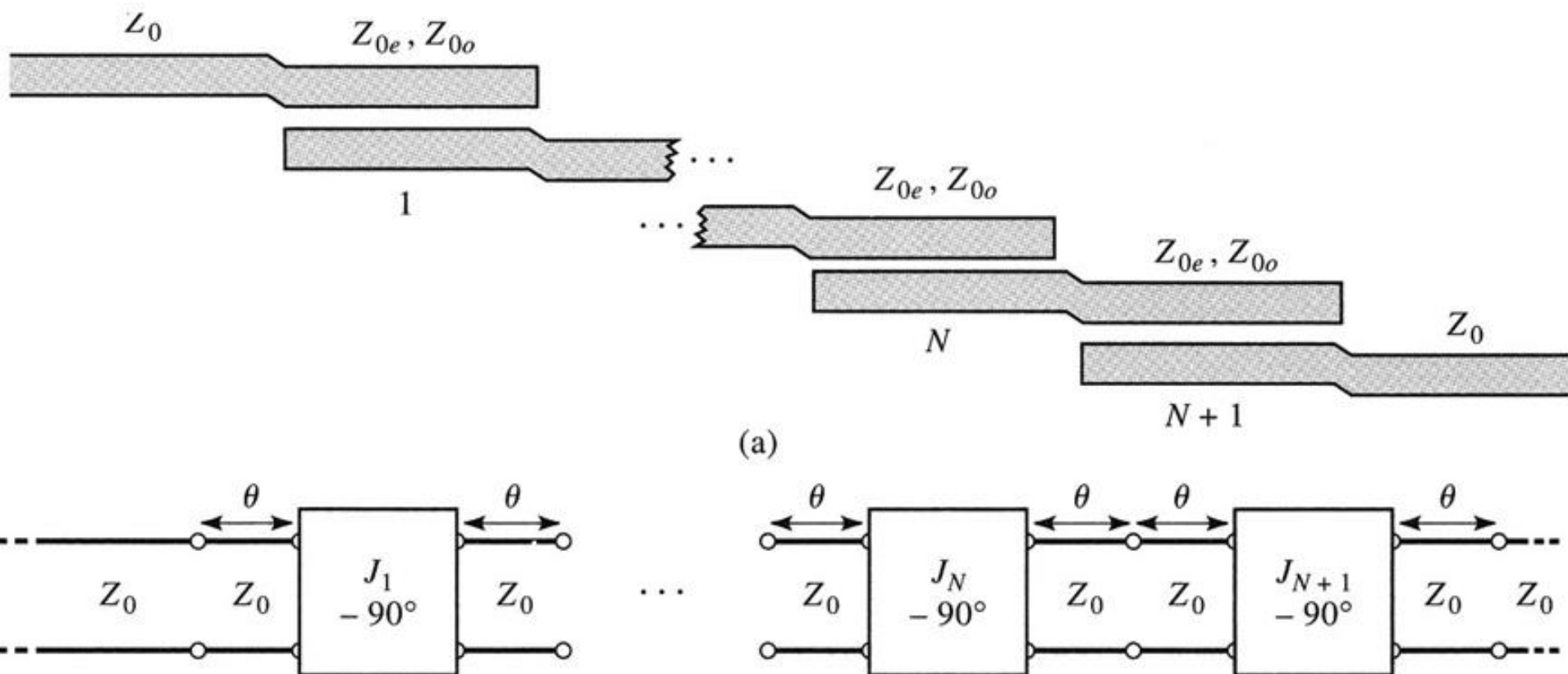


Figure 8.44

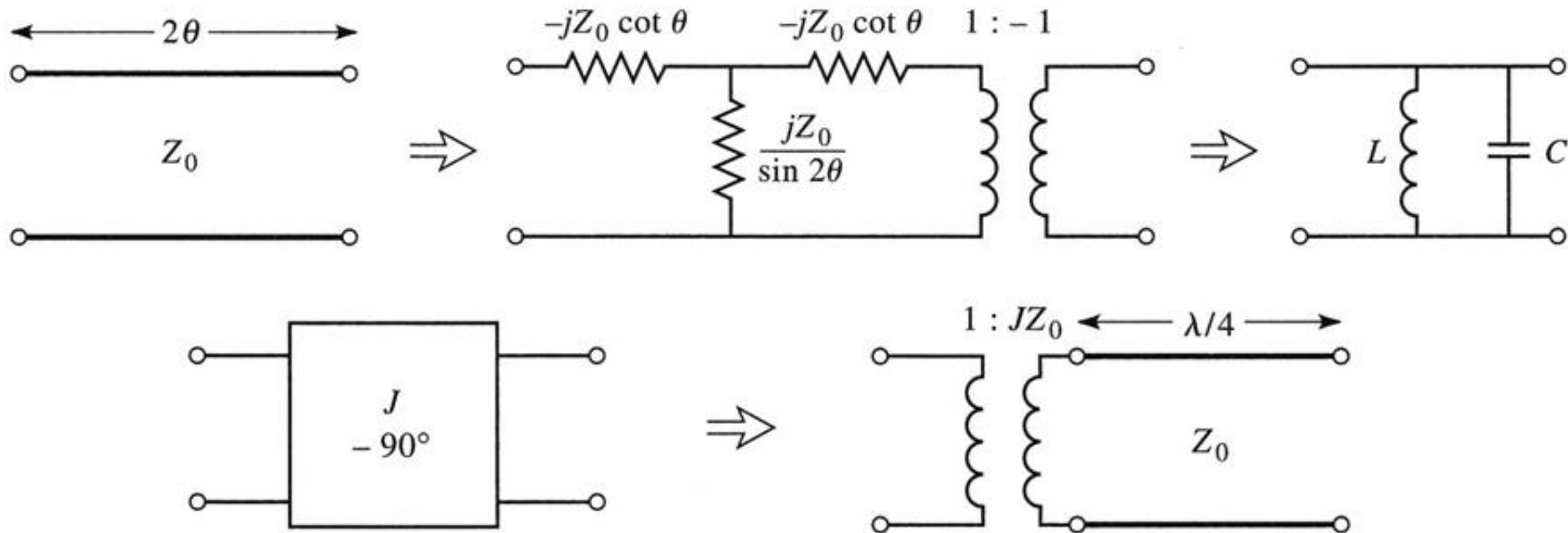
Filtre cu linii cuplate

- Un filtru cu $N+1$ sectiuni de linii cuplate



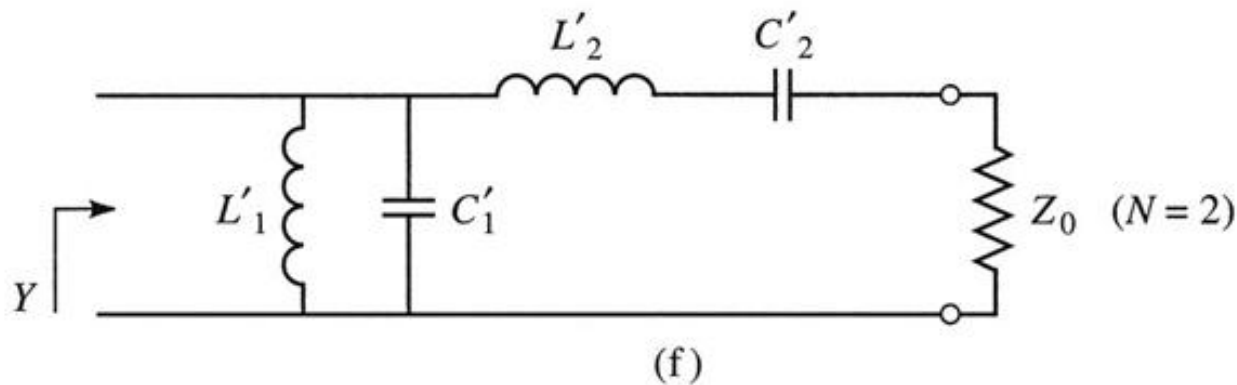
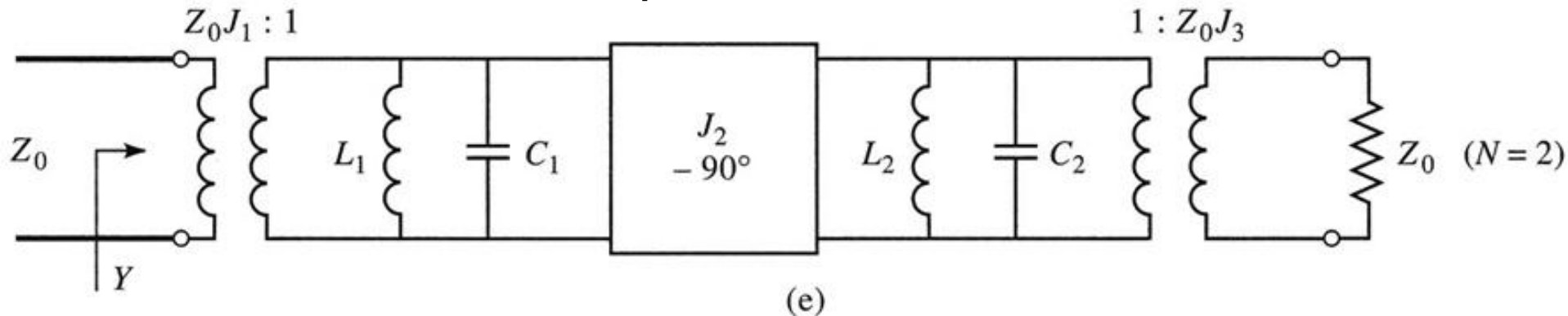
Filtre cu linii cuplate

- Se modeleaza
 - liniile
 - inversoarele



Filtre cu linii cuplate N=2

- Se obtine comportare de tip FTB de ordin 2 cu 3 sectiuni de linii cuplate



Proiectare filtre cu linii cuplate

- Se calculeaza inversoarele

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

- Se calculeaza liniile cuplate (toate de lungime $l = \lambda/4$)

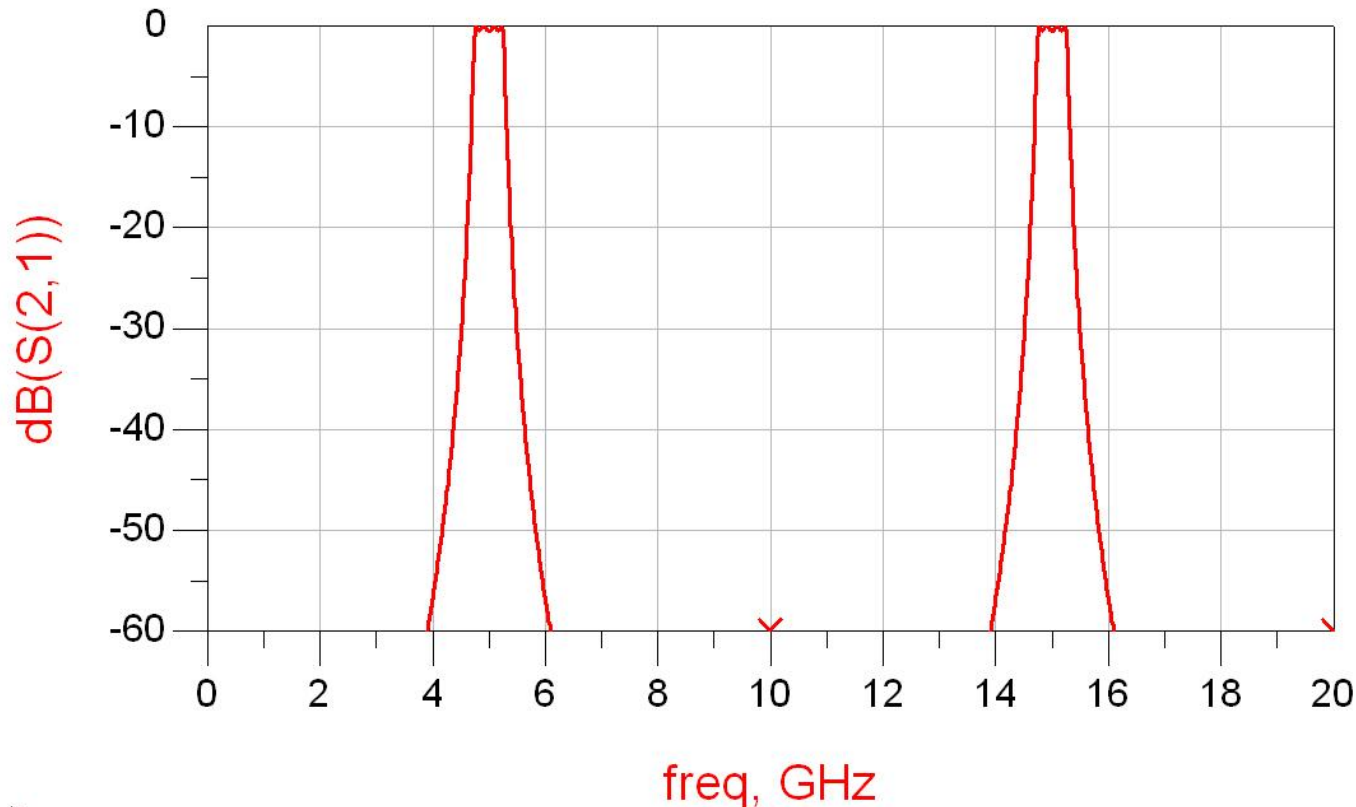
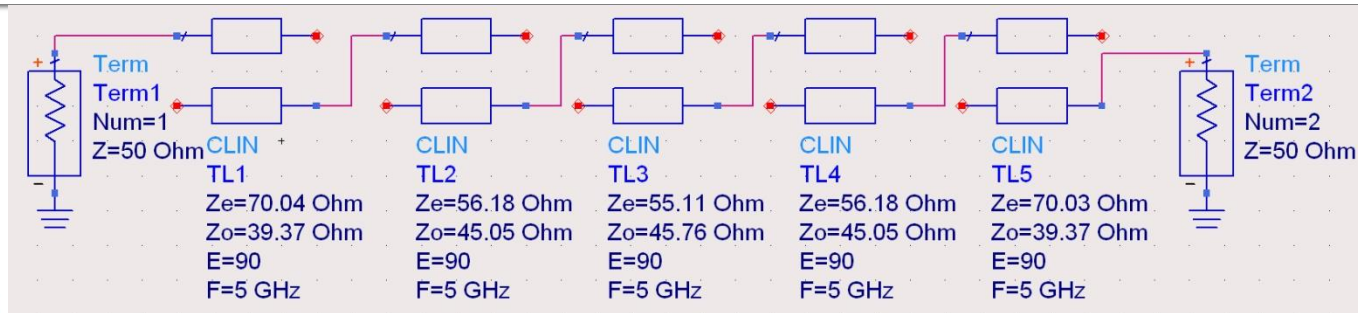
$$\begin{aligned} Z_{0e,n} &= Z_0 \cdot \left[1 + J_n \cdot Z_0 + (J_n \cdot Z_0)^2 \right] \\ Z_{0o,n} &= Z_0 \cdot \left[1 - J_n \cdot Z_0 + (J_n \cdot Z_0)^2 \right] \end{aligned} \quad n = \overline{1, N+1}$$

Exemplu

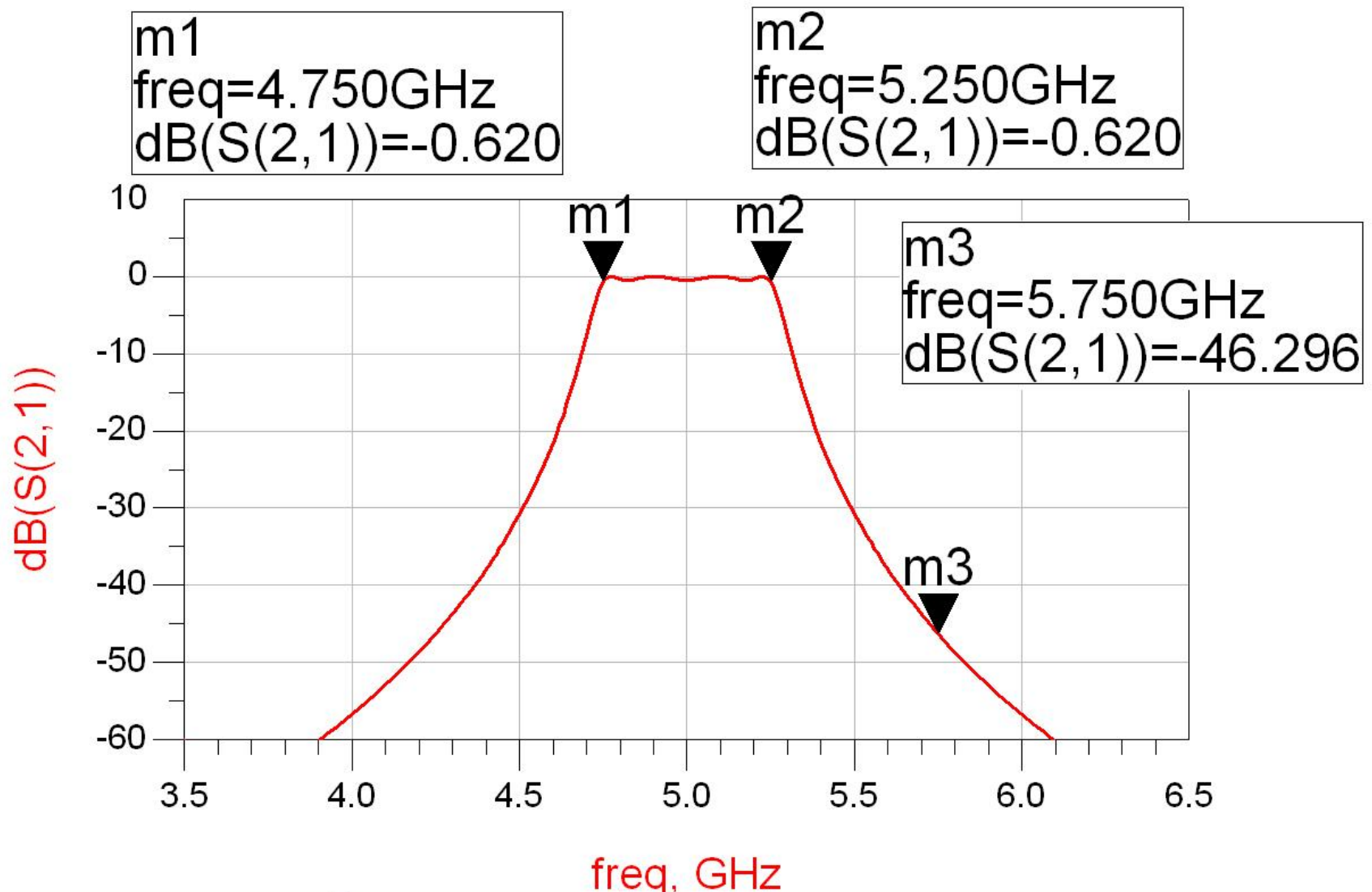
- Similar cu o tema de proiect
- Continuarea amplificatorului C11
- Filtru trece banda de ordinul 4, $f_0 = 5\text{GHz}$, banda 10 %
- Tabel echiriplu 0.5dB sau relatii de calcul: g_n

n	g	$Z_0 J_n$	Z_{0e}	Z_{0o}
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 – FTB coupled line



ADS – FTB coupled line



Example

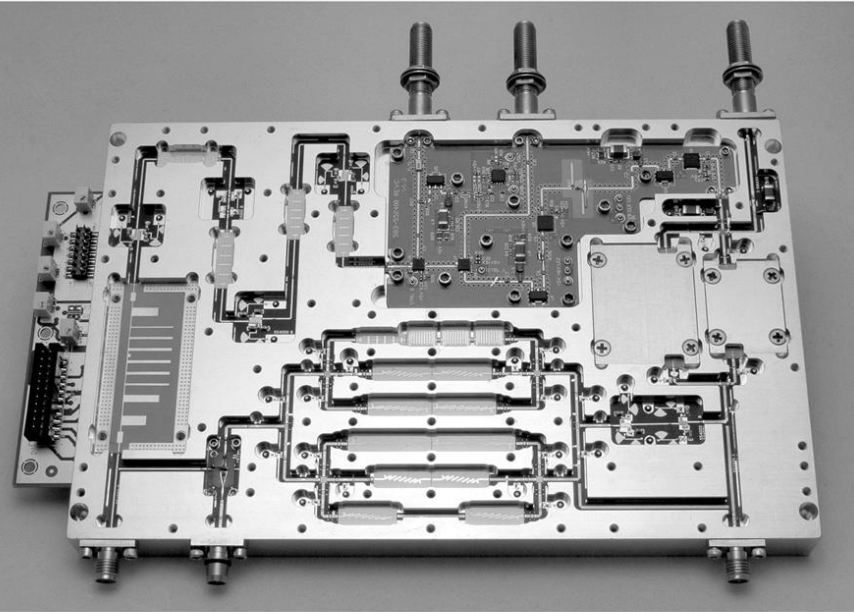
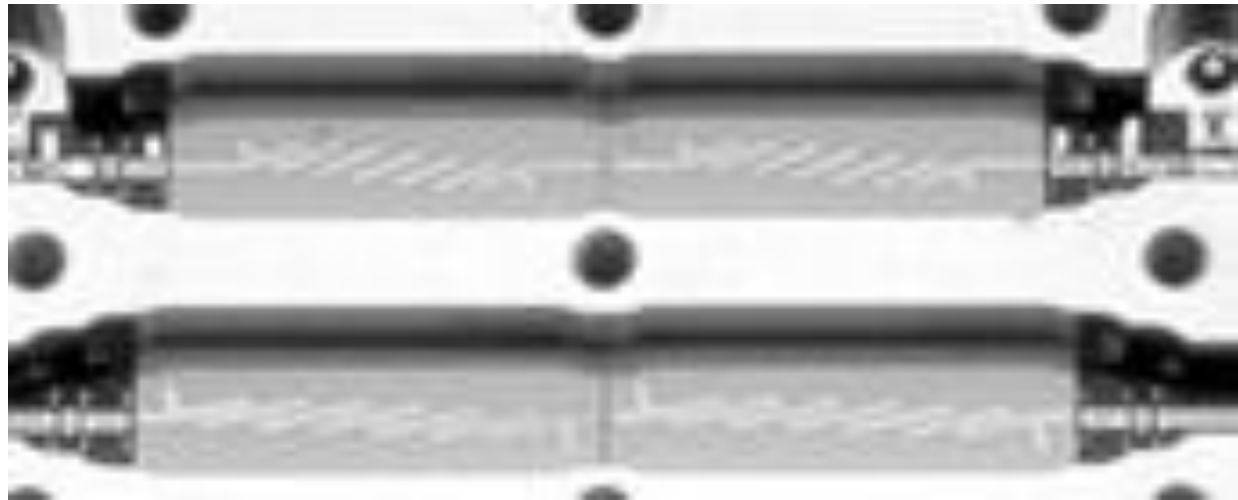
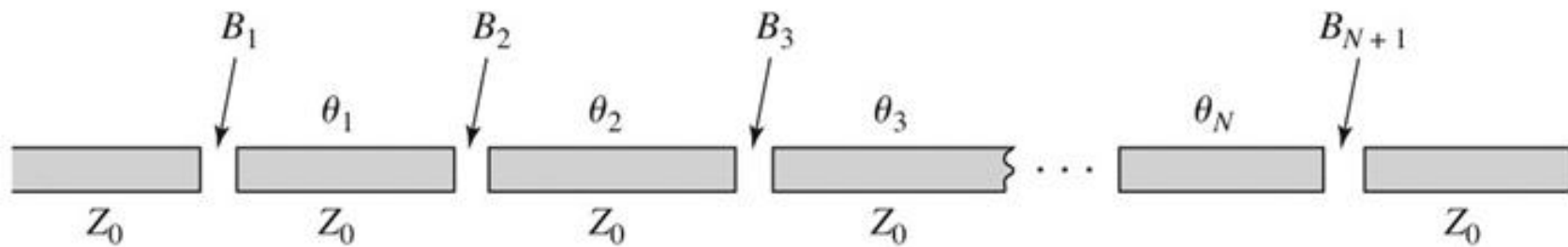


Figure 8.55
Courtesy of LNX Corporation, Salem, N.H.

