Microwave Circuits and Sub-Systems

A2M17MOS

Wide-Band Amplifier Design

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Contents

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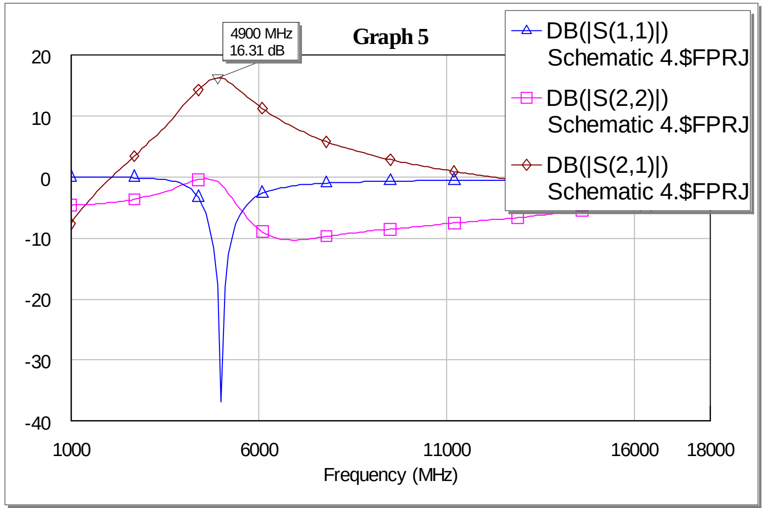
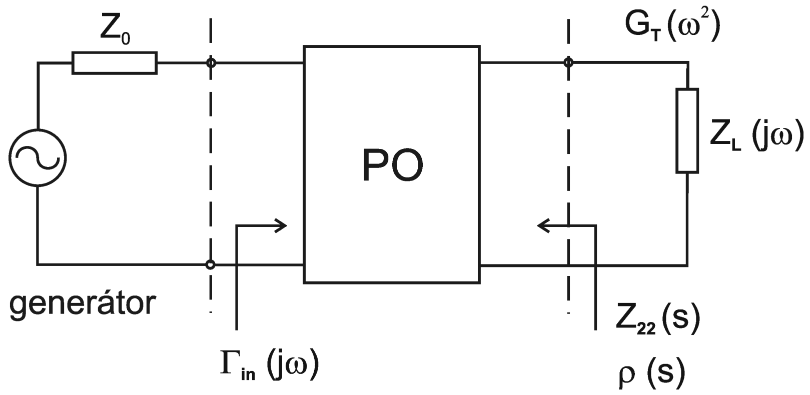
Narrow-Band Impedance Matching

- Design of narrow-band matching circuits → relatively simple
- Ideal impedance matching at 1 frequency can be reached: $|\Gamma_{in}| \to 0 \quad RL \to \infty$
- Solution exists always
- Numerous solutions exist
- Solutions can be calculated using direct design formulas
- Applicable for frequency bands $B \leq 10\%.f_0$
- The 10% frequency band comply with many communication applications
- But wider frequency bands are also required



Wide-Band Impedance Matching

- Substantially more complicated
- Ideal matching $\left|\Gamma_{in}(j\omega)\right| \to 0$ can never be reached in the whole frequency band
- Ideal matching can be reached only at separate and distanced frequencies



- **Attainable condition:**

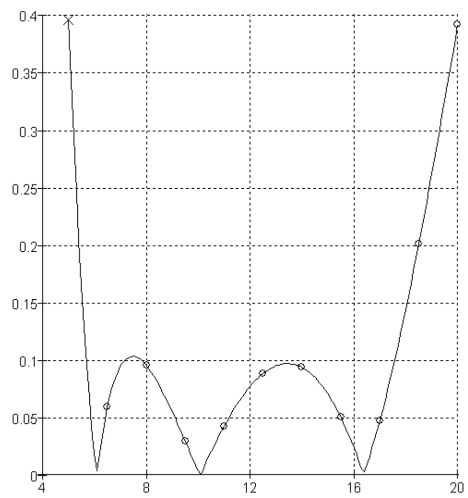
$$\left|\Gamma_{in}(j\omega)\right| \le const. = \left|\Gamma_{\max}\right|$$

- In the frequency band

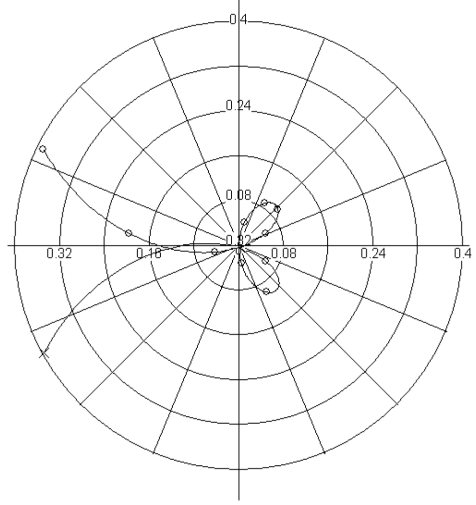
$$B = \Delta\omega = \omega_b - \omega_a$$

- Γ_{in} passes through 0 at distanced frequencies
- In many cases – does not pass through 0 at all

Graph 1: Amplitude of S11 [-] - frequency [GHz]

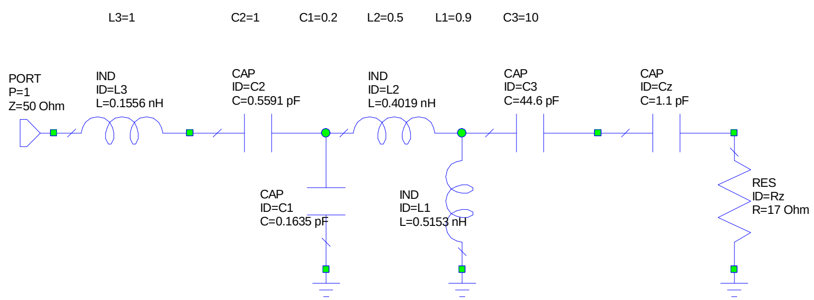


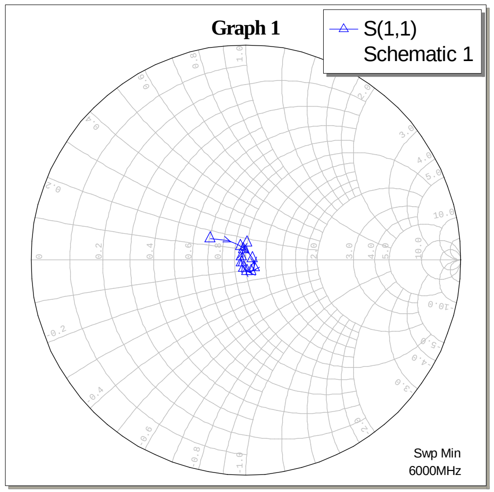
Graph 2: Polar S11 [-]



Wide-Band Impedance Matching

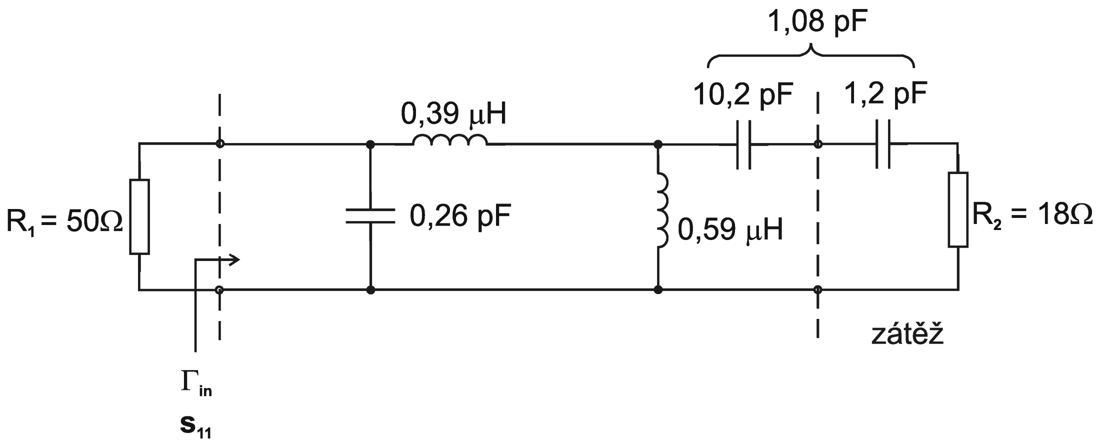
- Simple and direct solutions exist only rarely
- For the given requirements → **NO solution can exist**
- Solutions are based on **approximations and iterations** nearly always
- **CAD optimization** is recommended
- Numerous different approaches exist
- **Analytical solutions** → based on analog filter design
- **Iteration solutions** → based on optimization of the recommended wideband structures





Analytical Solutions

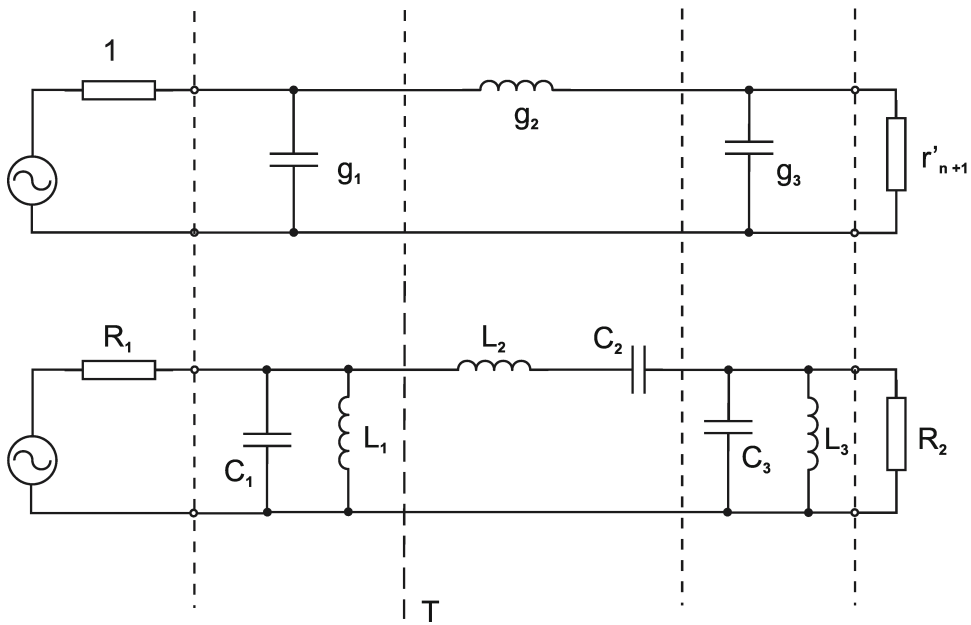
- Based on the **filter synthesis process**
- Filters with **different impedances at both ports** can be designed
- The filter synthesis process leads to **LC structures**
- The LC structures must be converted to structures practically realizable at microwave frequencies (real LC, microstrip lines, …) – the conversion is never ideal (B drops by 50% usually)
- During the design process, several **approximations** must be used
- Many different approaches can be found
- Wideband MC synthesizers are available as part of the RF CAD programs usually
- More details e.g. in: „Aktivní mikrovlnné obvody“, skriptum ČVUT.