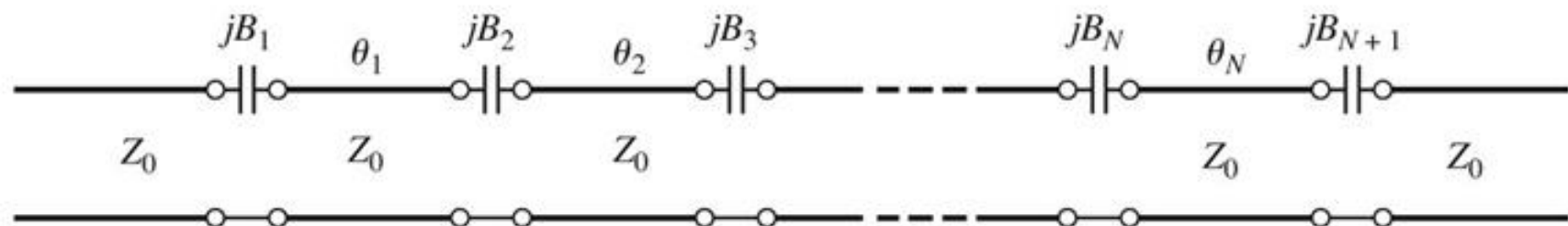


Filtru cu rezonatoare serie cuplate capacitiv

- Separarea fizica a doua sectiuni de linie produce un cuplaj capacitiv intre cele doua linii



(a)



(b)

Filtru cu rezonatoare serie cuplate capacitiv

- Din lungimea fizica a rezonatoarelor o portiune se foloseste pentru a crea schema de inversor ($\phi=\pi$, $l=\lambda/2$)

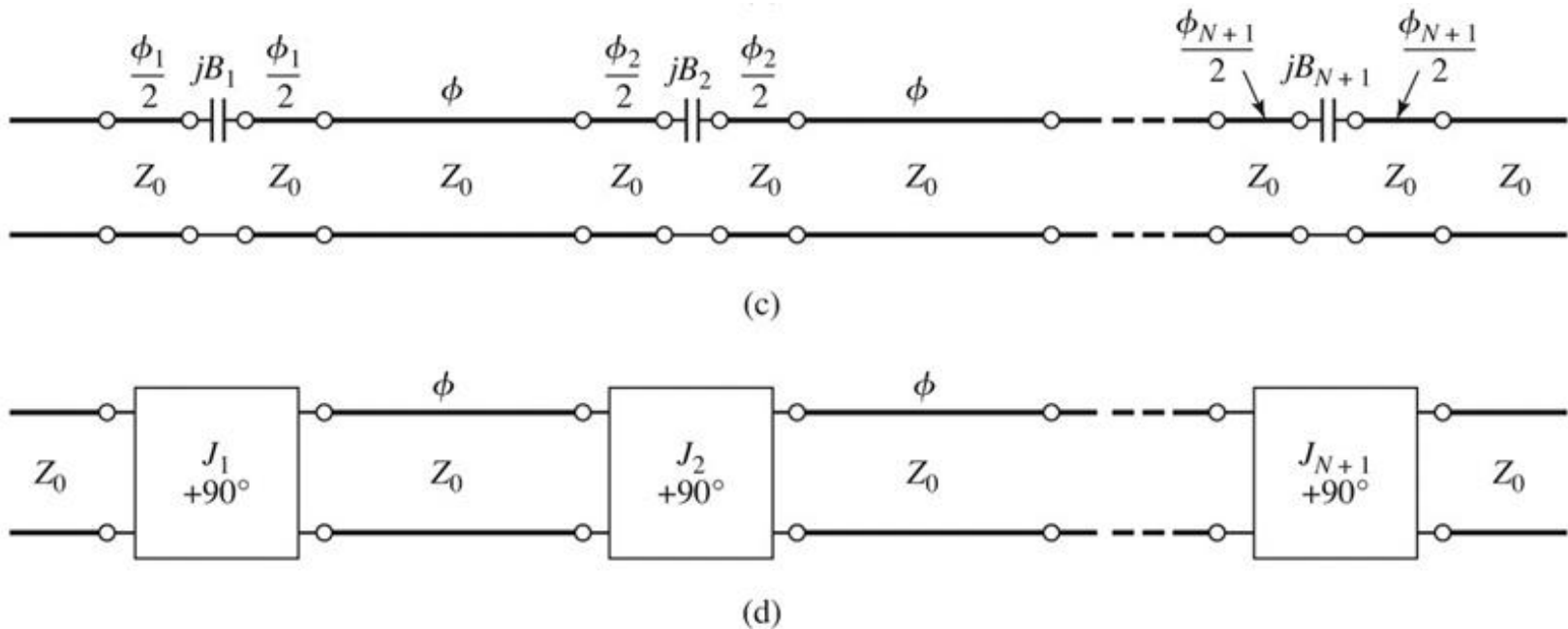


Figure 8.50

Proiectare filtru cu rezonatoare serie cuplate capacitiv

- Se calculeaza inversoarele (similar linii cuplate)

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

- Se calculeaza susceptantele cuplajului

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

- Se calculeaza lungimile de linii care trebuie “imprumutate” pentru realizarea inversoarelor

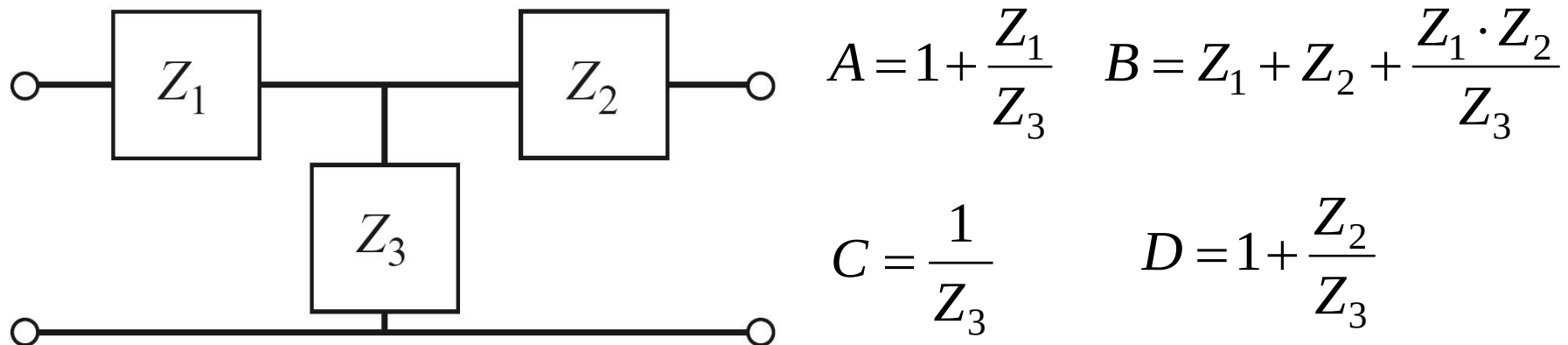
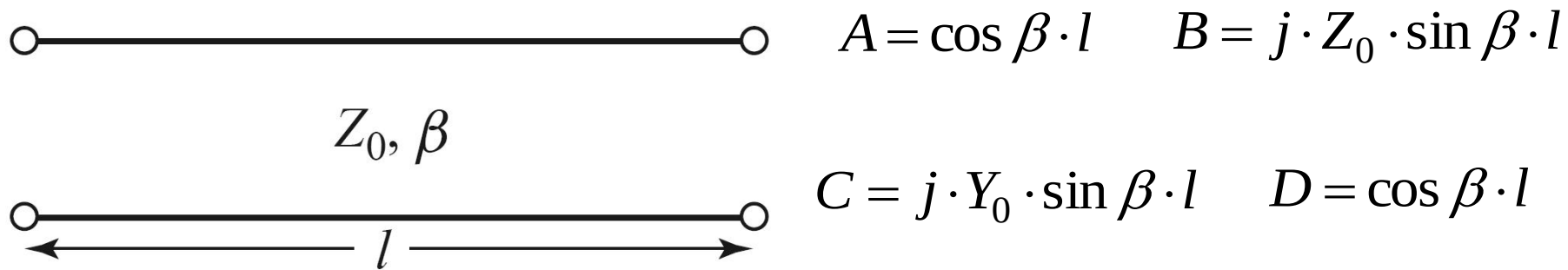
$$\phi_n = -\tan^{-1}(2 \cdot Z_0 \cdot B_n), n = \overline{1, N+1} \quad \phi_n < 0, n = \overline{1, N+1}$$

- Se calculeaza lungimile electrice ale liniilor

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

Circuite echivalente pentru sectiuni scurte de linii

- Parametri ABCD (C_4)
- linie scurta, model cu elemente concentrate valid



Circuite echivalente pentru sectiuni scurte de linii

- Element paralel capacitiv

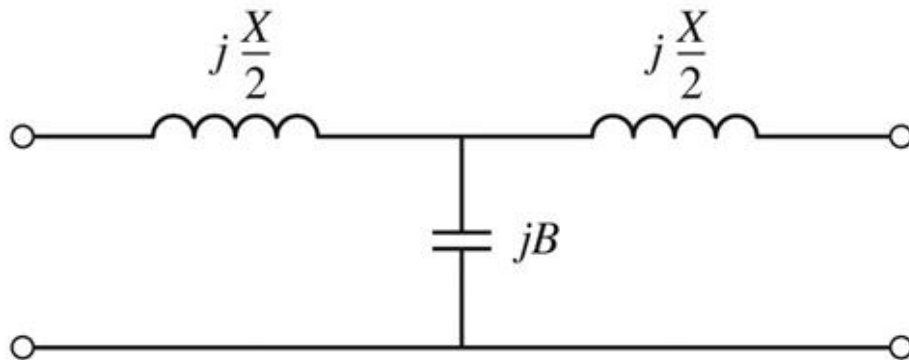
$$Z_3 = \frac{1}{j \cdot Y_0 \cdot \sin \beta \cdot l}$$

- Elementele in serie egale, inductive

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

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

- Schema echivalenta

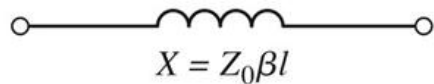


$$\frac{X}{2} = Z_0 \cdot \tan \frac{\beta \cdot l}{2}$$
$$B = \frac{1}{Z_0} \cdot \sin \beta \cdot l$$

Circuite echivalente pentru sectiuni scurte de linii

- In functie de valoarea impedantei caracteristice

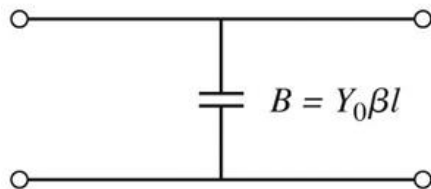
- impedanta ridicata $Z_0 \gg$



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



- impedanta scazuta $Z_0 \ll$



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

Filtre cu variatii treapta a impedantei

- Se pot crea filtre trece jos
- Se utilizeaza
 - linii cu impedanta caracteristica mare pentru a implementa o inductanta

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

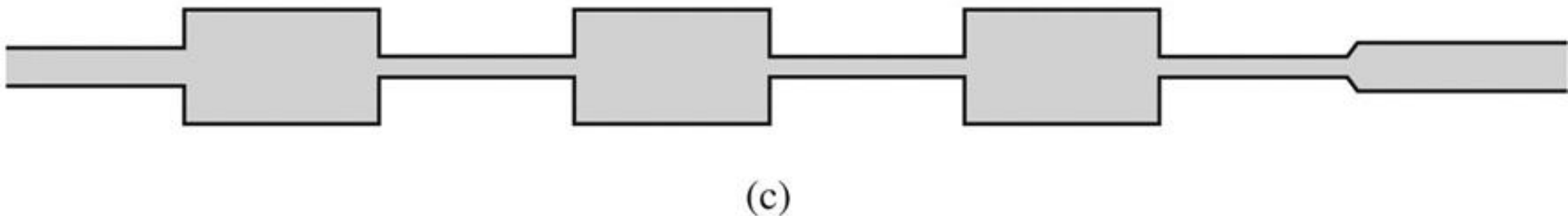
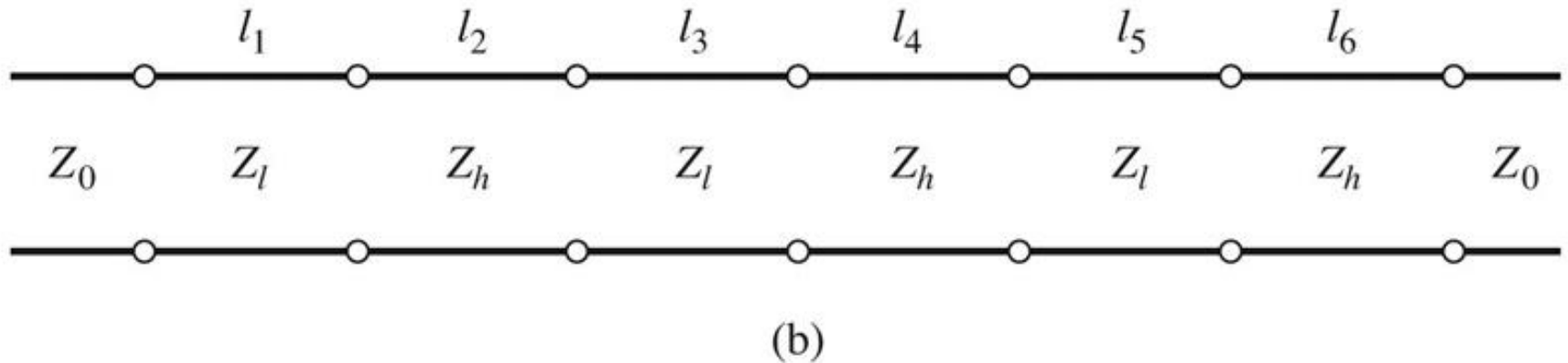
- linii cu impedanta caracteristica mica pentru a implementa o capacitate

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

- De obicei se utilizeaza cea mai mare si cea mai mica impedanta permisa de tehnologie

Filtre cu variatii treapta a impedantei

- Nu toate liniile au aceeași lungime deci problema periodicității în frecvență a răspunsului este mai puțin importantă

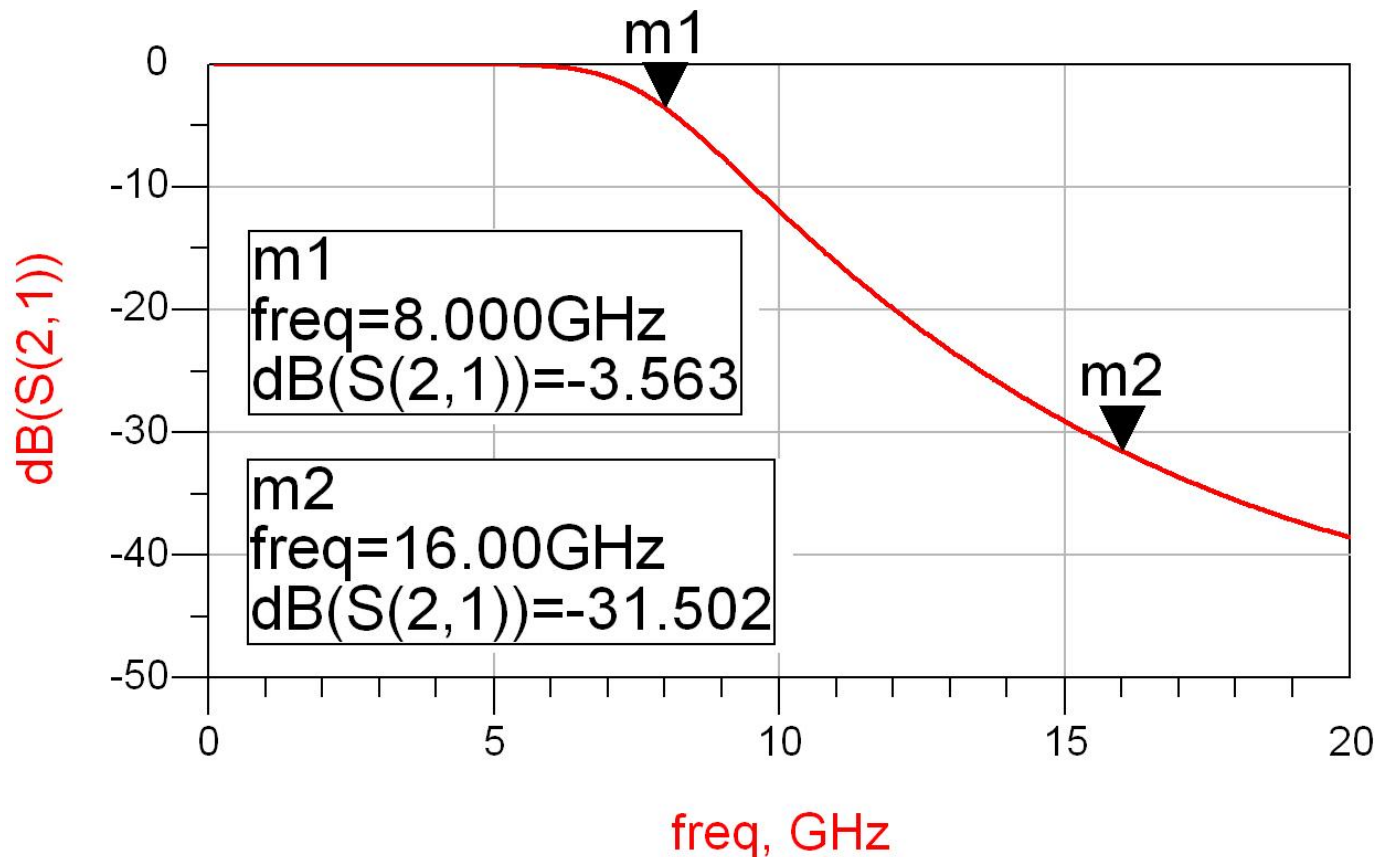
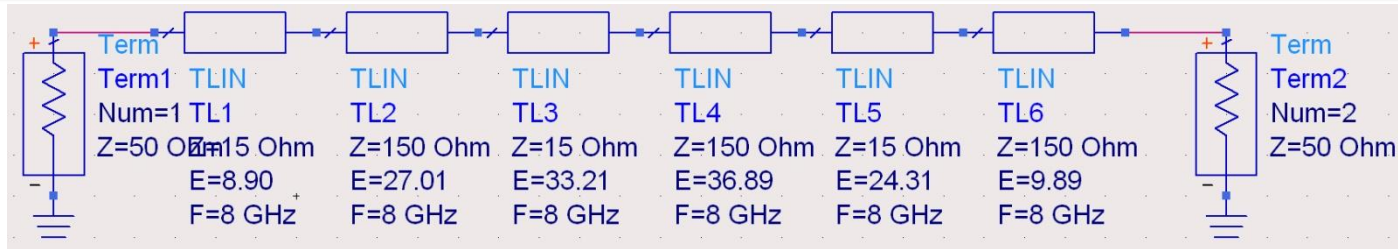