Example of Analytical Design Method

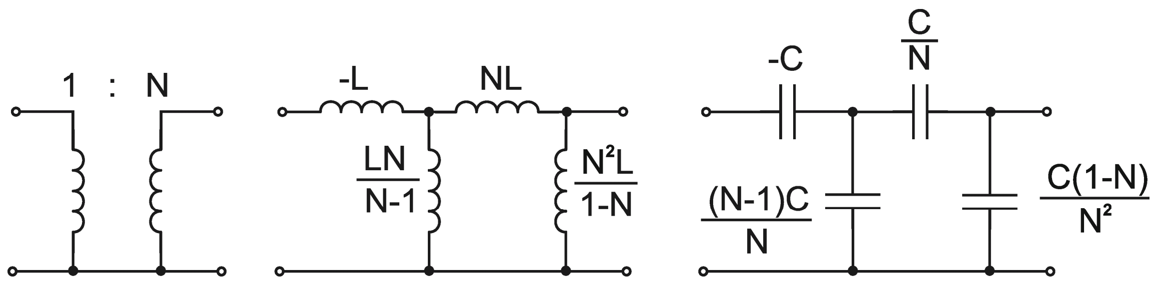
- Wide-band MCs = filters with **maximum impedance transformation**
- One of possible solutions with very good results
- Based on standard analog filter synthesis:
  - Normalized **low-pass filter prototype** based on the required Γ_{max} value
  - **Thebyschev or Butterworth transfer-function approximations**
  - Conversion of the low-pass prototype to the band-pass filter
- Example: Individual task No.5.

- The synthesized filters show different load impedances R_1 and R_2 at both ends
- The filter is able to perform the $N=R_2/R_1$ **impedance transformation**

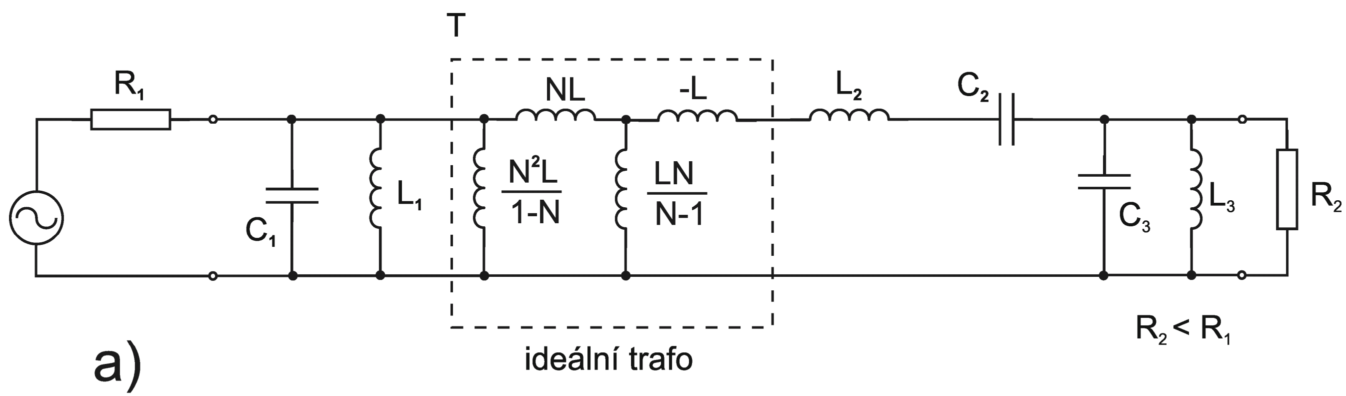


Additional Impedance Transformation

- In order to raise the $N=R_2/R_1$ ratio, an **additional impedance transformer** can be connected to the filter structure
- The standard transformer can be realized by equivalent L or C circuits

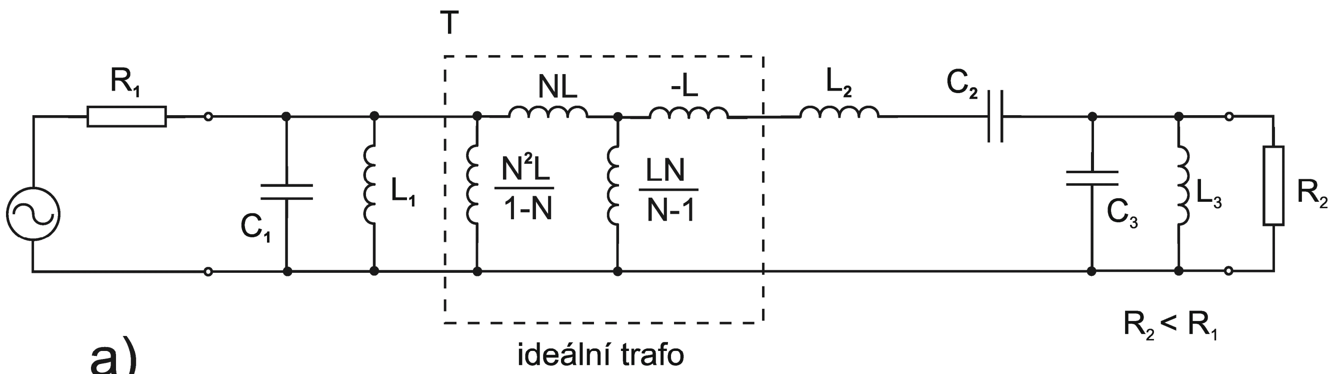


- Values of the side L, C elements are negative and can be absorbed in the filter structure

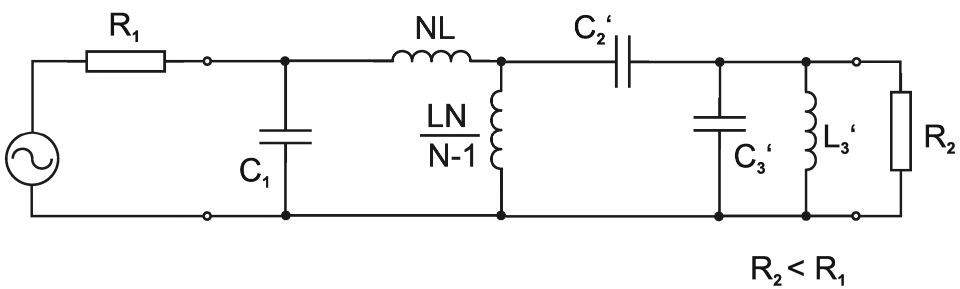


Resulting Wide-Band MC

- The negative equivalent circuit values can be **canceled out** by the neighboring positive filter elements
- These conditions lead to L and N values (2 equations with 2 unknowns)
- The resulting MC provides the **maximum available impedance transformation**
- Or for the fixed transformation ratio → **the maximum frequency band-width**



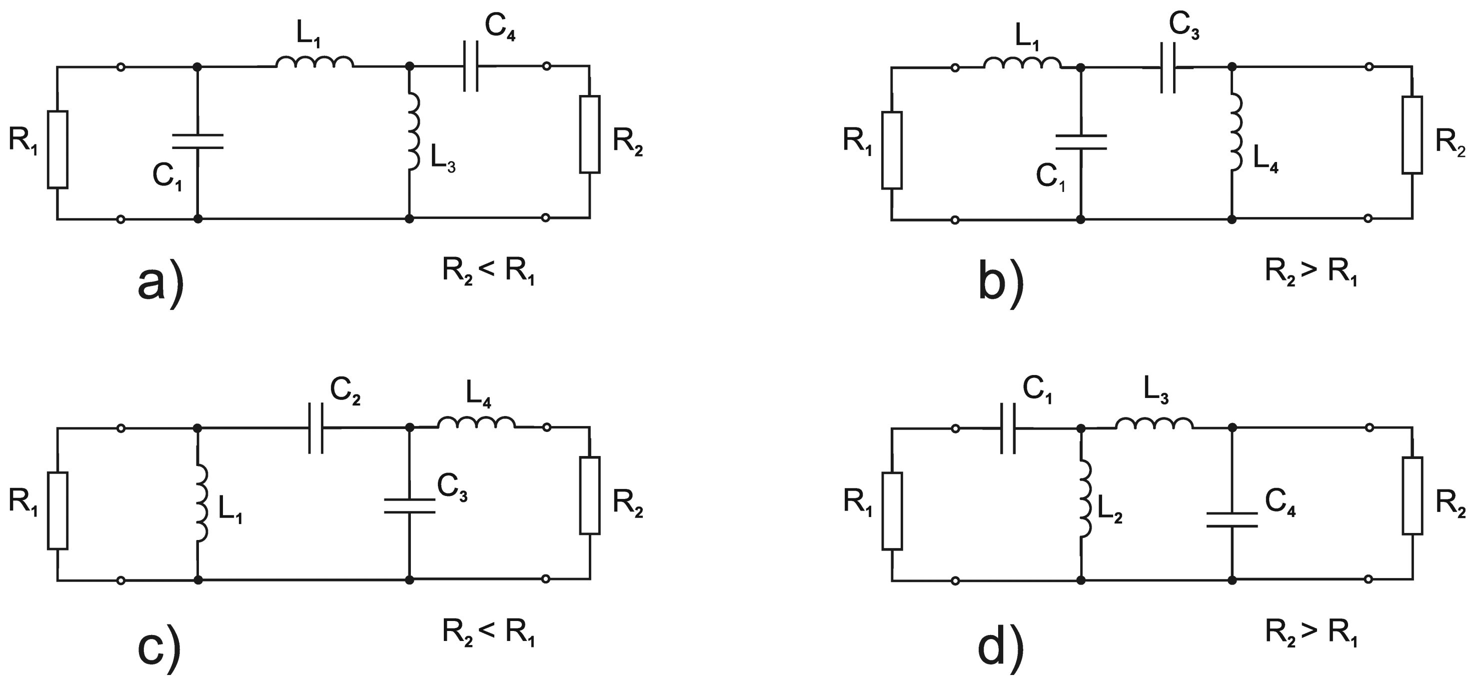
a)



b)