Exemplu

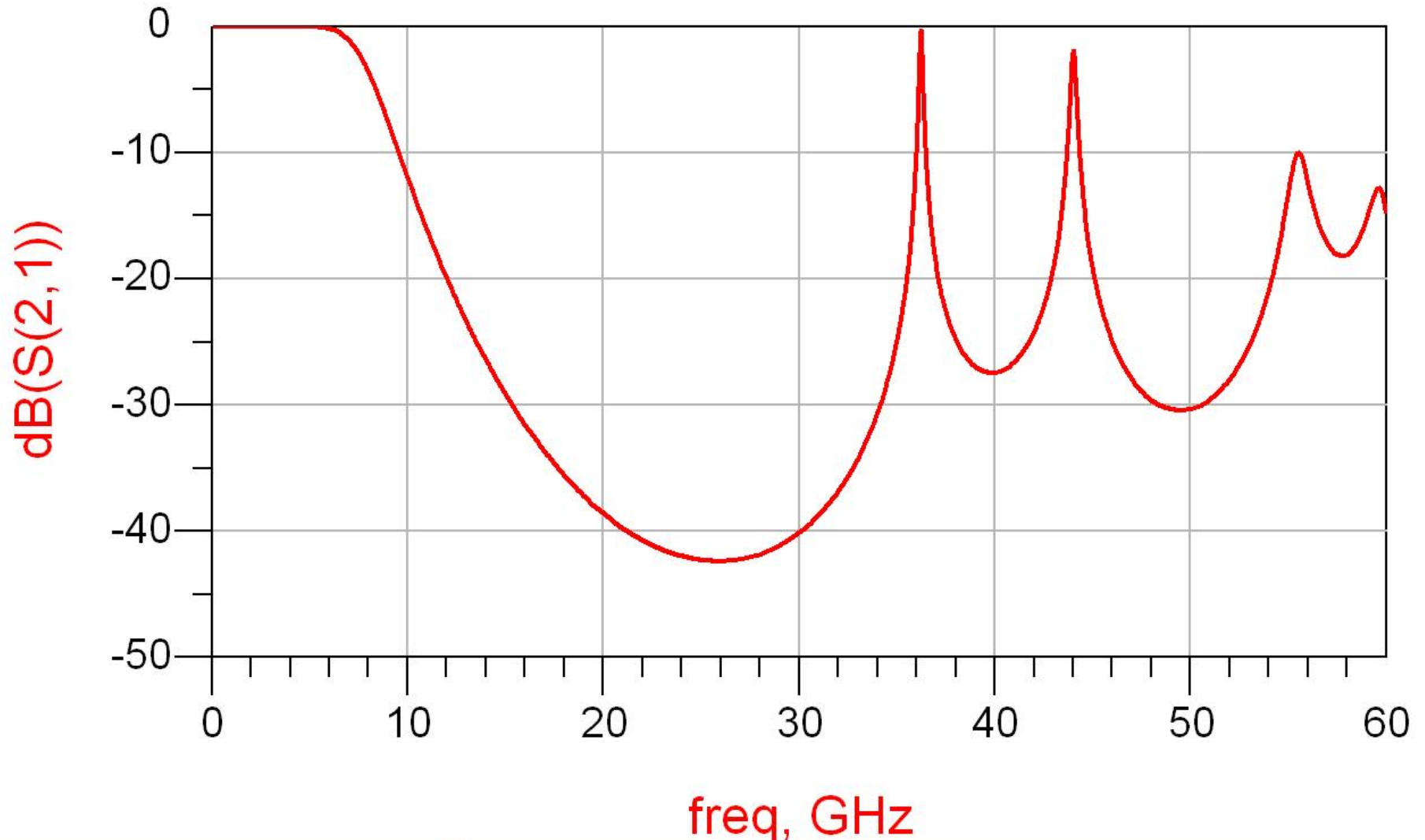
- FTJ cu frecventa de taiere 8GHz, de ordinul 6. Impedanta maxima realizabila este 150Ω iar cea minima 15Ω .

n	g_n	L/C_n	Z	$\theta_n[\text{rad}]$	$\theta_n[^\circ]$
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

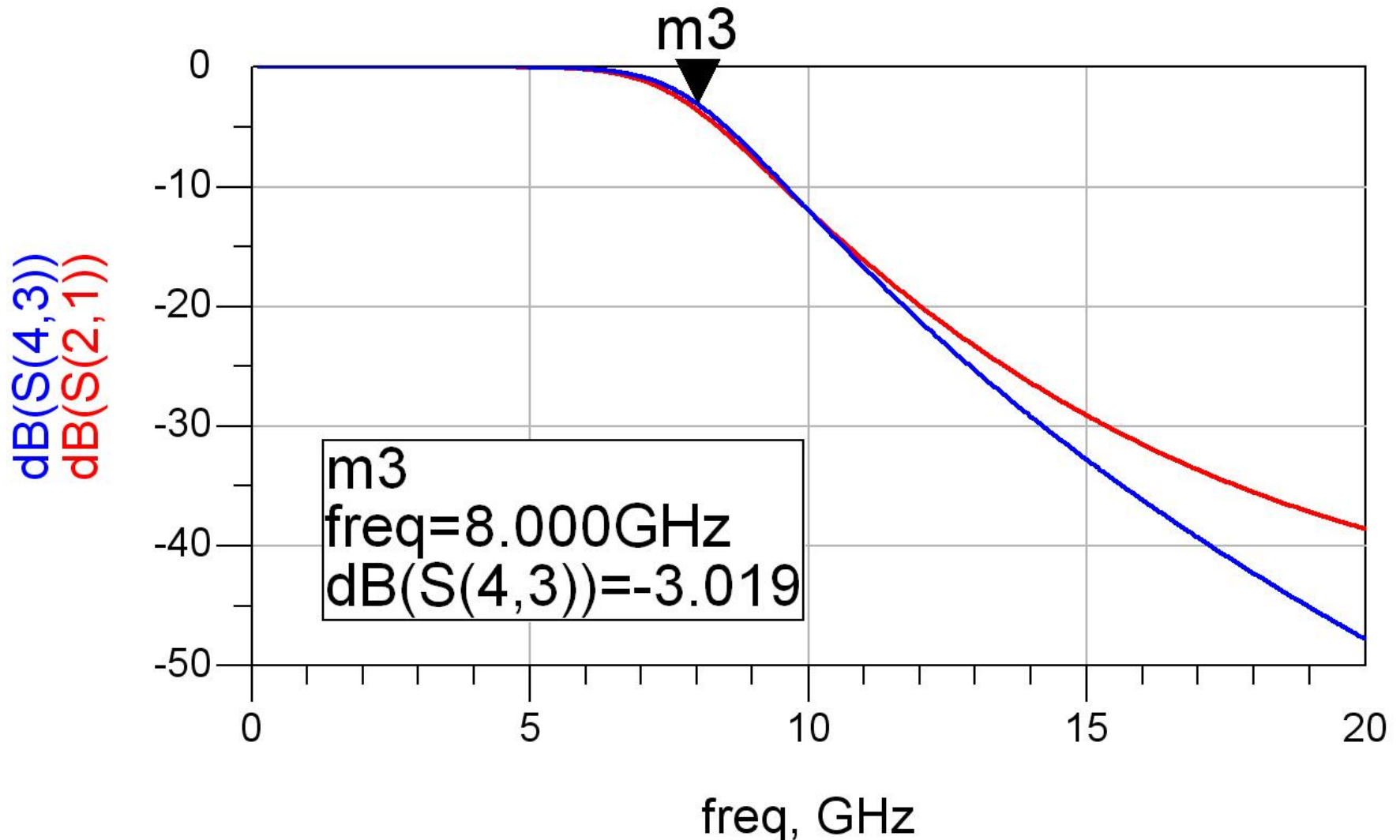
ADS – FTJ stepped impedance



ADS – FTJ stepped impedance



ADS – FTJ comparatie cu elemente concentrate



Example

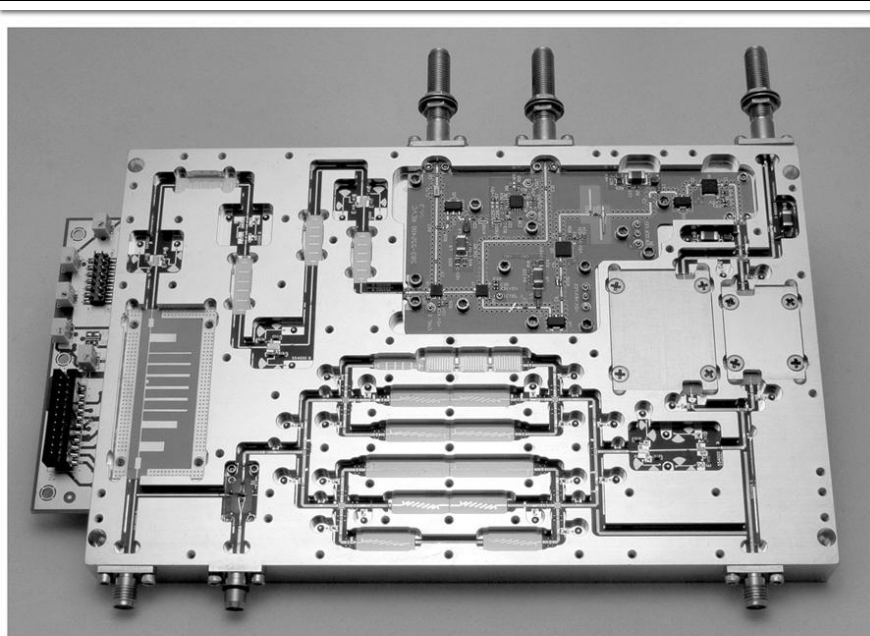
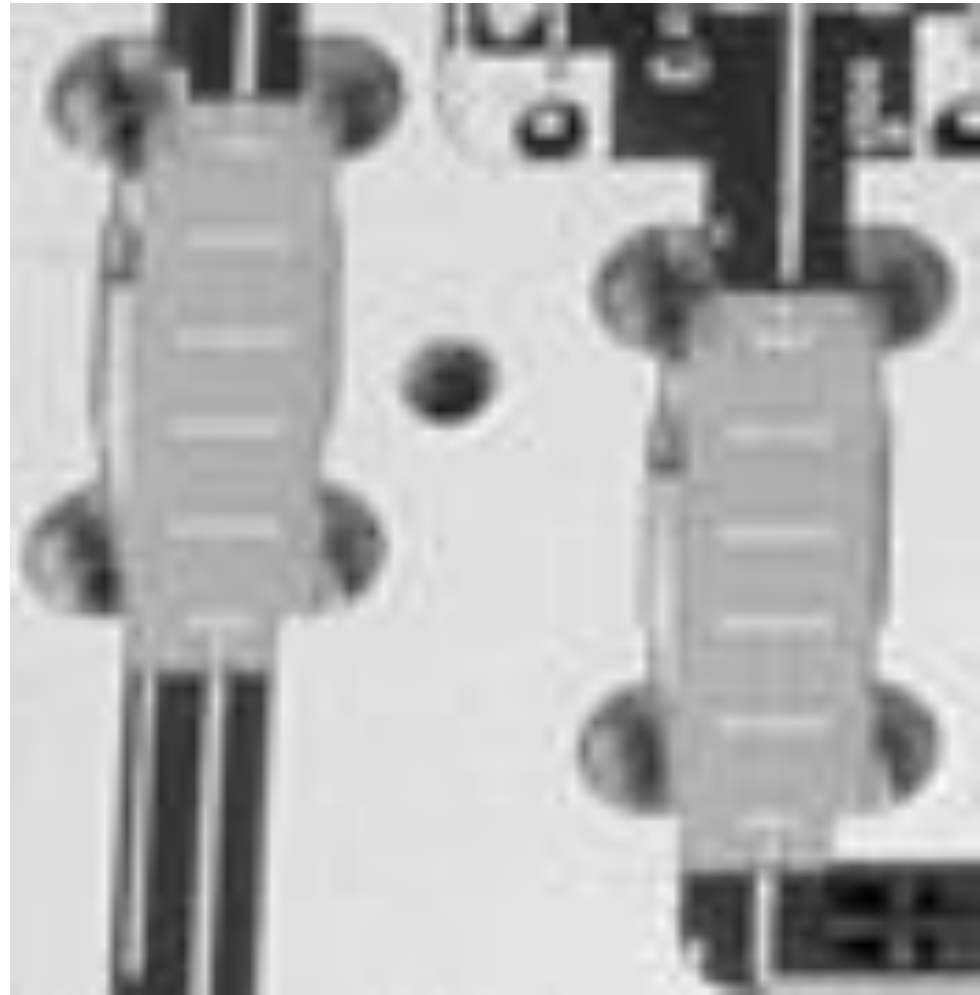


Figure 8.55
Courtesy of LNX Corporation, Salem, N.H.



Supliment Mini Proiect

Implementare cu linii microstrip

- linii microstrip
 - strat dielectric
 - metalizare totala (plan de masa)
 - trasee care fixeaza
 - impedanta caracteristica
 - lungime fizica/electrica

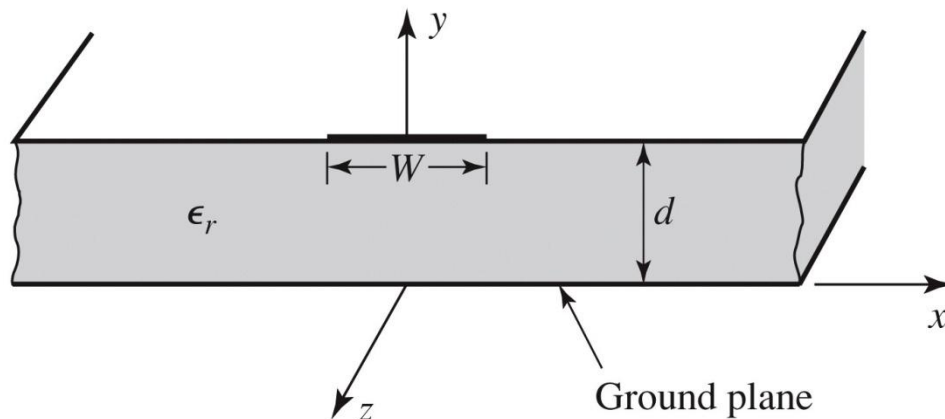


Figure 3.25a
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Implementare cu linii microstrip

- Linie quasi TEM

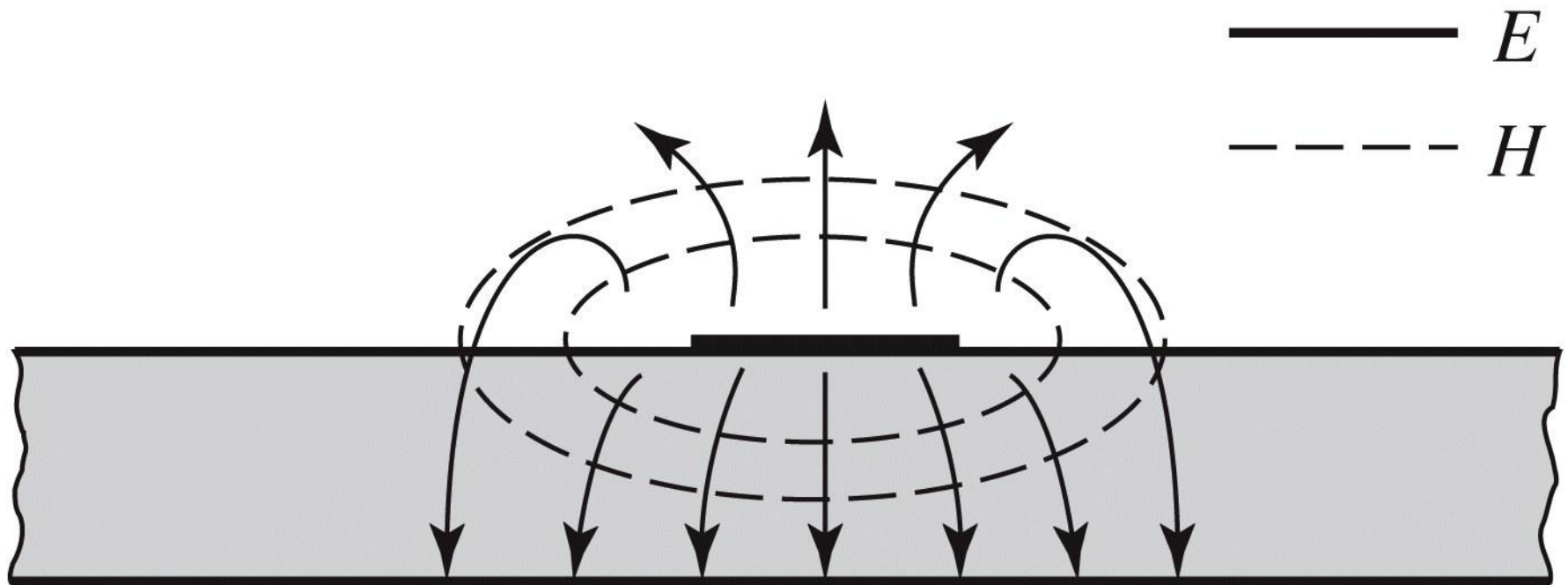
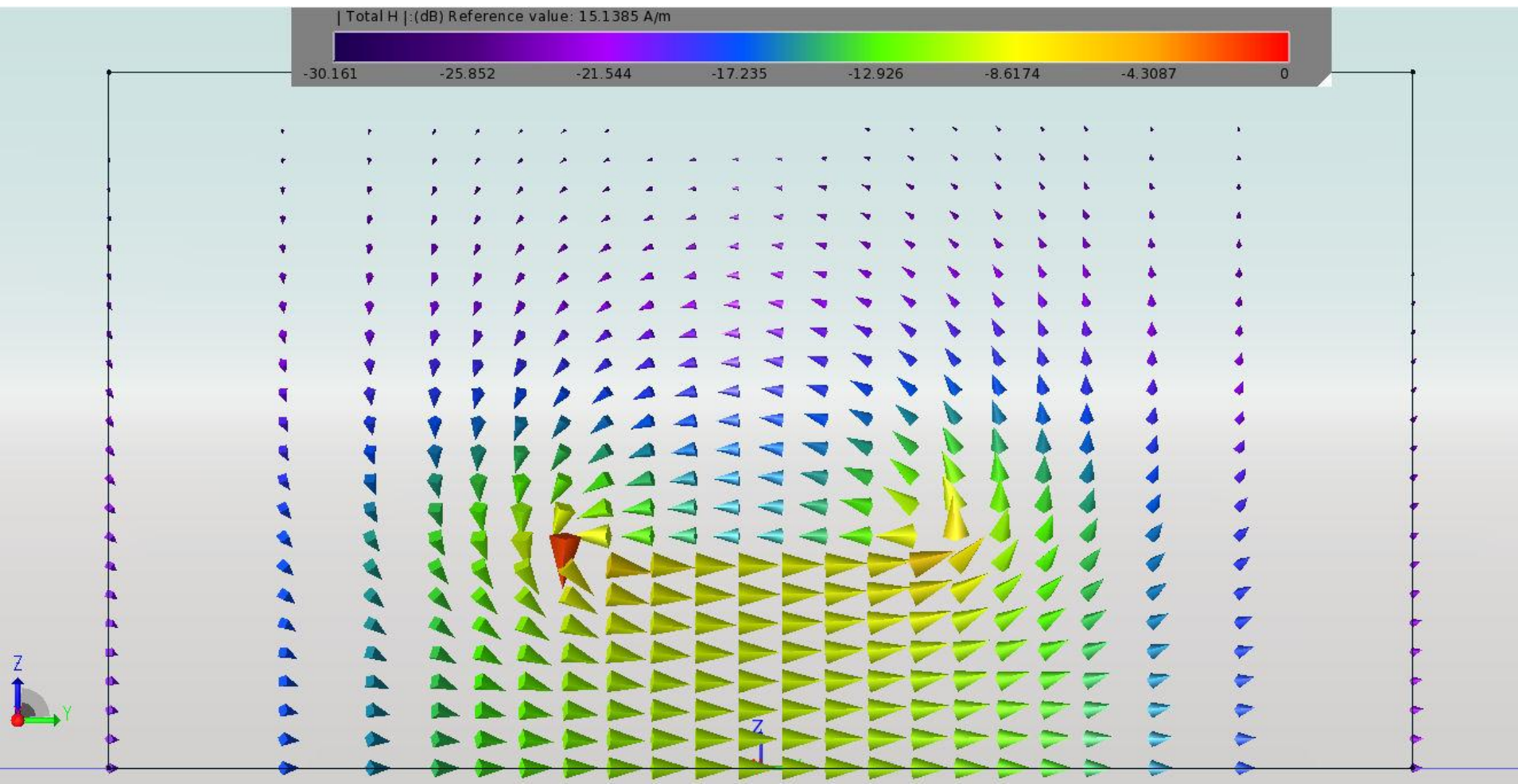


Figure 3.25b

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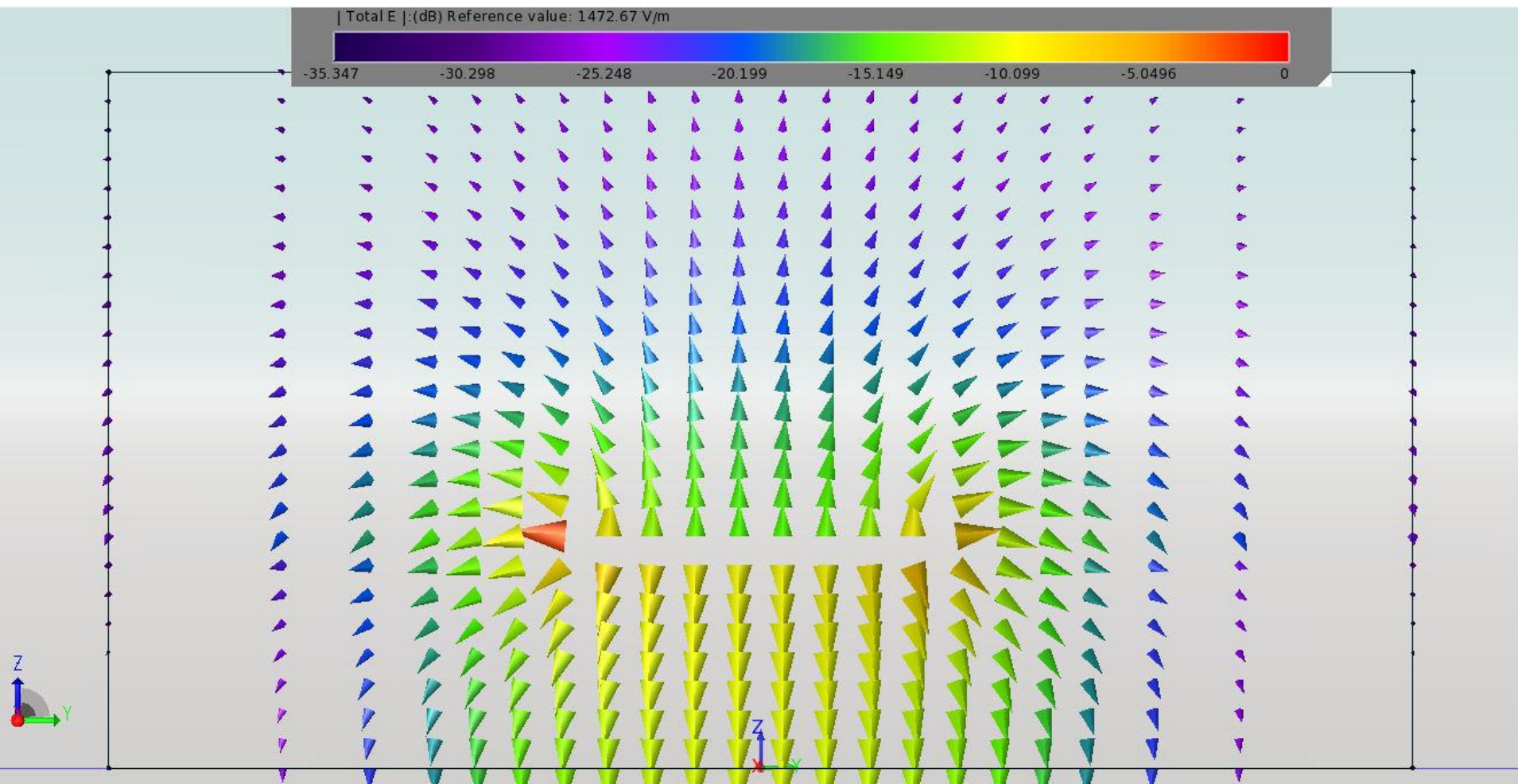
Implementare cu linii microstrip