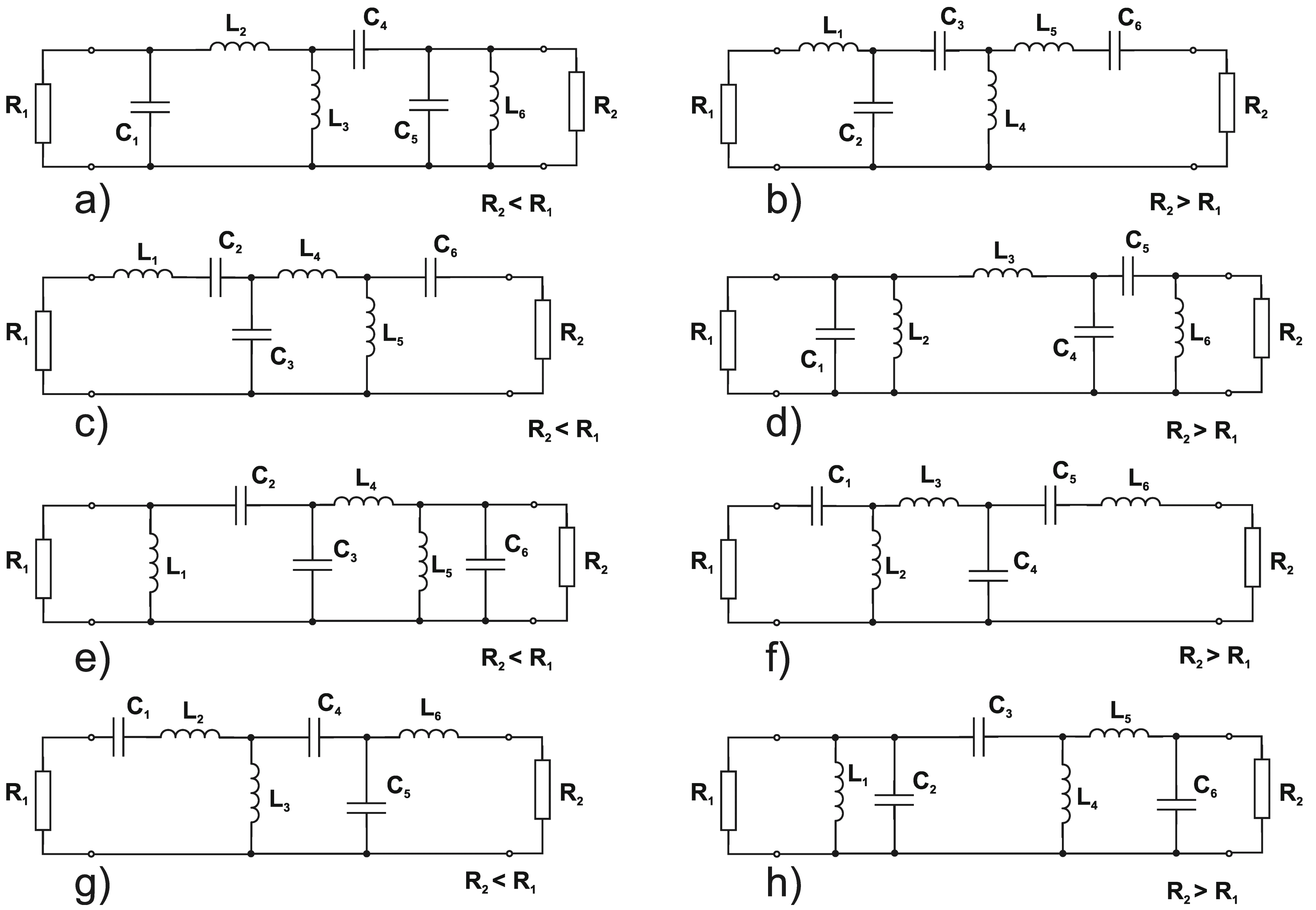
Recommended Structures n=2

- Recommended structures based on the **2-nd order filter**
- Structures a) and c) provide the **down-transformation $R_2<R_1$** (e.g. from 50Ω to 15Ω)
- Structures b) and d) provide the **up-transformation $R_2>R_1$** (e.g. from 50Ω to 250Ω)
- Values of the LC elements can be calculated or obtained by the CAD optimization



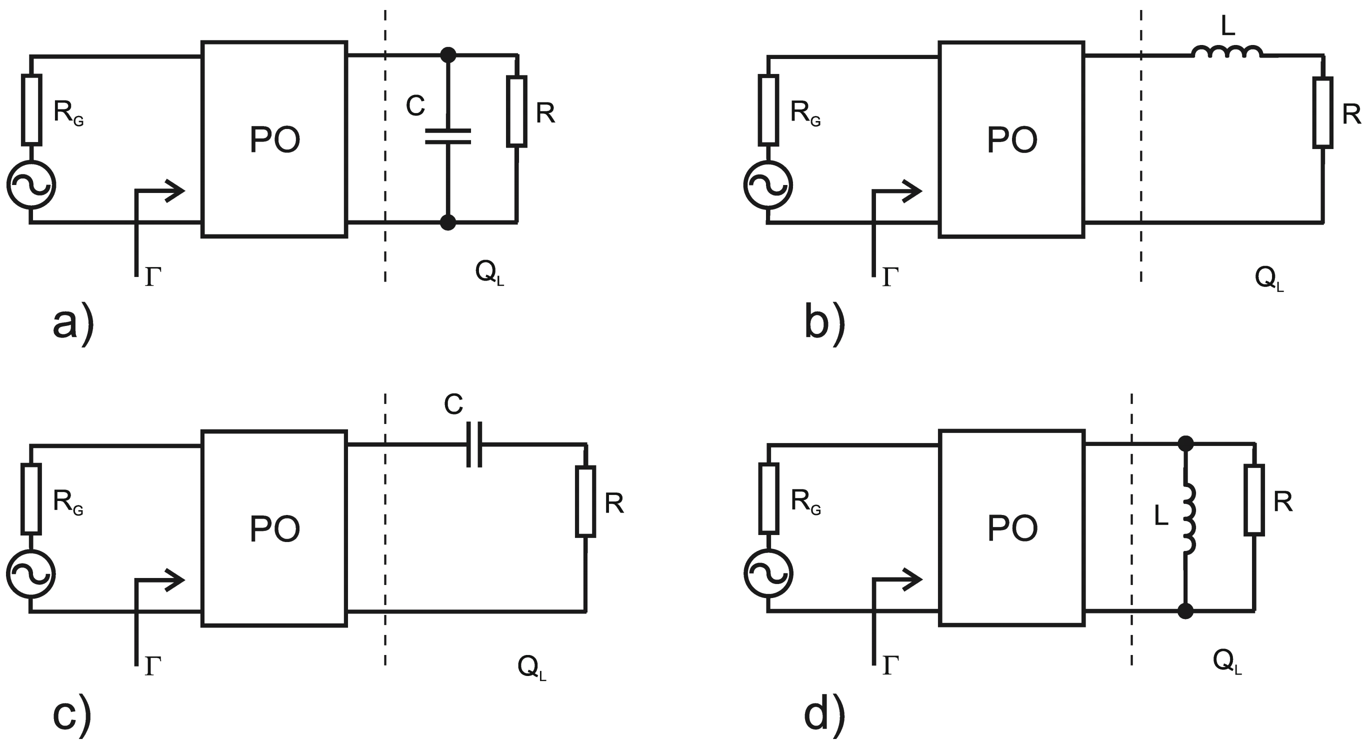
Recommended Structures n=3

- Recommended structures based on the **3-rd order filter**



Problems

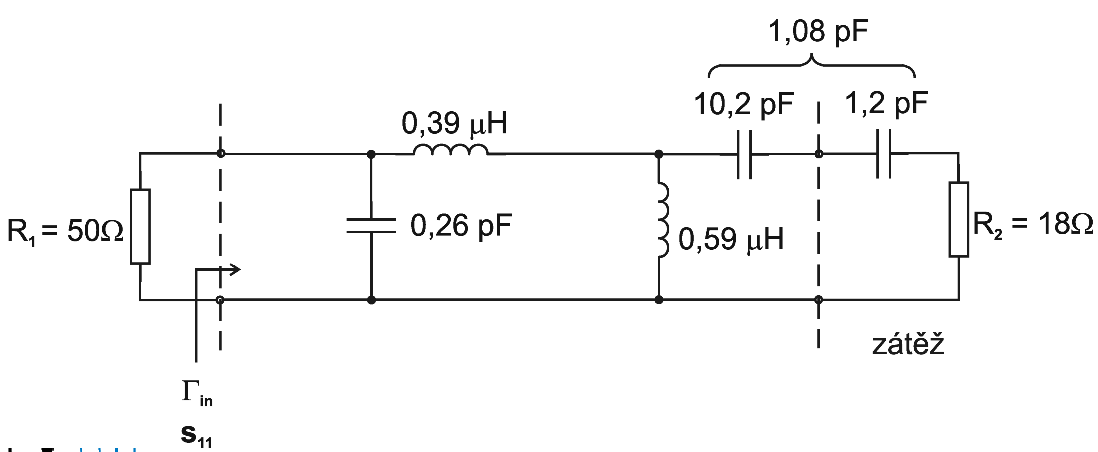
- Usually, only one of impedances of the wide-band MC is real (e.g. $R_1=Z_0=50\Omega$)
- The latter impedance is generally complex
- **RLC approximation can be used for modeling the complex load**
- If lucky, the reactance part of the load can be **absorbed into the matching filter**
- Often used wide-band load equivalent circuits



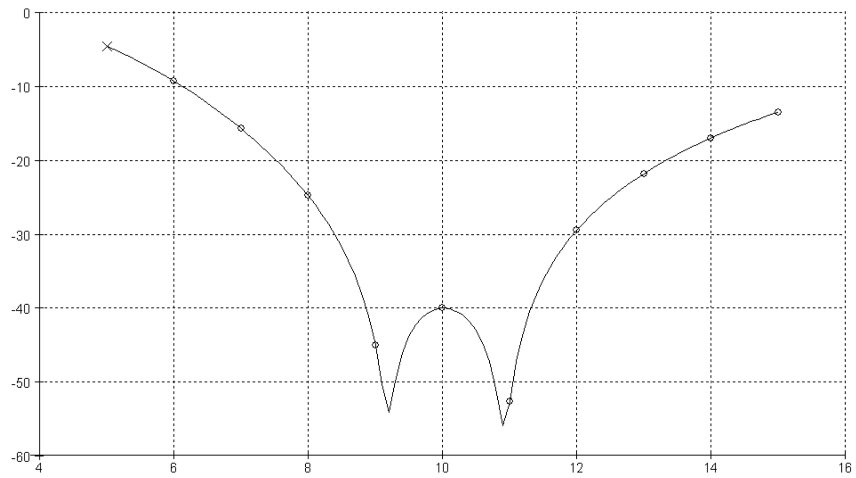
Example

- The input of the microwave transistor can be modeled as a **series combination of the 18Ω resistance and 1,2pF capacitance**
- Structure n=2 and down-converting MC can be chosen
- Using design formulas, the C_1, L_2, L_3 and C_4 values can be calculated → C_4=1,08pF
- This capacitance can be realized by the series combination of the C_L=1,2pF load capacitance and C_{ext}=10,2pF external filter capacitance
- Formulas used

$$C_4 = \frac{C_{ext} C_L}{C_{ext} + C_L} \qquad C_{ext} = \frac{C_4 C_L}{C_L - C_4}$$

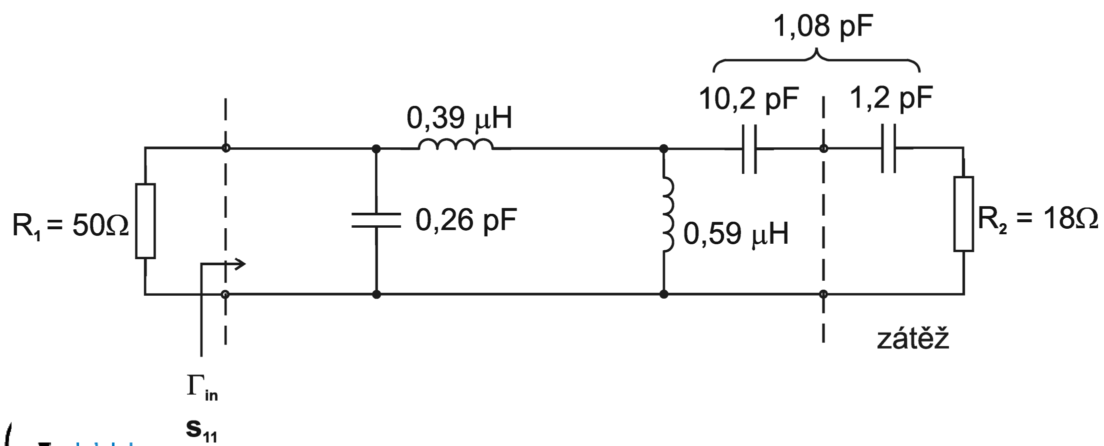


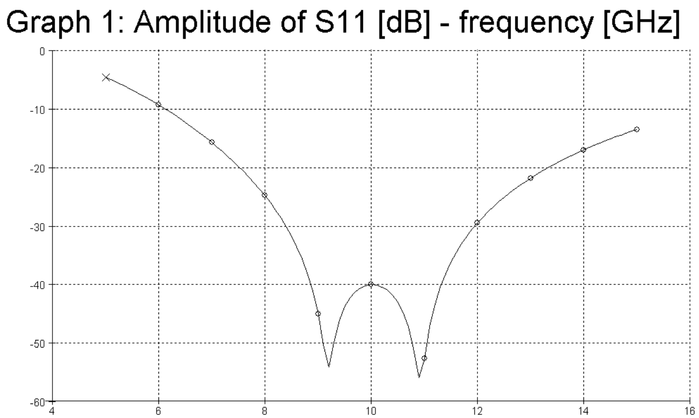
Graph 1: Amplitude of S11 [dB] - frequency [GHz]