- Linie quasi TEM, EmPro



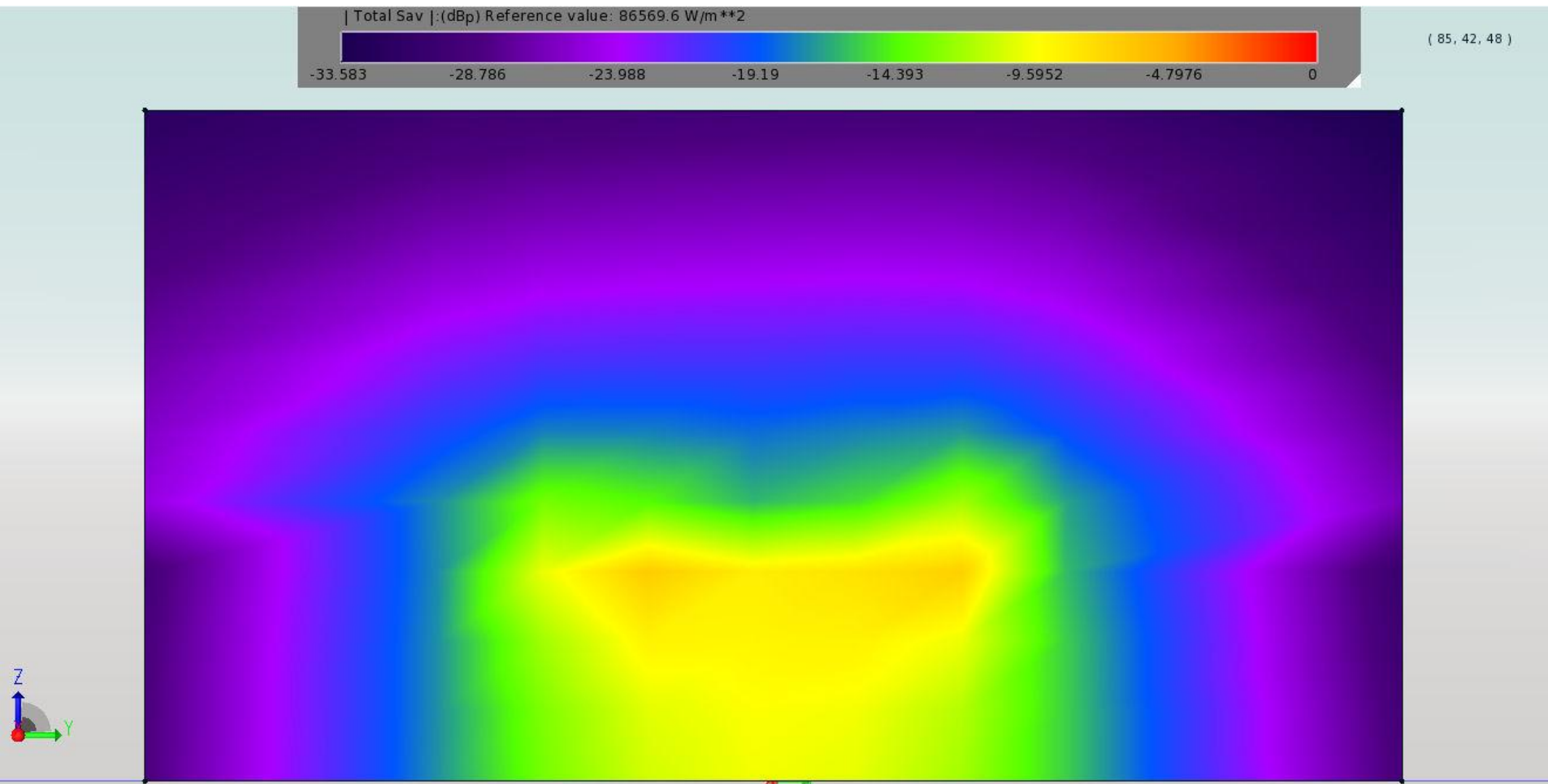
Implementare cu linii microstrip

- Linie quasi TEM, EmPro



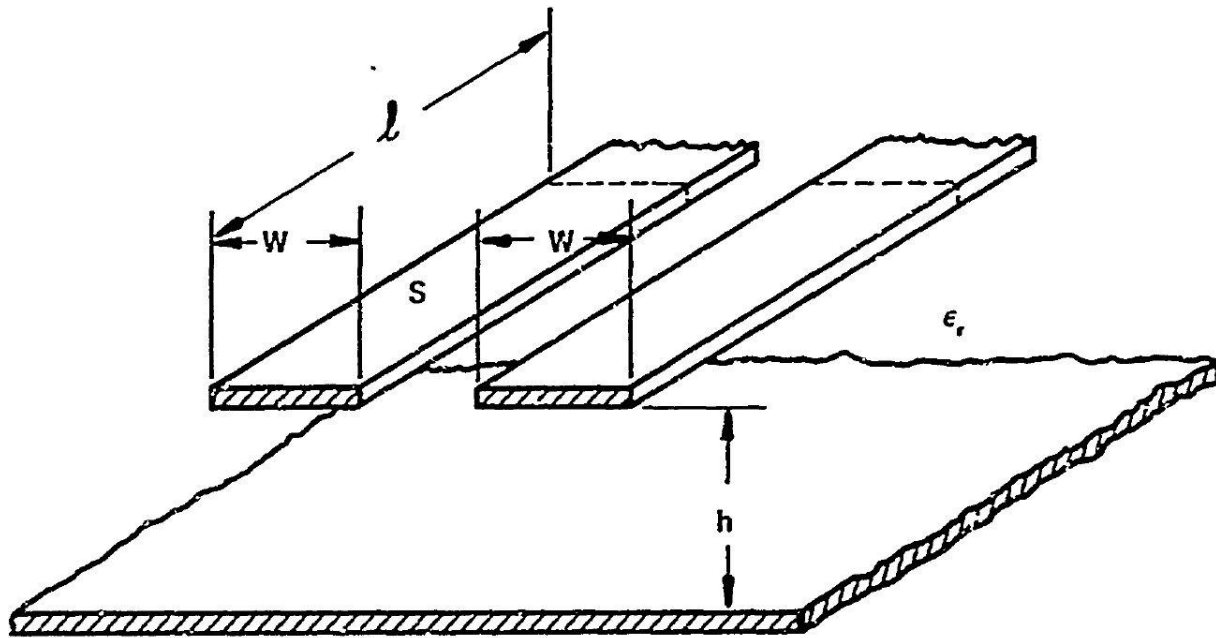
Implementare cu linii microstrip

- Linie quasi TEM, EmPro



Implementare cu linii microstrip

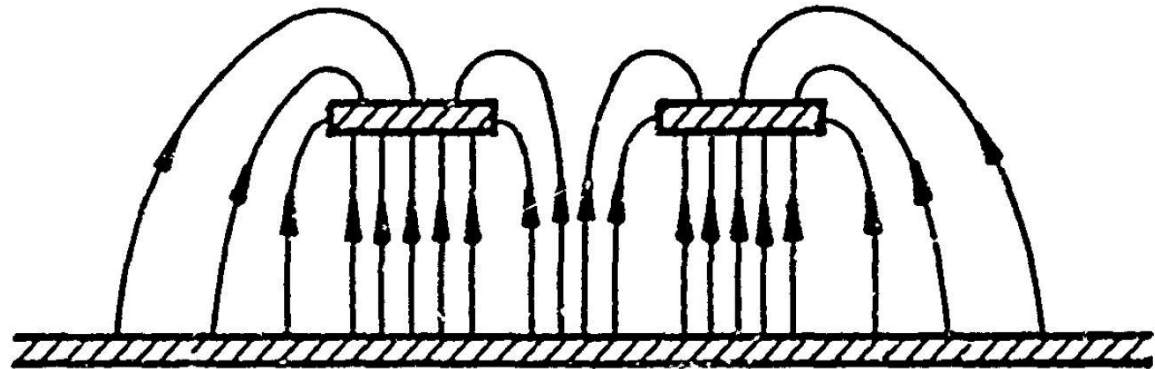
- ~ Aproximativ TEM



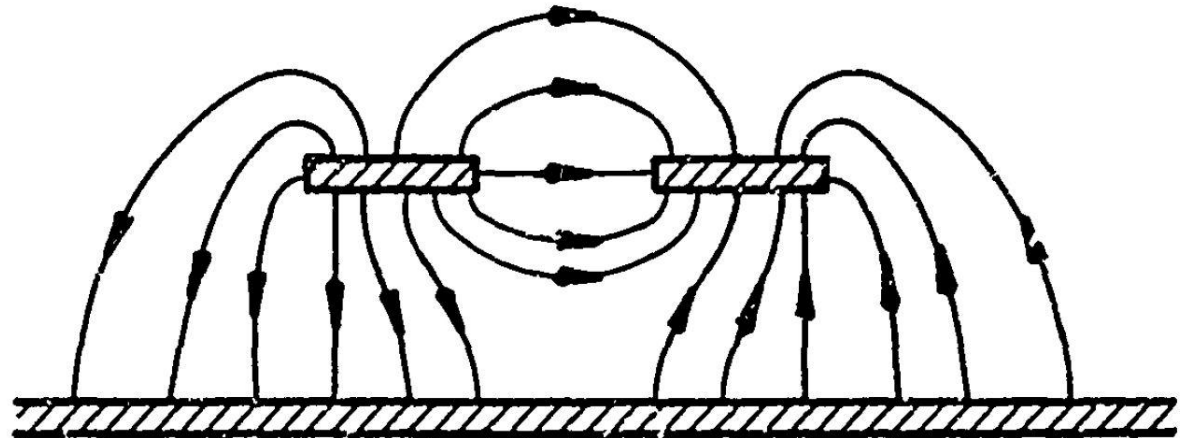
a) COUPLED STRIP GEOMETRY

Implementare cu linii microstrip

- ~ Aproximativ TEM



b) EVEN MODE ELECTRIC FIELD PATTERN (SCHEMATIC)



c) ODD MODE ELECTRIC FIELD PATTERN (SCHEMATIC)

Implementare cu linii microstrip

- Se echivaleaza linia cu o linie cu dielectric omogen echivalent

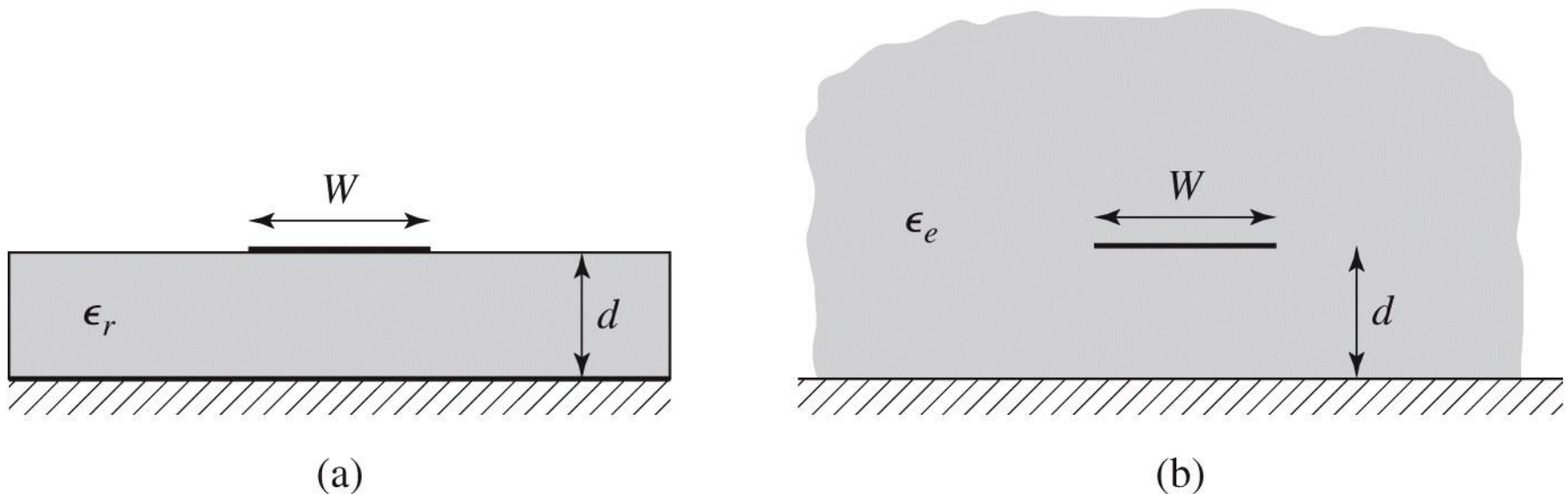


Figure 3.26
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Calcul

■ Calcul empiric

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

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

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

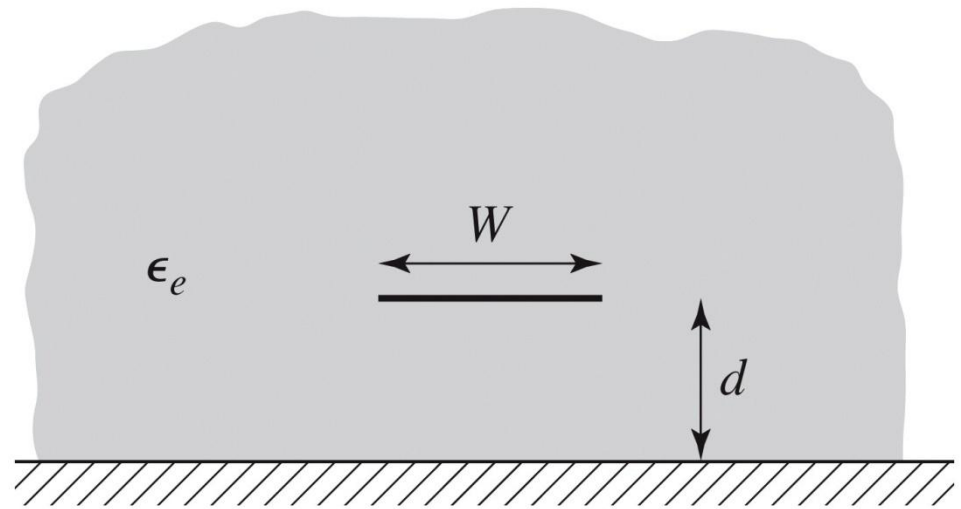


Figure 3.26b
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$$Z_0 = \begin{cases} \frac{60}{\sqrt{\epsilon_e}} \ln \left(\frac{8d}{W} + \frac{W}{4d} \right) & \text{for } W/d \leq 1 \\ \frac{120\pi}{\sqrt{\epsilon_e} [W/d + 1.393 + 0.667 \ln (W/d + 1.444)]} & \text{for } W/d \geq 1. \end{cases}$$

Calcul

■ Calcul empiric

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

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

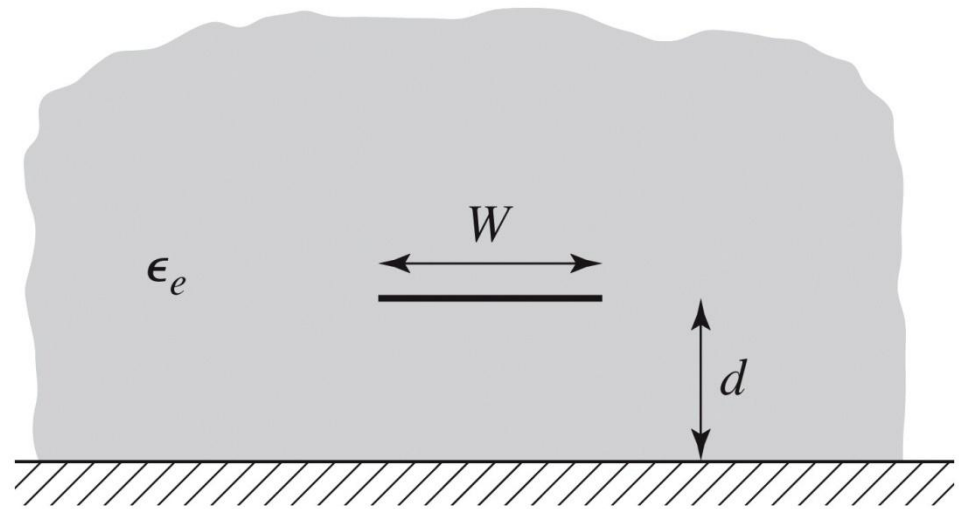
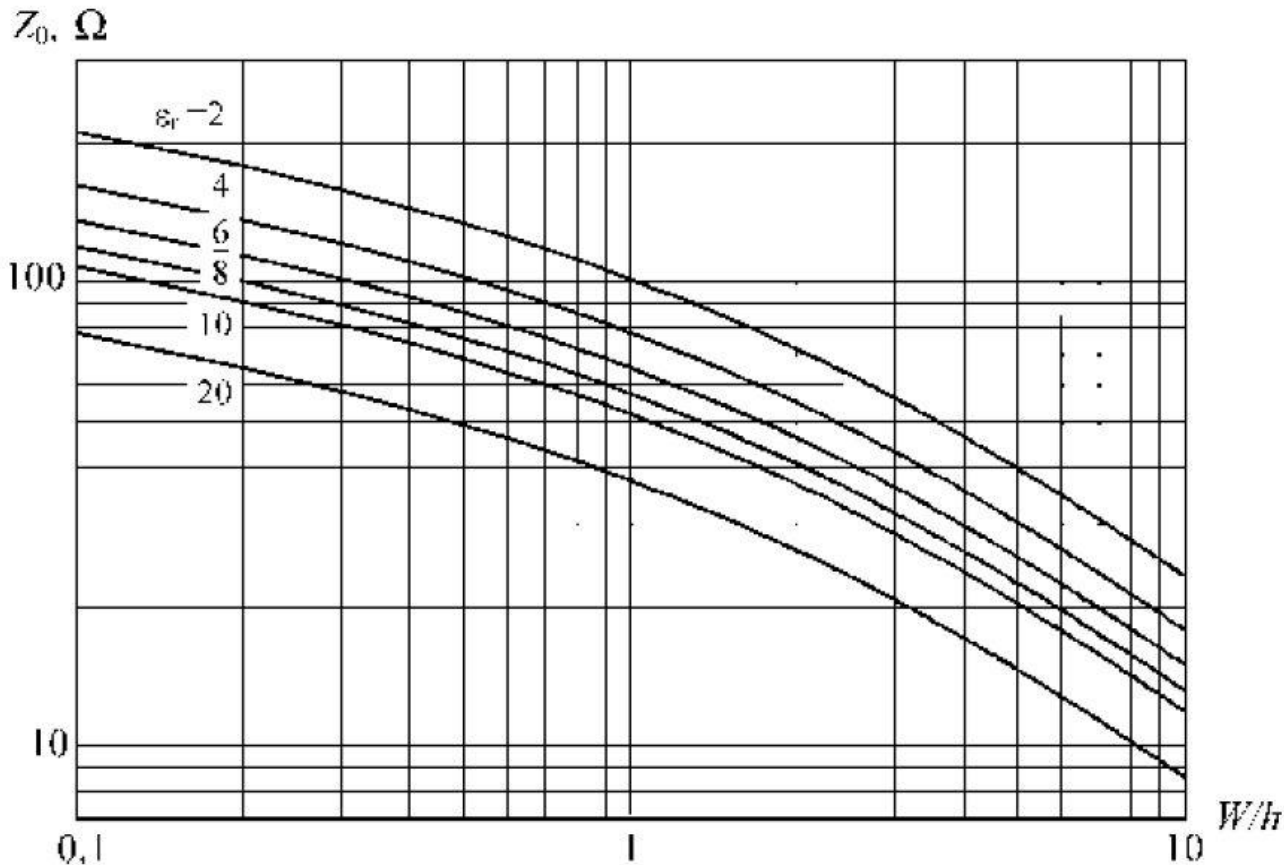


Figure 3.26b
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$$\frac{W}{d} = \begin{cases} \frac{8e^A}{e^{2A} - 2} & \text{for } W/d < 2 \\ \frac{2}{\pi} \left[B - 1 - \ln(2B - 1) + \frac{\epsilon_r - 1}{2\epsilon_r} \left\{ \ln(B - 1) + 0.39 - \frac{0.61}{\epsilon_r} \right\} \right] & \text{for } W/d > 2, \end{cases}$$

Impedanta caracteristica

- Pentru **impedante mari** e nevoie de **latimi mici** ale traseelor
- Pentru **impedante mici** e nevoie de **latimi mari** ale traseelor



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

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

Microstrip standardizare

- Standardizare
 - dimensiuni in **mil**
 - 1 mil = 10^{-3} inch
 - 1 inch = 2.54 cm
- Inaltimea conductoarelor
 - in functie de greutatea cuprului
 - uncii / picioare pătrate (oz/ft²)
 - 10z=28.35g și 1ft=30.48cm

Greutatea cuprului depus		Grosimea stratului	
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 standardizare