Problems

- If $C_4 \geq C_L$ → ideal absorption cannot be reached
- It is recommended to neglect the C_{ext} and optimize the circuit using the CAD tool
- **Lumped LC matching circuits** are applicable at low frequencies (<1GHz) or inside MMICs
- Parasitic properties of real LC elements must be taken into account
- At higher frequencies, **distributed** (transmission line based) **solutions** must be used
- The lumped→distributed conversion can be performed
- But any lumped→distributed conversion **narrows B substantially** ($B_{distributed} \approx 0{,}5 B_{lumped}$)
- Direct synthesis of wide-band distributed MCs is advantageous

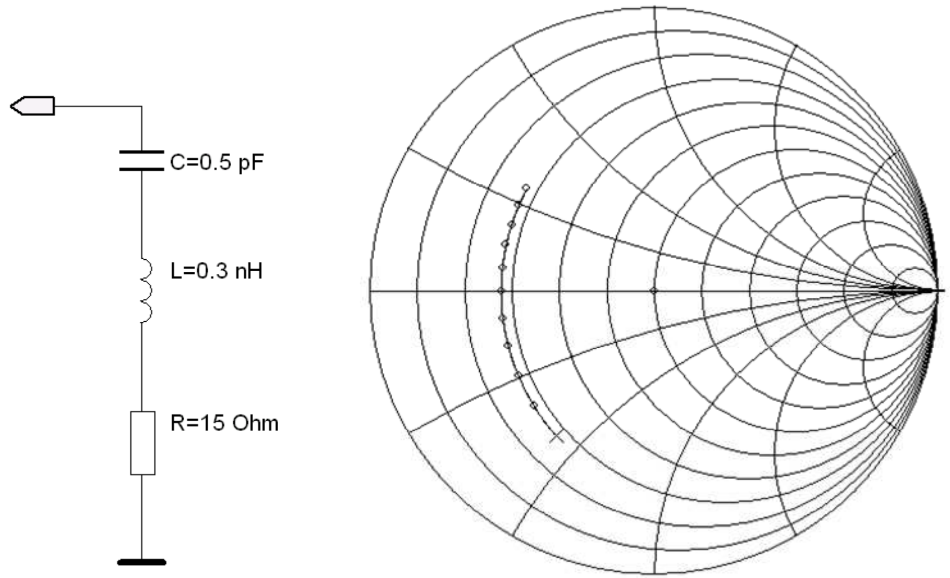




Recommended Distributed MC – Step No.1

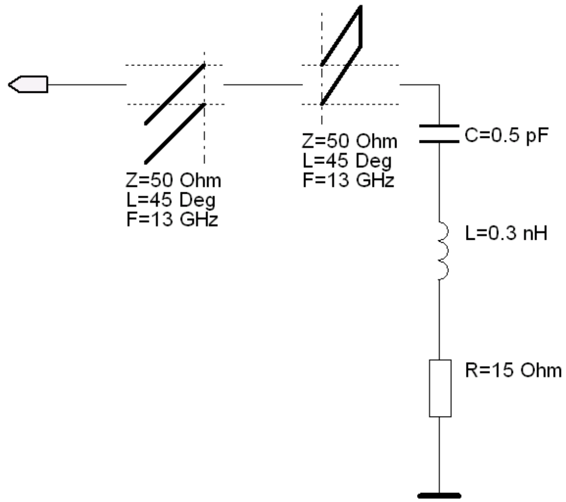
- Direct synthesis of a wide-band distributed MC – can be performed by an **iterative process**
- Based on the calculated or measured $\Gamma_L(j\omega)$ frequency plot
- The MC can be synthesized in several relatively simple steps supported by a suitable CAD tool
- Recommended for octave band-width requirements
- Individual task No.5

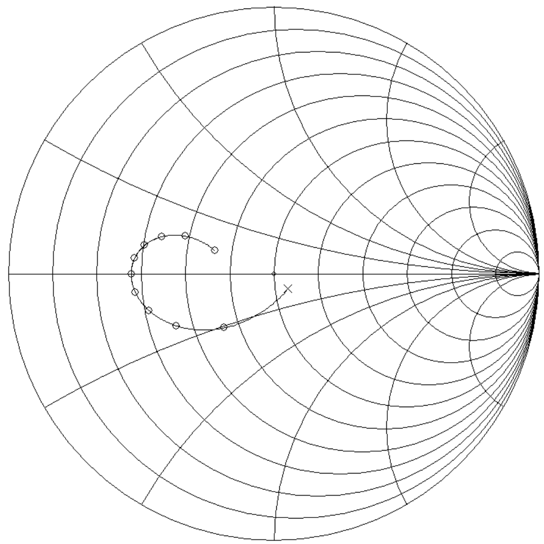
- Example: RLC load 8-18GHz
- **Step No. 1:**
- Center of the $\Gamma_L(j\omega)$ plot @13GHz should be transformed to the X=0 axis by using a section of the transmission line



Recommended Distributed MC – Step No. 2

- Add parallel connection of 2 stubs:
    - o $\lambda/8$ long open stub
    - o $\lambda/8$ long short stub
- Open/short stubs enable to **curl ends of the initial $\Gamma_L(j\omega)$ frequency plot**
- The area of the $\Gamma_L(j\omega)$ gets smaller, but still at a wrong position in the Smith chart