- Tipic inaltimea straturilor de dielectric de asemenea standardizat in mil

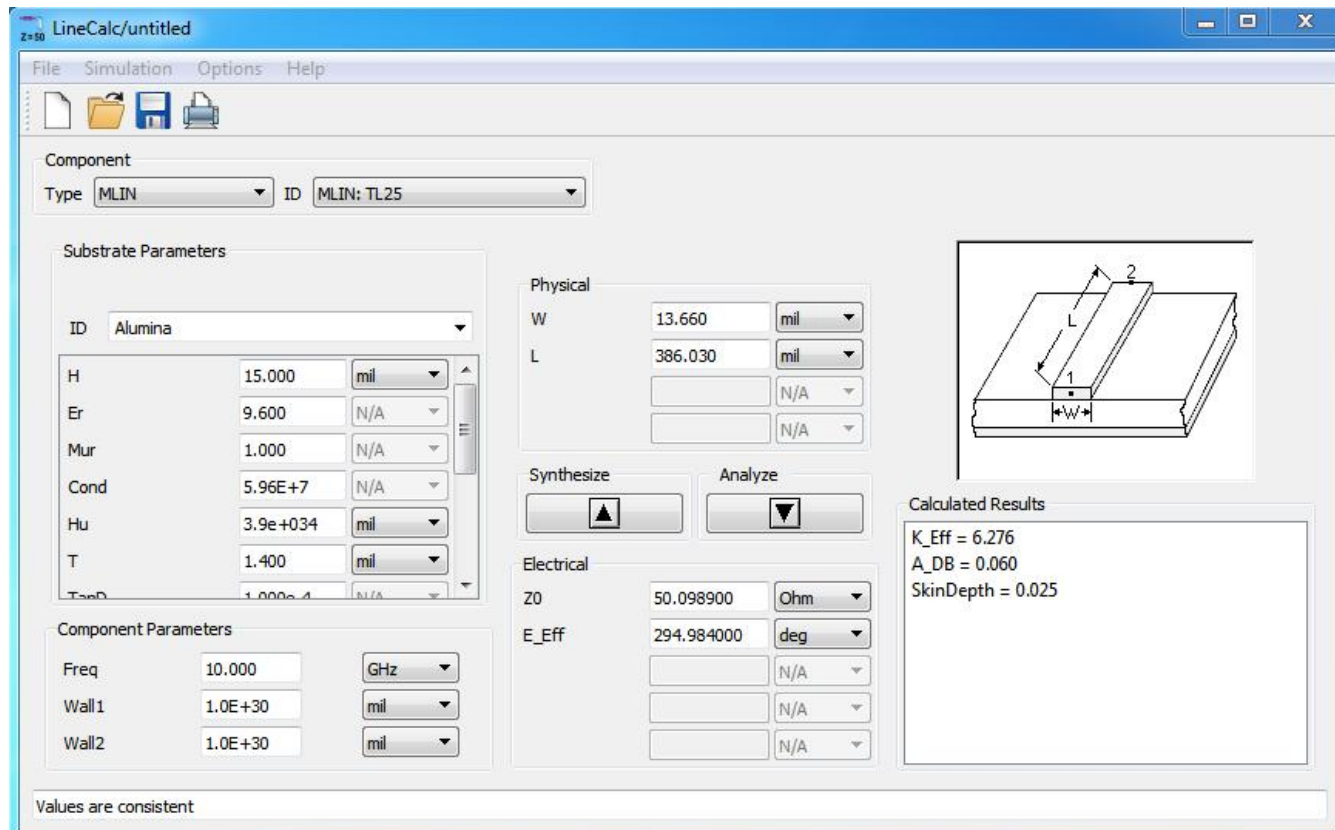
Standard Thickness
RO4003C: 0.008" (0.203mm), 0.012 (0.305mm), 0.016" (0.406mm), 0.020" (0.508mm) 0.032" (0.813mm), 0.060" (1.524mm) RO4350B: *0.004" (0.101mm), 0.0066" (0.168mm) 0.010" (0.254mm), 0.0133 (0.338mm), 0.0166 (0.422mm), 0.020" (0.508mm) 0.030" (0.762mm), 0.060" (1.524mm)

Microstrip, materiale substrat

	Constanta dielectrică relativă	Factorul de pierderi dielectrice	Conductivitate termică	Coeficient liniar de expansiune	Coeficient de temperatură a lui ϵ_r
Material	-	-	W/cm/K	ppm/K	ppm/K
Al_2O_3 (99.5%)	9.8	0.0001	0.37	6.3	+136
Al_2O_3 (96%)	9.4	0.001	0.35	6.4	-
Safir	9.4;11.6	0.0001	0.42	6.0	+110-+140
Sticlă cuarț	3.78	0.0001	0.017	0.55	+13
Sticlă Corning 7059	5.75	0.0036	0.012	4.6	-
BeO Ceramic (98%)	6.3	0.006	2.1	6.1	+107
TiO_2	85	0.004	0.05	7.5	-575
Tetratitanat de Ba (BaTi_4O_9)	37	0.0005	0.02	9.4	-26
Zirconat	20-40	0.002	-	5.0	-130-+100
GaAs	12.9	0.002	0.46	5.7	-
Si	11.9	0.015	1.45	4.2	-
Ferită	9-16	0.001	-	-	-

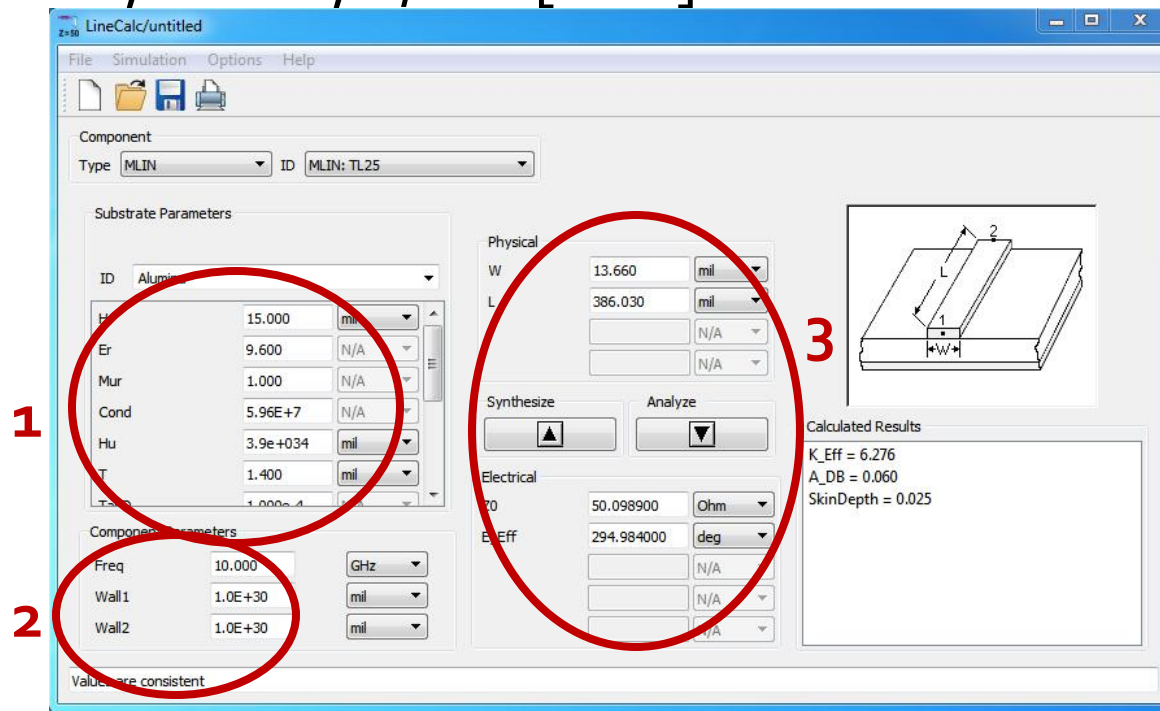
ADS linecalc

- In scheme:
 - >Tools>LineCalc>Start
 - Pentru linii Microstrip >Tools>LineCalc>Send to Linecalc



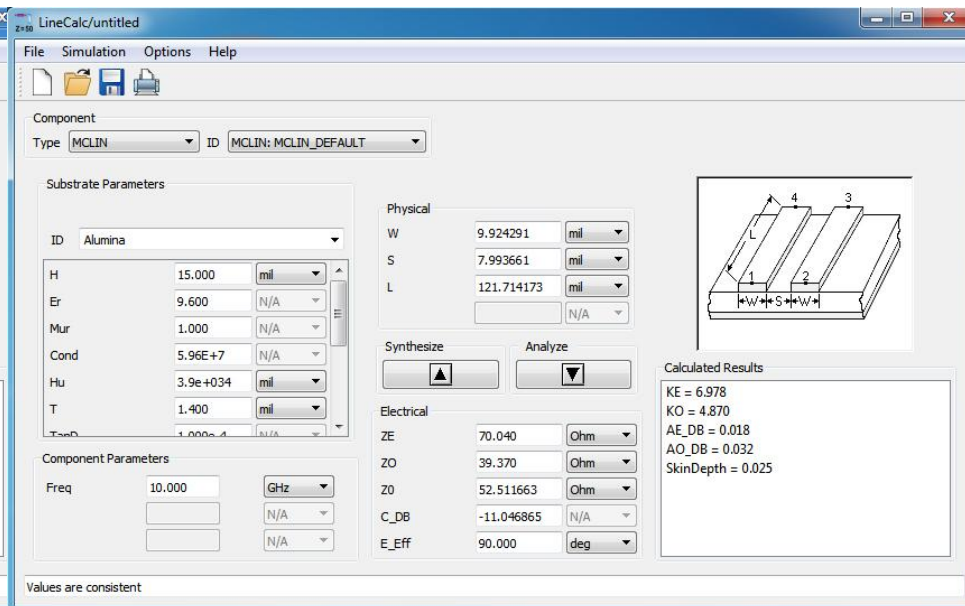
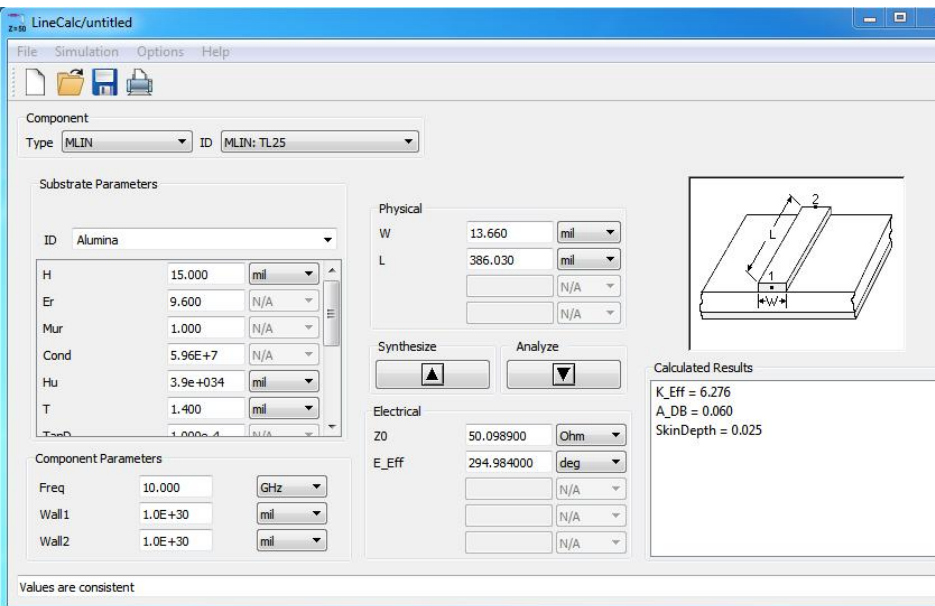
ADS linecalc

- 1. Definire (receptie din schema) substrat
- 2. Introducere frecventa
- 3. Introducere date de intrare
 - Analiza: $W, L \rightarrow Z_0, E$ sau $Z_e, Z_0, E / la f [GHz]$
 - Sinteza: $Z_0, E \rightarrow W, L / la f [GHz]$



ADS linecalc

- Se poate utiliza pentru:
 - linii microstrip MLIN: $W, L \Leftrightarrow Z_0, E$
 - linii cuplate microstrip MCLIN: $W, L \Leftrightarrow Z_e, Z_0, E$



ADS linecalc

LineCalc/untitled

File Simulation Options Help

Component
Type: MCLIN ID: MCLIN: MCLIN_DEFAULT

Substrate Parameters

ID	Alumina
H	15.000 mil
Er	9.600 N/A
Mur	1.000 N/A
Cond	5.96E+7 N/A
Hu	3.9e+034 mil
T	1.400 mil
TanD	1.000e-4 N/A

Component Parameters

Freq	10.000 GHz
	N/A
	N/A

Physical

W	9.924291 mil
S	7.993661 mil
L	121.714173 mil
	N/A

Synthesize Analyze

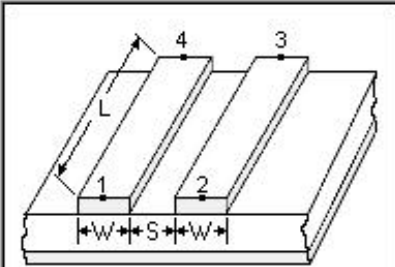
Electrical

ZE	70.040 Ohm
ZO	39.370 Ohm
Z0	52.511663 Ohm
C_DB	-11.046865 N/A
E_Eff	90.000 deg

Calculated Results

KE = 6.978
KO = 4.870
AE_DB = 0.018
AO_DB = 0.032
SkinDepth = 0.025

Values are consistent



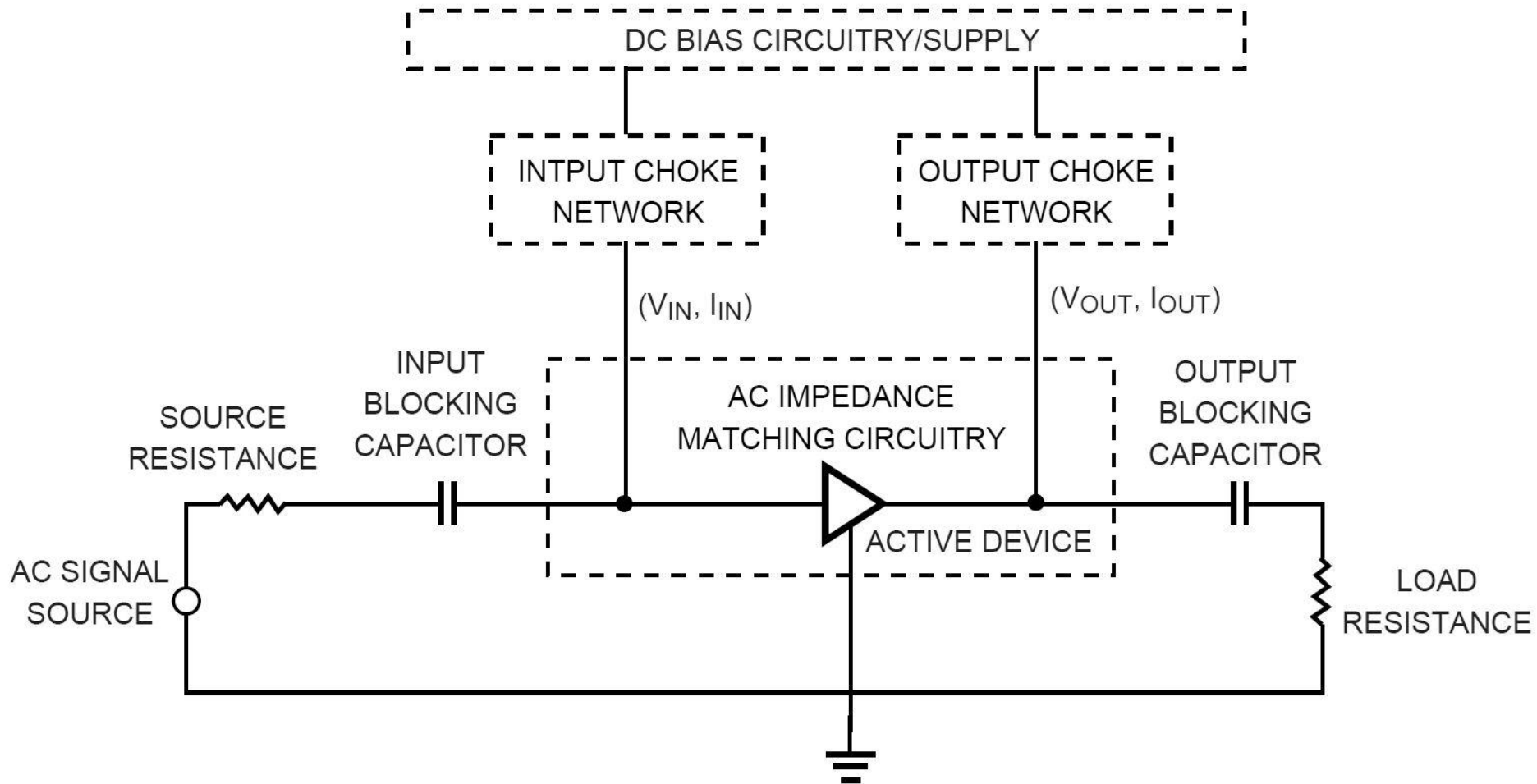
Linii de transmisie

- <http://rf-opto.etti.tuiasi.ro>
- linii de transmisie Rogers
 - relatii dependente de
 - t , inaltimea metalizarilor
 - f , frecventa
 - relatii pentru
 - microstrip
 - strip
 - linii cuplate

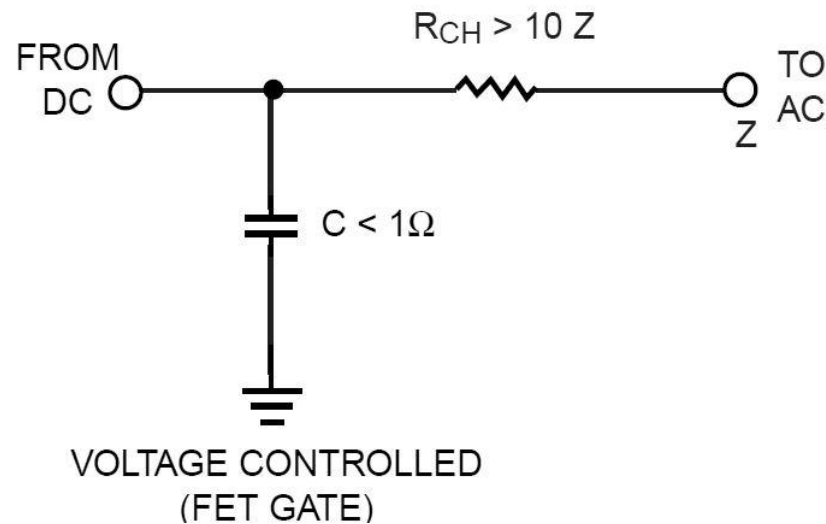
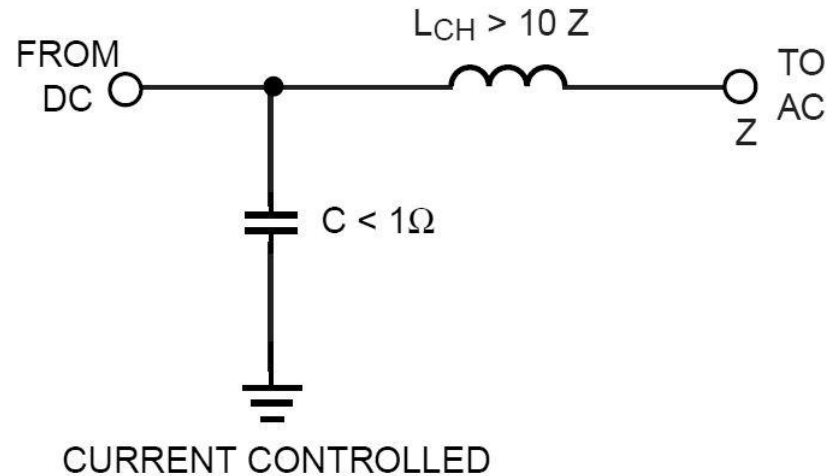
Polarizare

- <http://rf-opto.etti.tuiasi.ro>
- note de aplicatii importante Agilent
 - decuplarea circuit de semnal/circuit de polarizare
 - detalii de implementare a circuitelor de polarizare pentru tranzistoarele cu microunde
- Appcad contine instrumente pentru calculul schemelor de polarizare

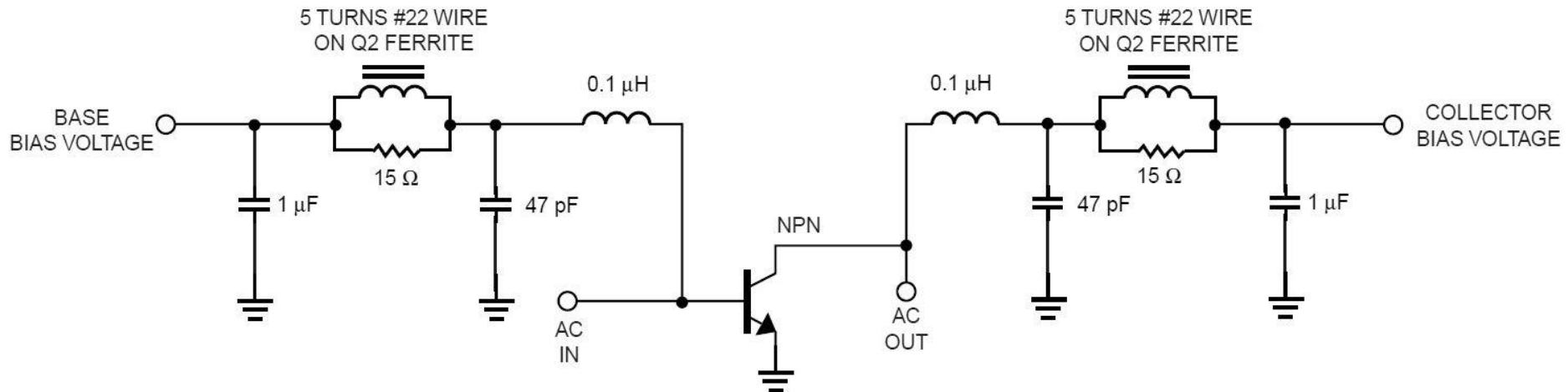
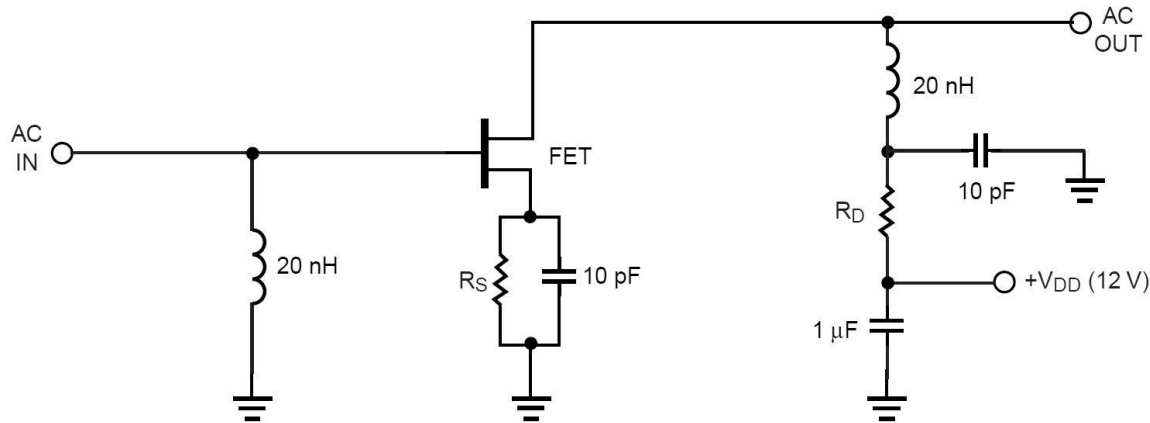
Polarizare



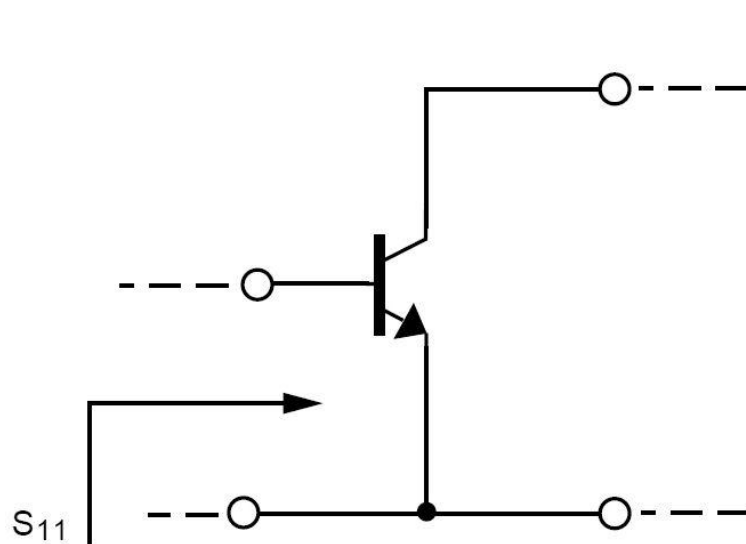
Polarizare, soc tipic



Polarizare, scheme/valori tipice

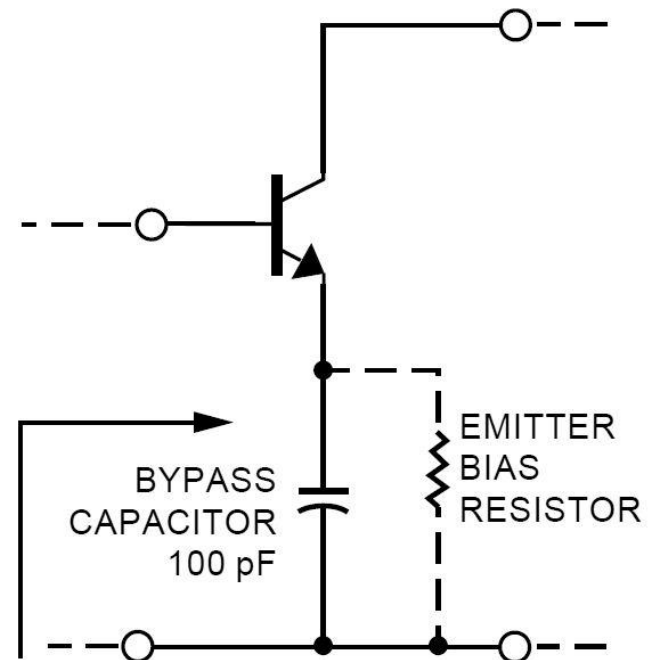


Polarizare, problema elementelor conectate in E/S



$$S_{11} \text{ (AT 4 GHz)} = 0.52 \angle 154^\circ$$

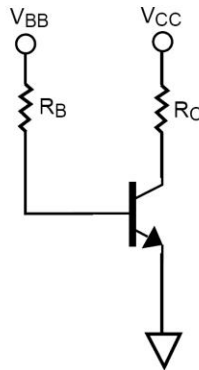
$$S_{11} \text{ (AT 0.1 GHz)} = 0.901 \angle -14.9^\circ$$



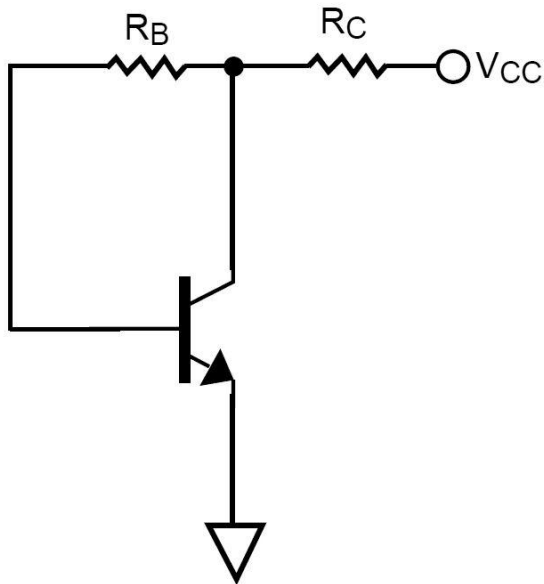
$$S'_{11} \text{ (AT 4 GHz)} = 0.52 \angle 154^\circ \text{ UNCHANGED AT 4 GHz}$$

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

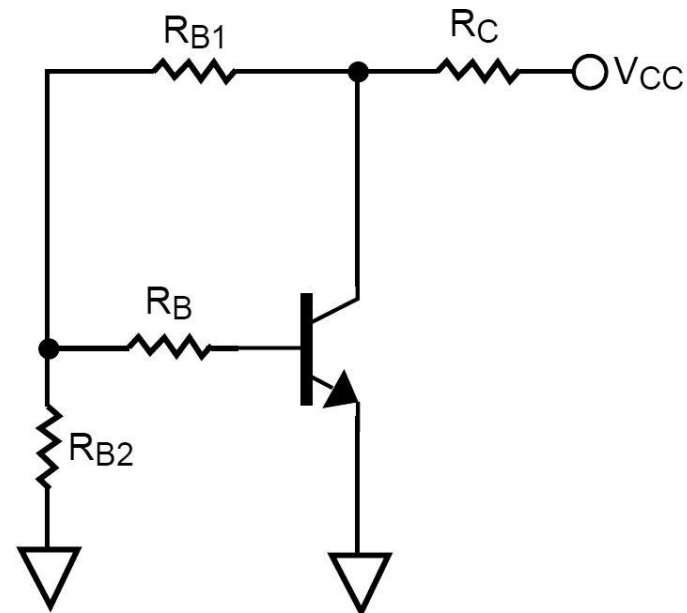
Polarizare TB scheme de polarizare



NON-STABILIZED



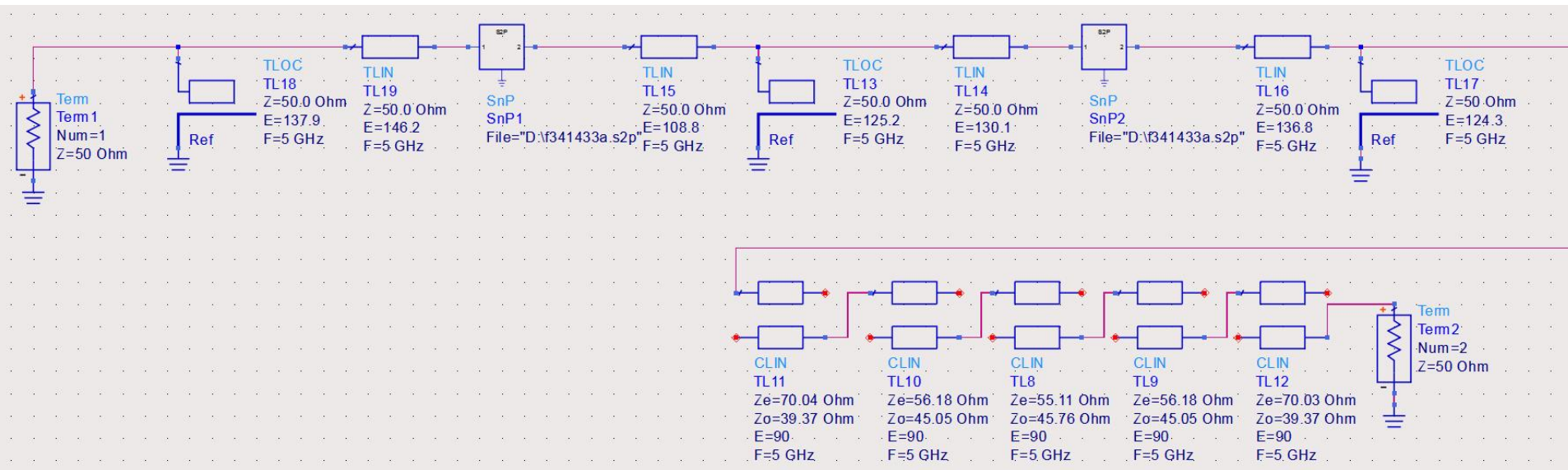
VOLTAGE FEEDBACK



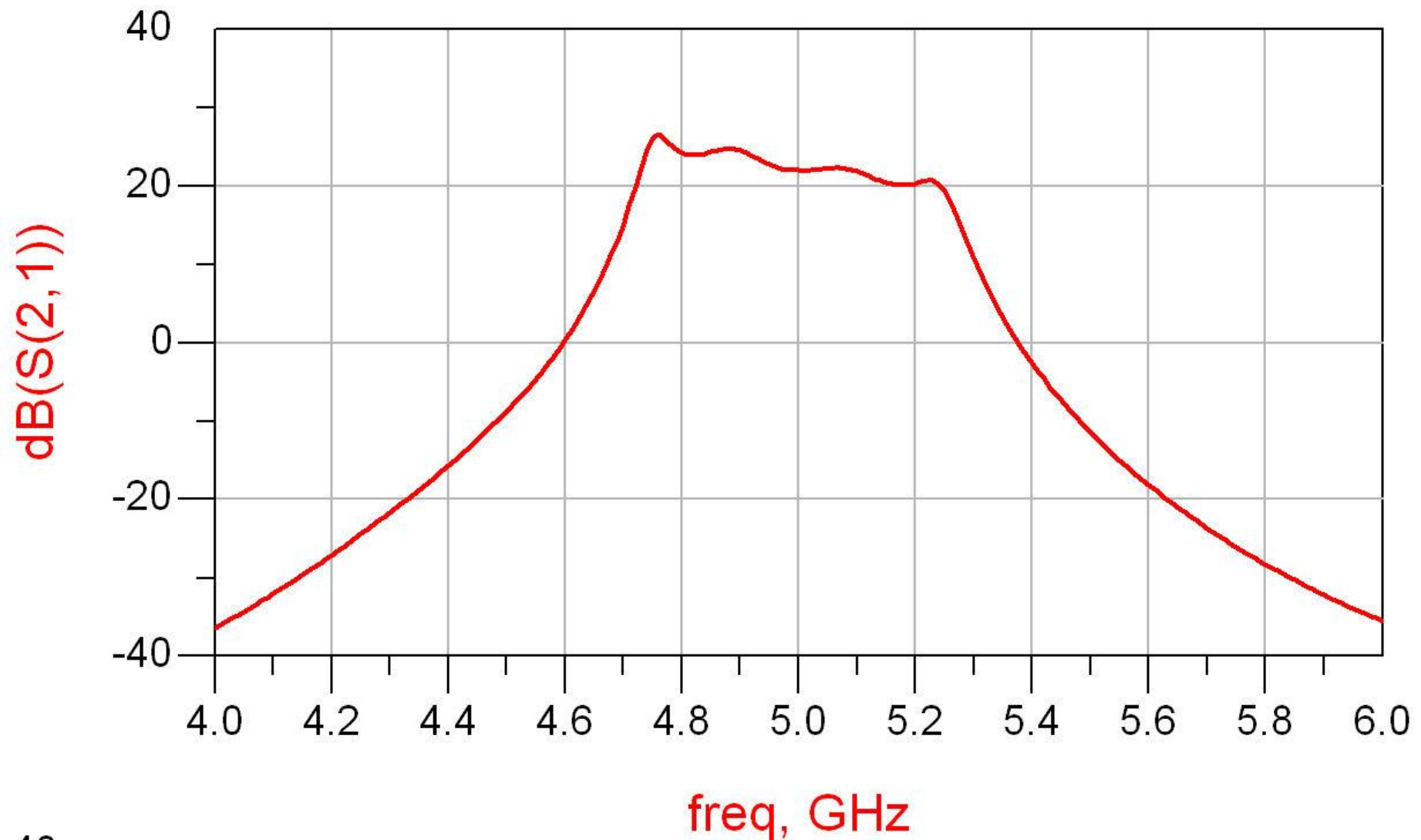
VOLTAGE FEEDBACK AND CONSTANT
BASE CURRENT SOURCE

Exemplu proiect

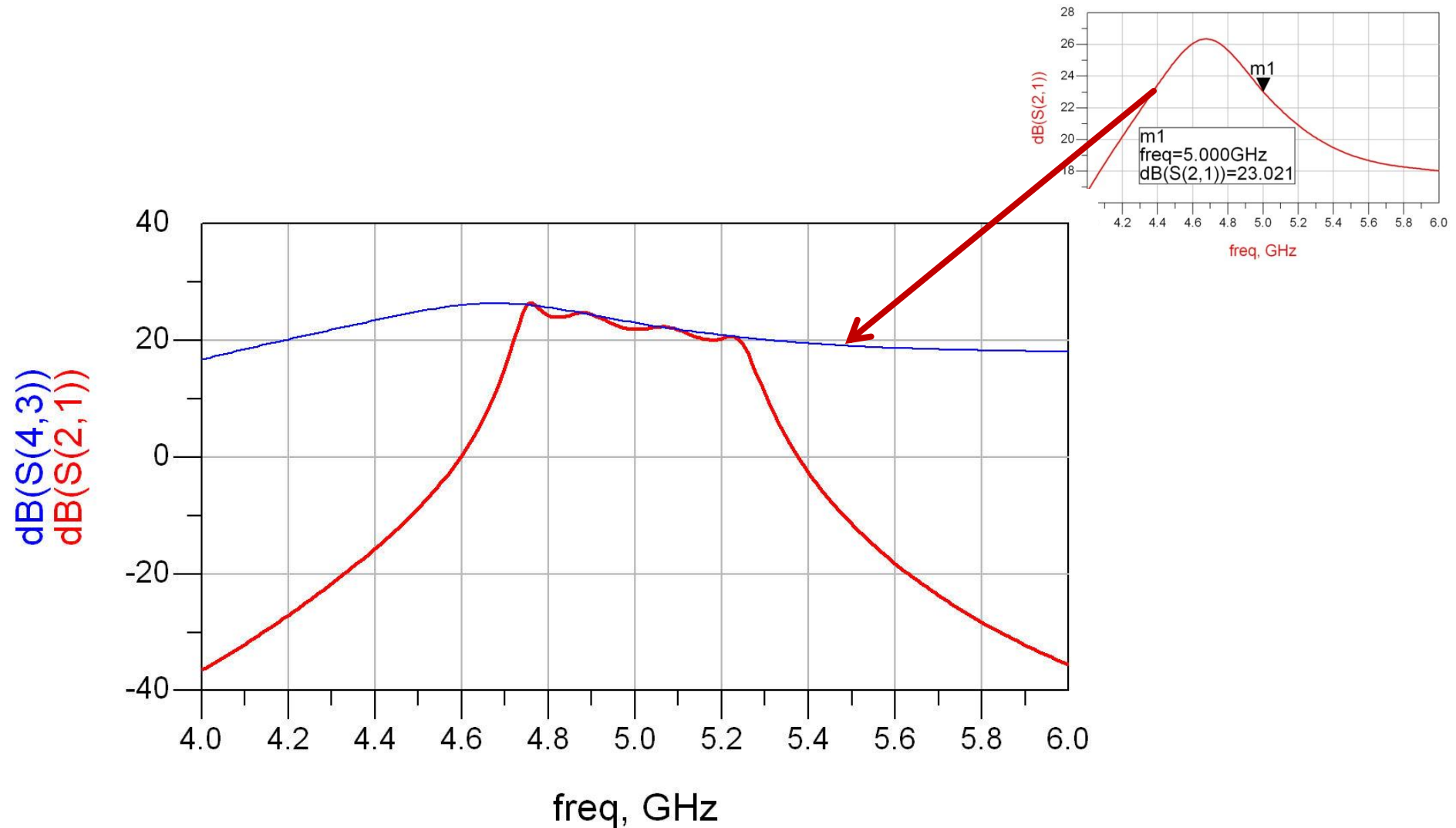
- Unirea celor doua scheme
 - C11 – amplificator (var 4/S36-37)
 - C12 – filtre



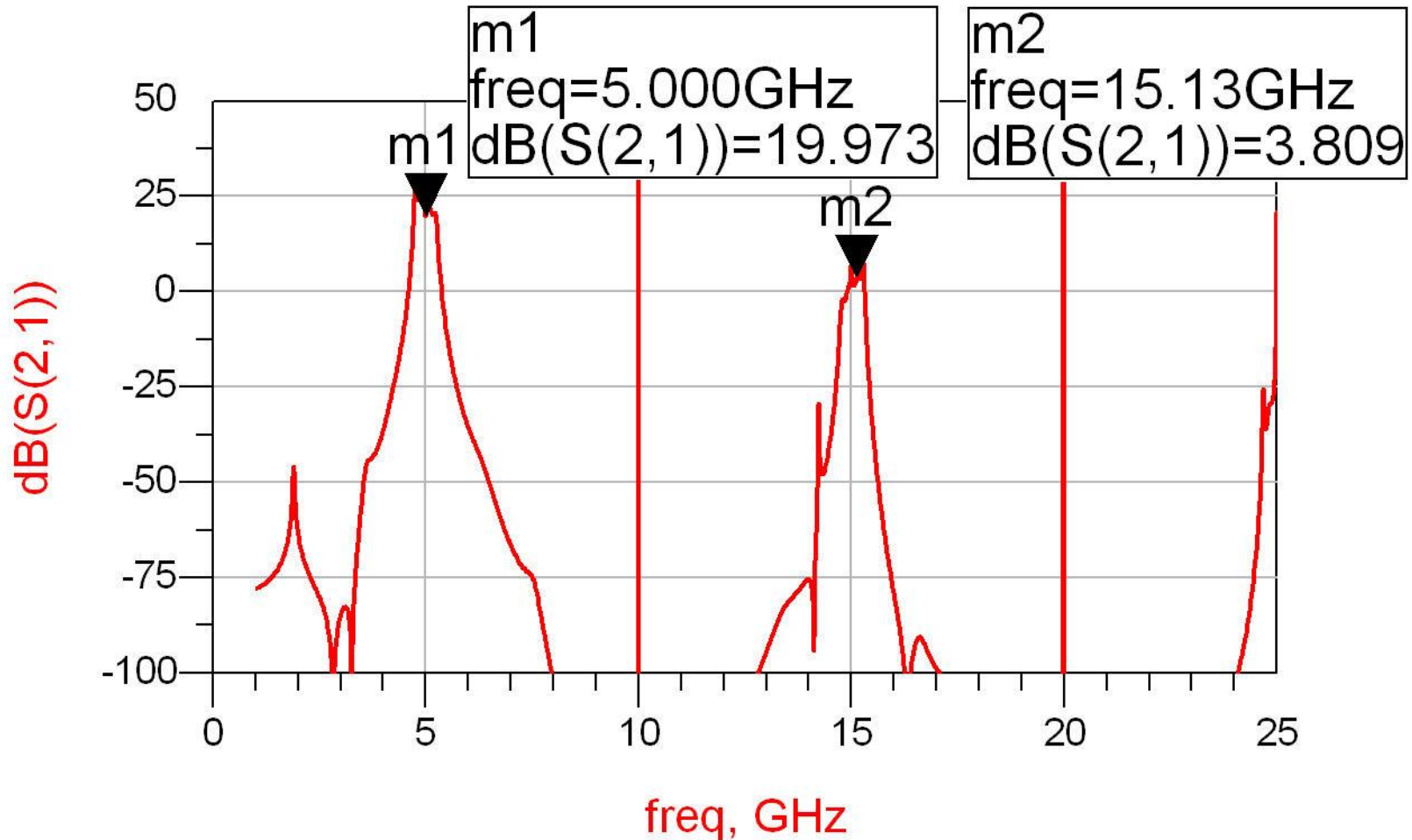
Rezultat (dezechilibrat)



Rezultat (dezechilibrat)

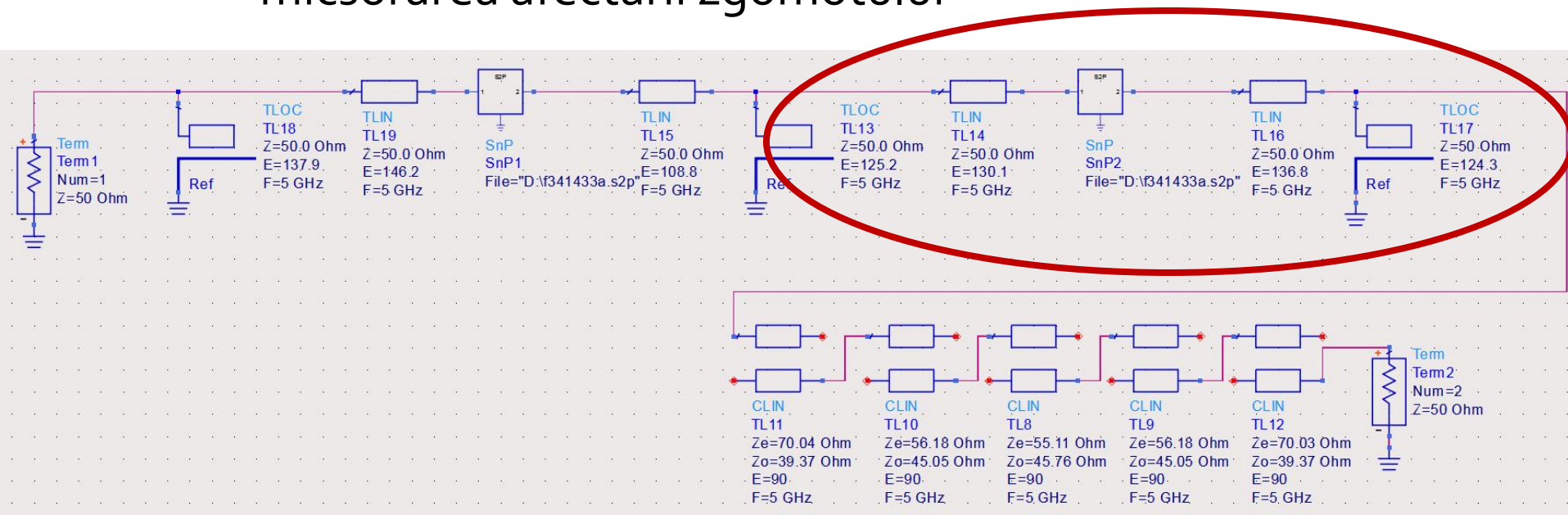


Rezultat (periodic in frecventa)

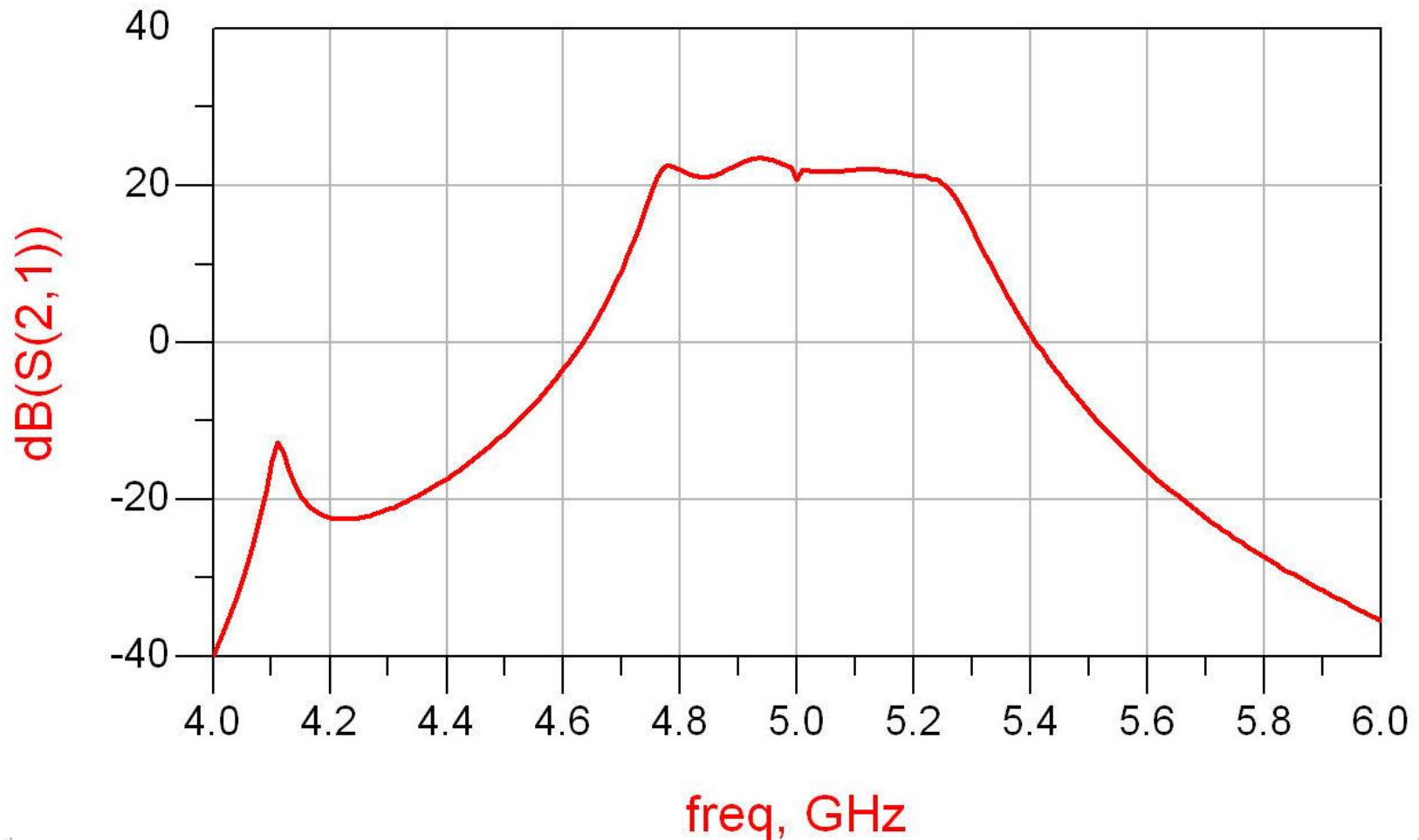


Reglaj -> echilibrare

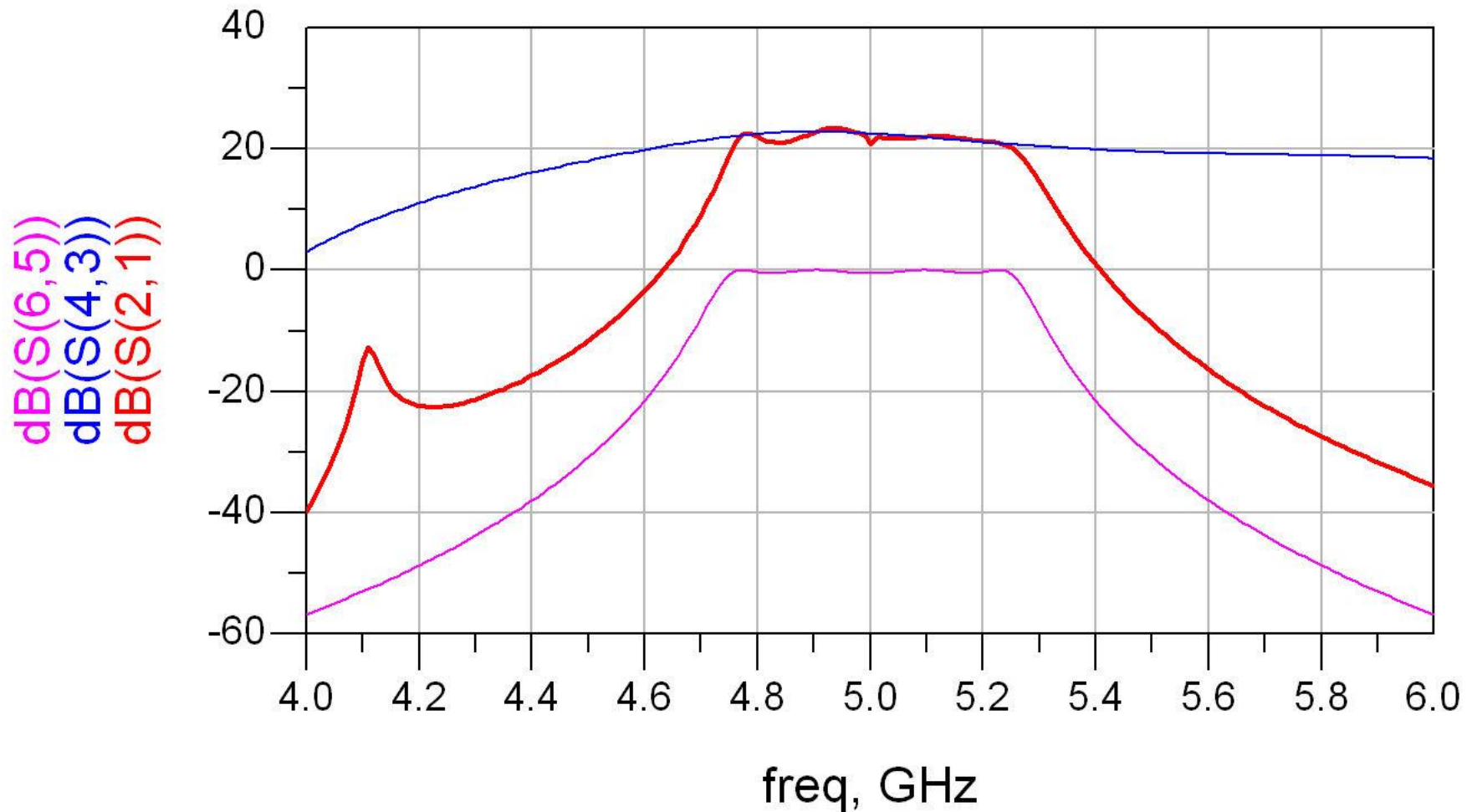
- scopul: echilibrarea caracteristicii amplificatorului (maxim la frecventa centrala)
 - se prefera reglarea lungimii liniilor de la iesirea amplificatorului
 - micsorarea afectarii zgomotului



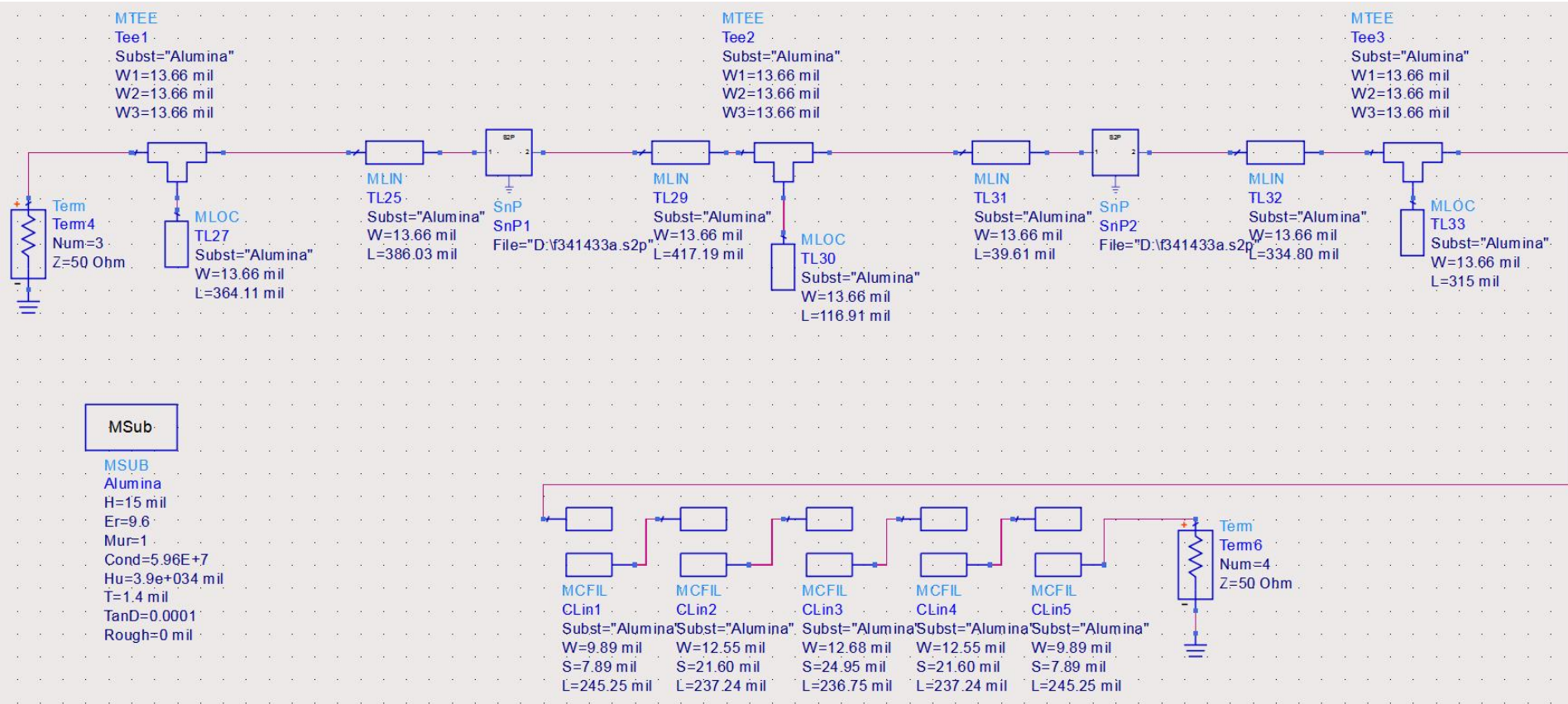
Reglaj -> echilibrare, efect



Amplificator, Filtru, Total

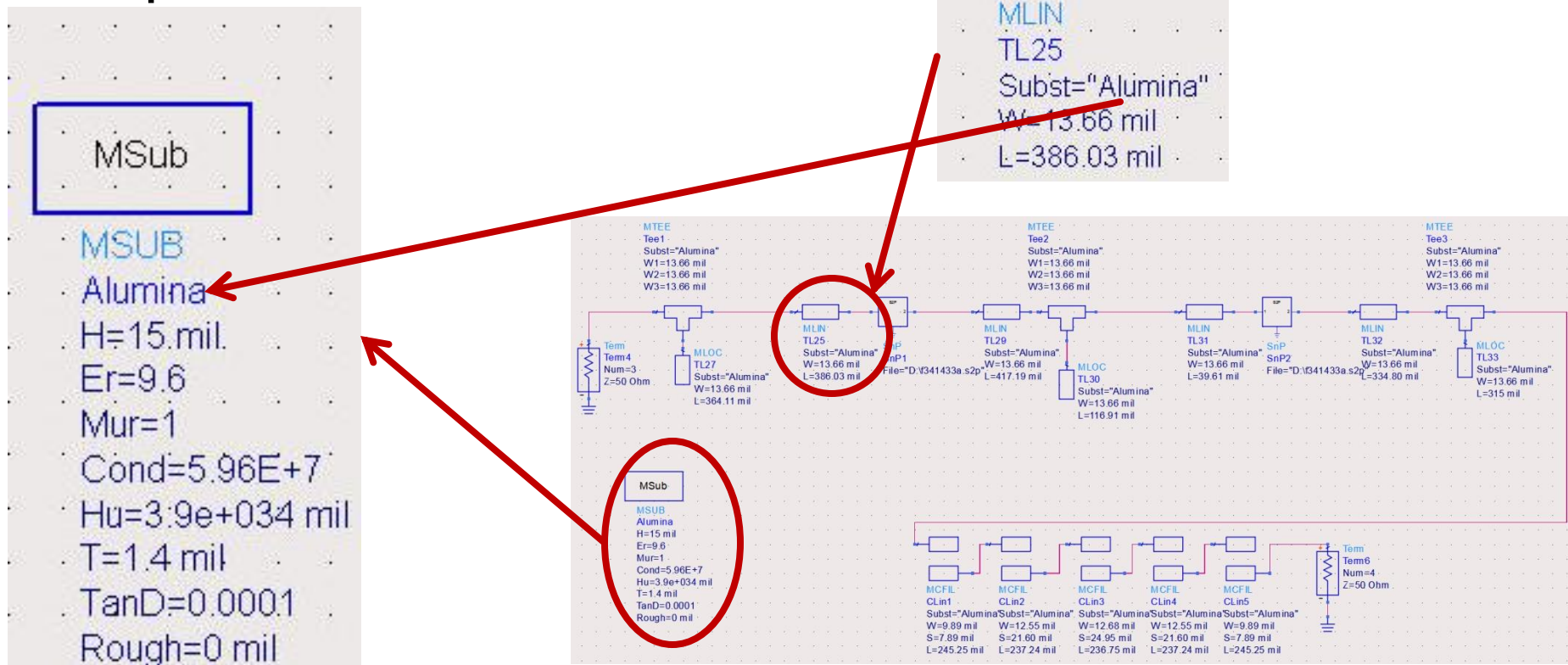


Implementare cu linii microstrip



Implementare cu linii microstrip

- Se introduce modelul de substrat
- Liniile/liniile cuplate se calculeaza cu Linecalc pentru acelasi substrat

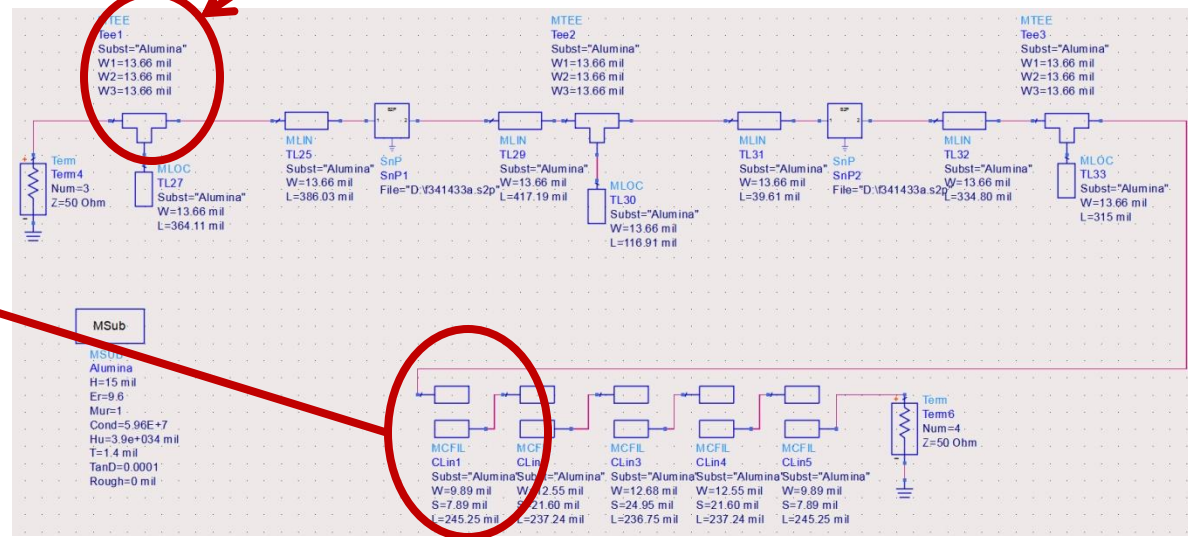
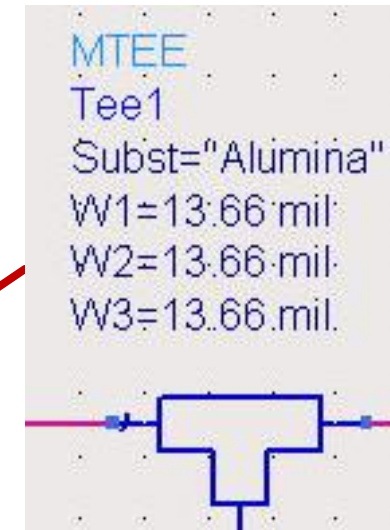


Implementare cu linii microstrip

- Se folosesc componente din paleta Transmission Lines – Microstrip
 - MSUB - substrat
 - MLIN – linie serie
 - MLOC – stub paralel in gol
 - MTEE – modelare conexiune cu stub in paralel
 - MCFIL – sectiune de filtru cu linii cuplate
(alternativa mai precisa decat MCLIN – se tine cont de faptul ca doua sectiuni succesive sunt in fizic alaturate)

Implementare cu linii microstrip

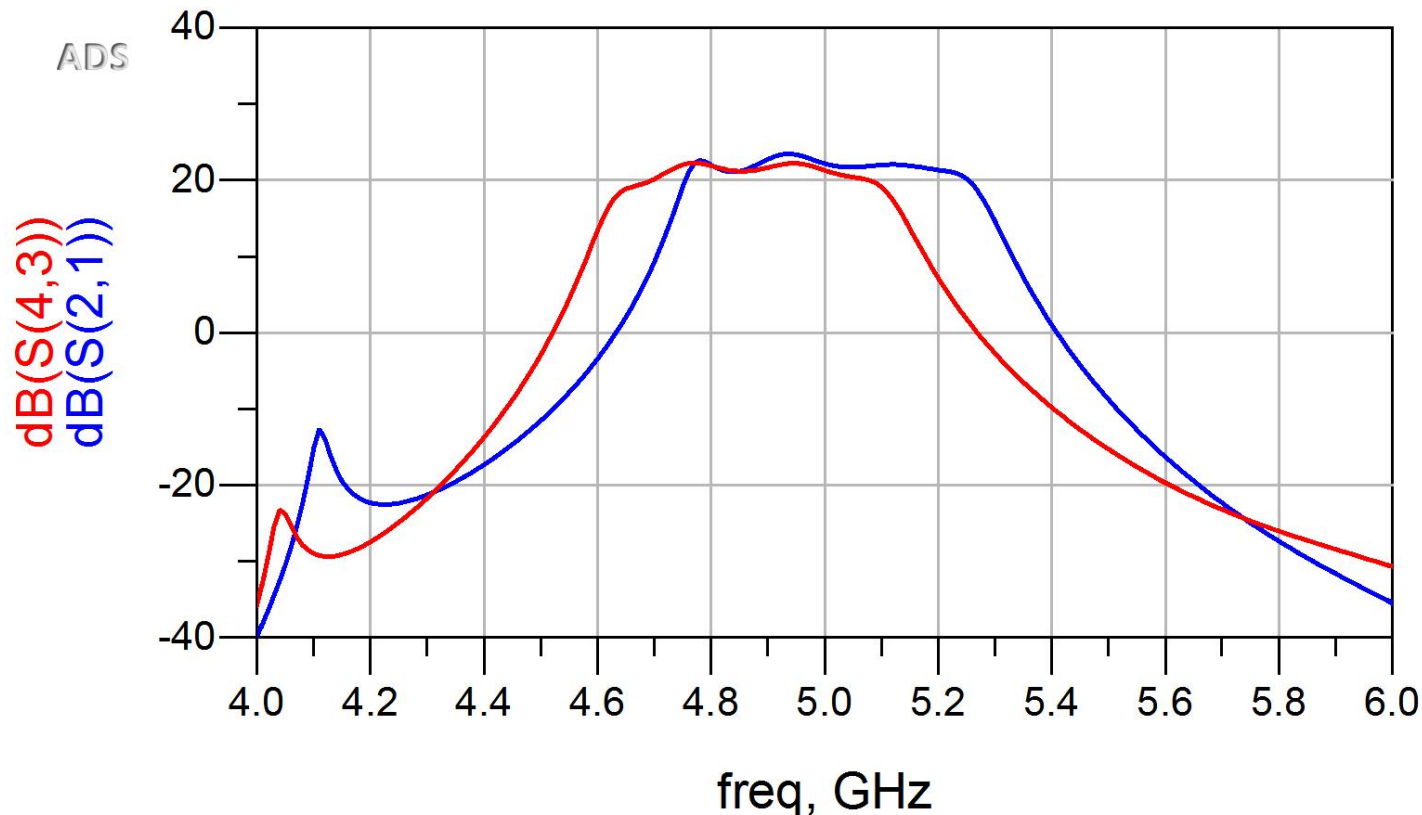
- E necesara atentie la completarea parametrilor pentru MTEE si MCFIL prin verificarea in schema a latimii liniilor conectate la fiecare terminal



Implementare cu linii microstrip

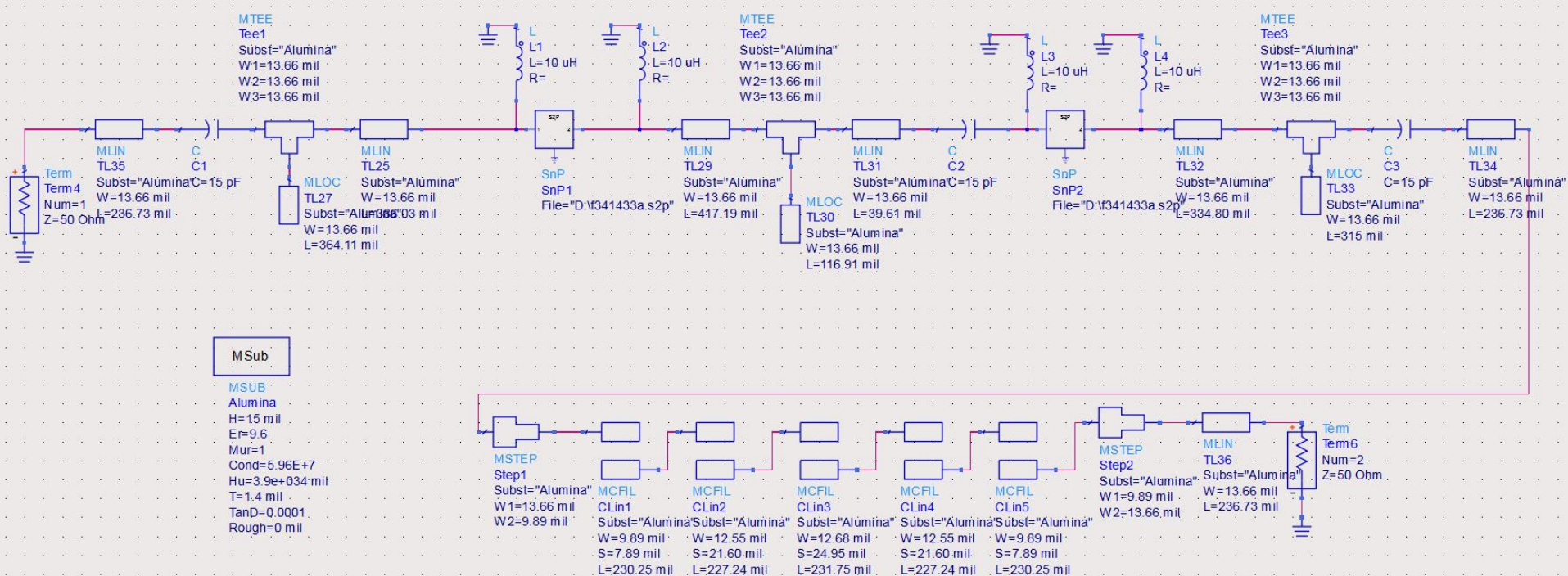
Rezultat

- Se constata o deplasare a benzii obtinute (albastru) spre frecvente mai mici fata de modelele ideale (rosu)
 - datorat diferentei MCFIL / MCLIN

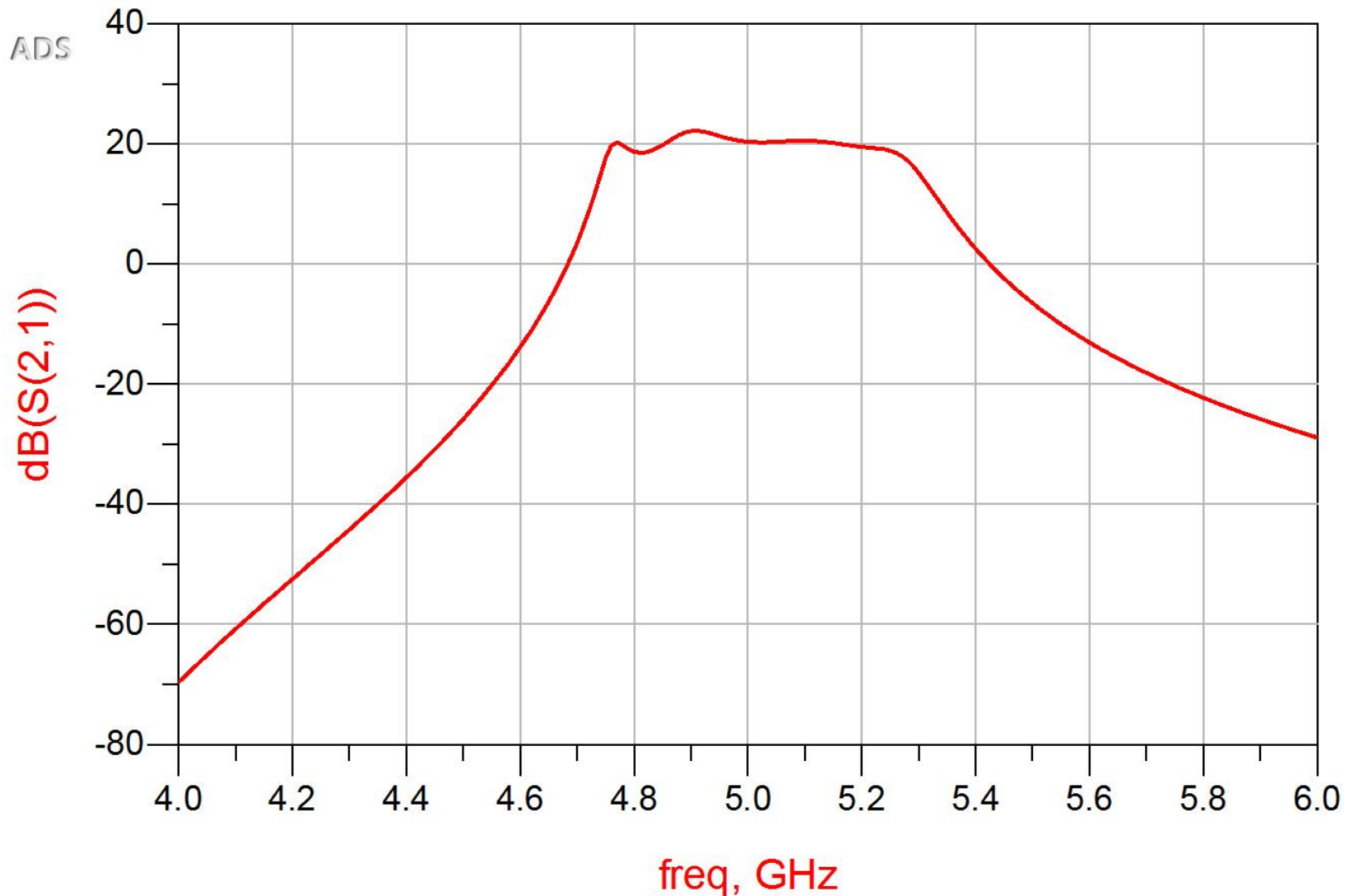


Introdúcere elemente de polarizare

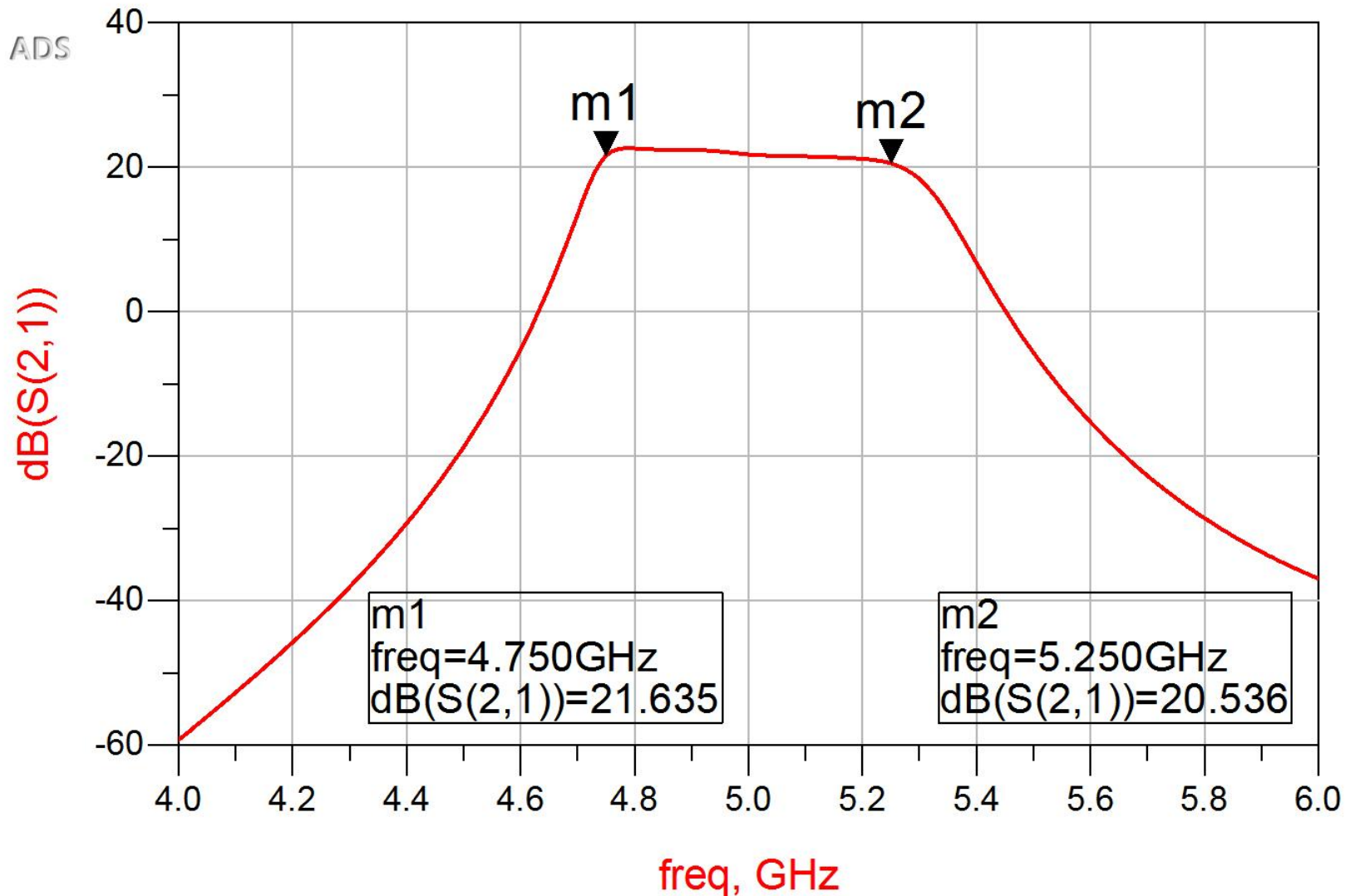
- Reglaj de lungimi la elementele filtrului pentru reglarea frecventelor in jurul $f_0 = 5\text{GHz}$
- Introducere L (soc RF) si C (decuplare)



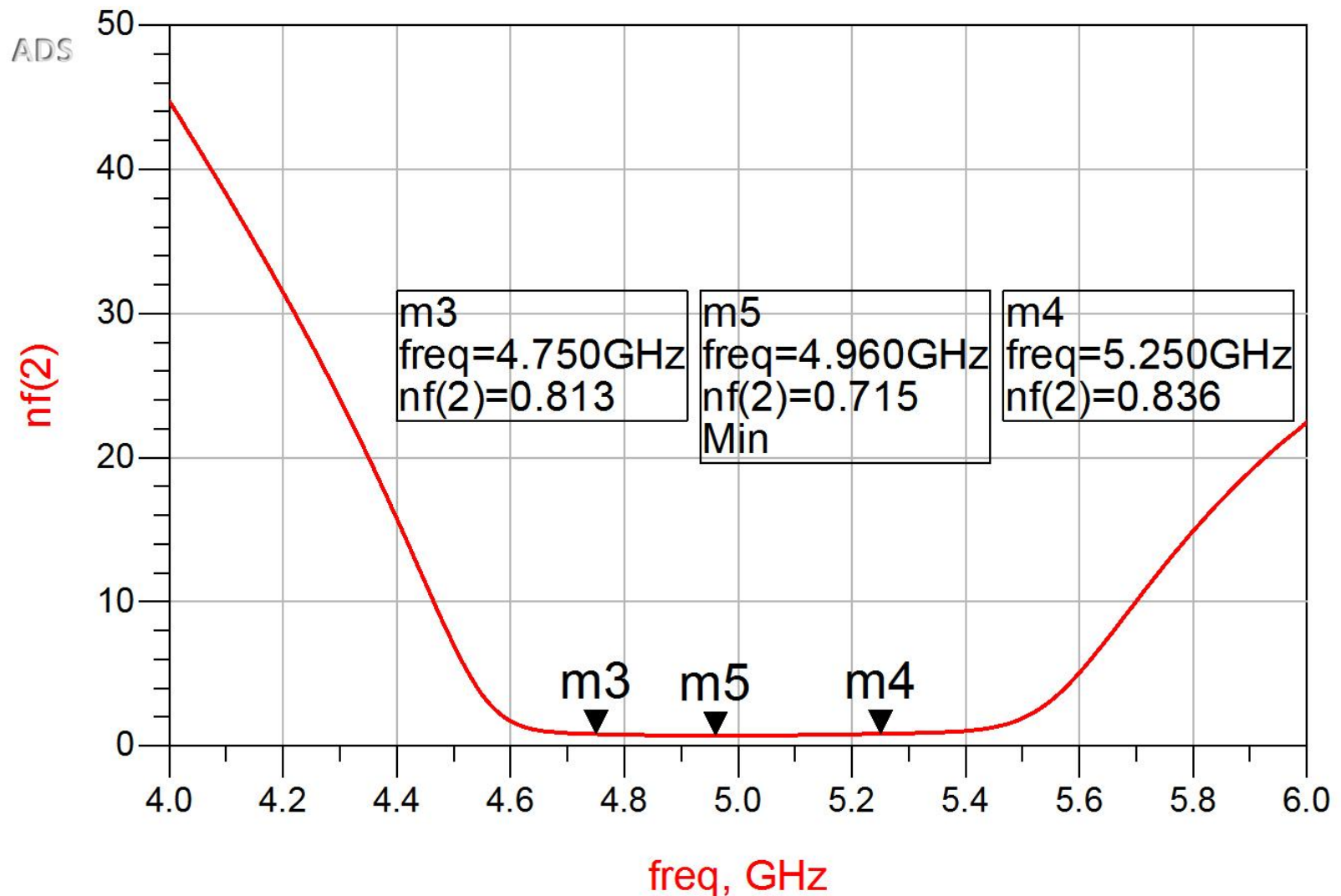
Castig -> Reglaj/Optimizare



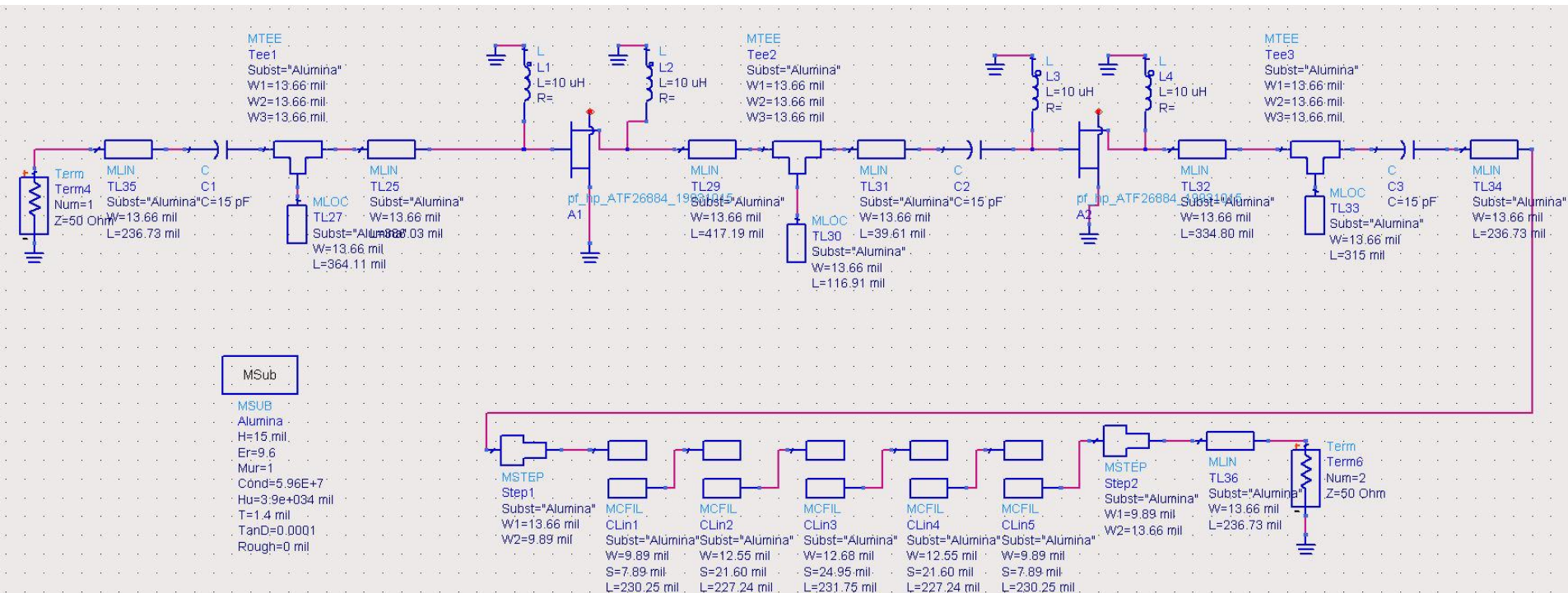
Rezultat final (Castig)



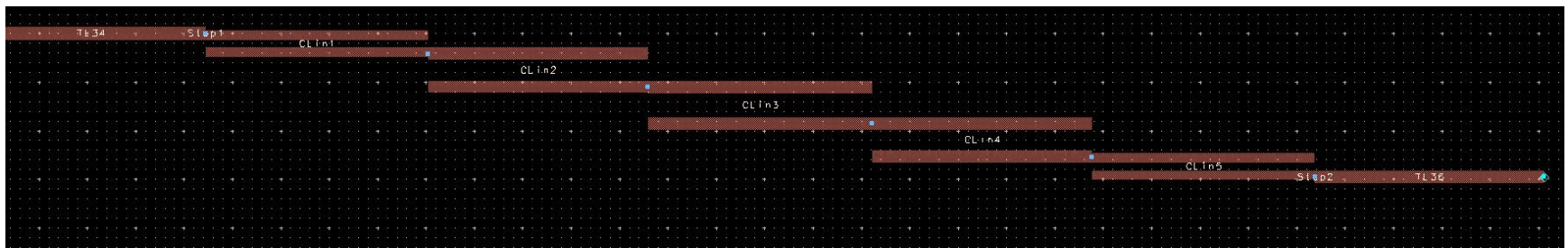
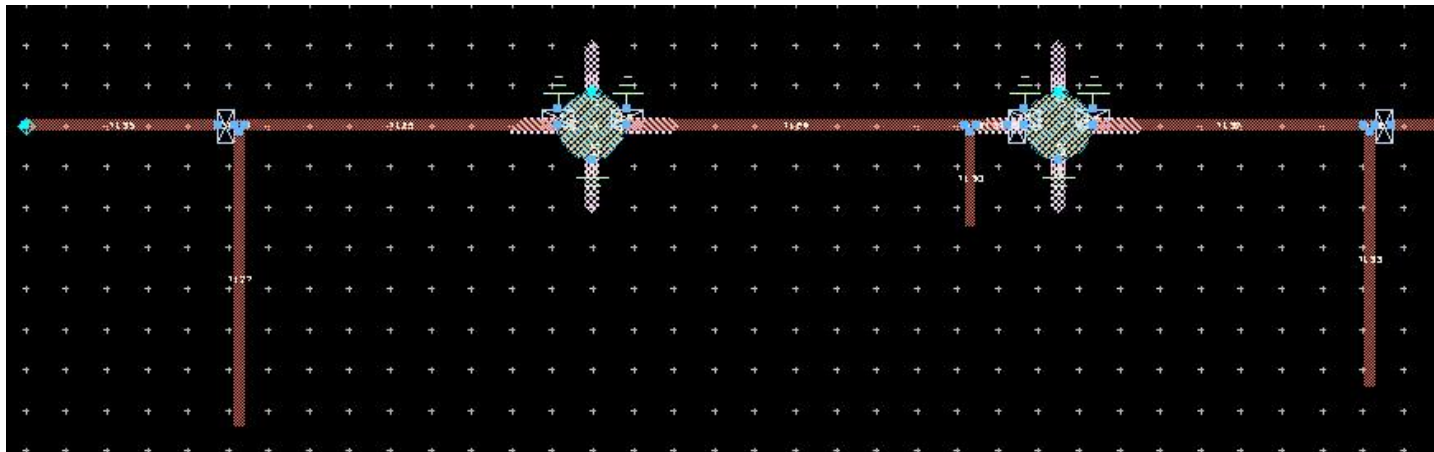
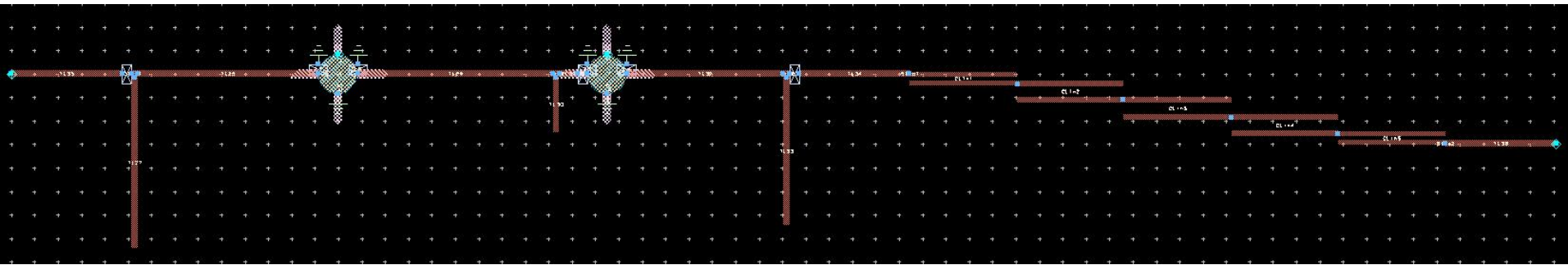
Rezultat final (Castig)



- Inlocuirea (fictiva) a tranzistoarelor si elementelor concentrate (LC) cu elemente pentru care ADS are informatii despre capsule



Layout (Exemplu)



Contact

- Microwave and Optoelectronics Laboratory
- <http://rf-opto.etti.tuiasi.ro>
- rdamian@etti.tuiasi.ro