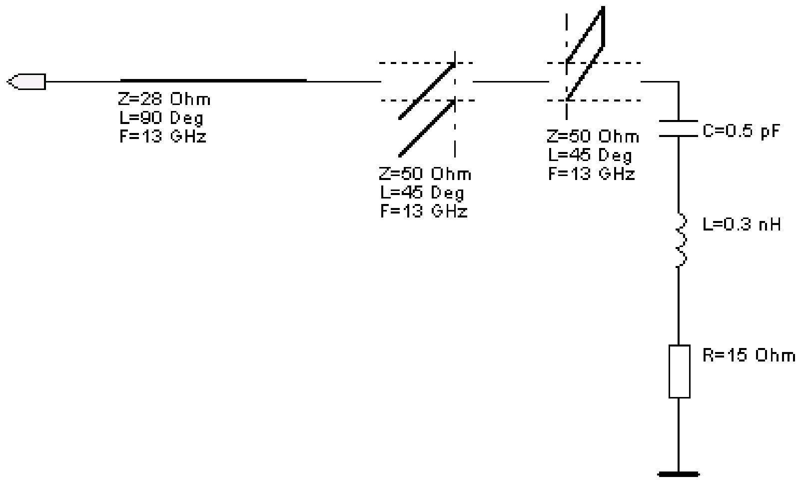


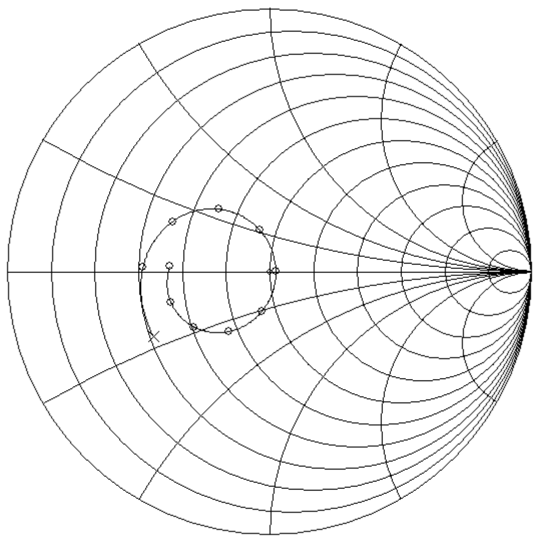
Recommended Distributed MC – Step No. 3

- In order to shift the Γ(jω) frequency plot closer to the Smith chard center, it is advantageous to use the **λ/4 impedance transformer**
- Using the λ/4 long microstrip line with impedance Z_1, it is possible to transform the Z_x impedance to the Z_0 impedance:

$$Z_1 = \sqrt{Z_x Z_0} = \sqrt{15.50} = 27{,}4\Omega$$

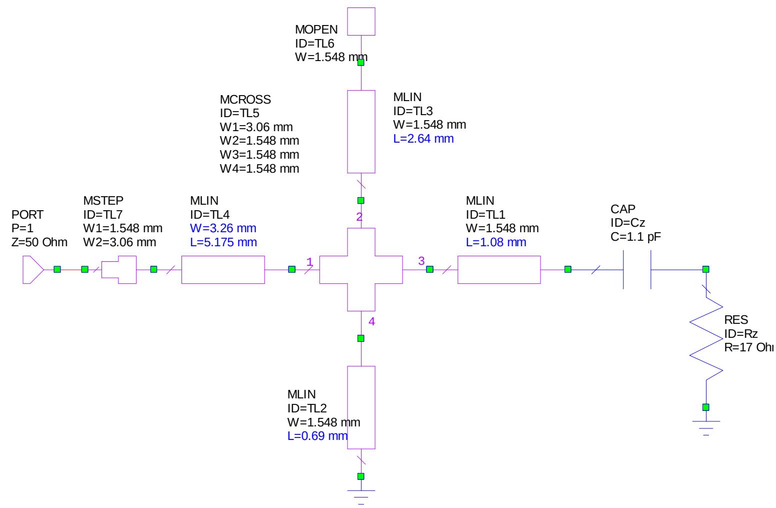
- The λ/4 transformer enables to shift the curled Γ(jω) frequency plot to the surroundings of the Smith chart center

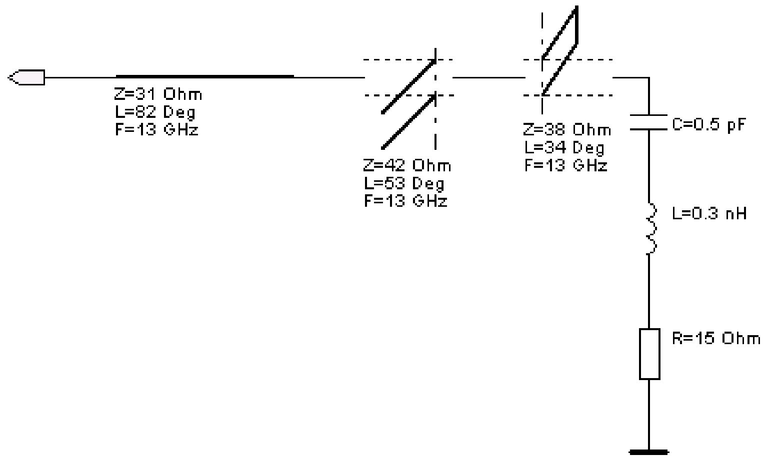


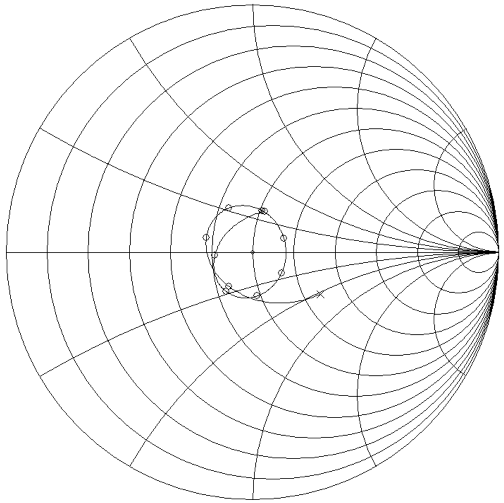


Recommended Distributed MC – Step No. 4

- **Final CAD optimization:**
  - Fine tuning of all transmission line sections
  - Add models of discontinuities (MCROSS, MTEE, MSTEP, …)
- The resulting MC is:
  - Wideband
  - Shows a directly realizable microstrip structure
- Example: RCL load 8-18 GHz → $\Gamma_{in}(j\omega) \le 0{,}3$





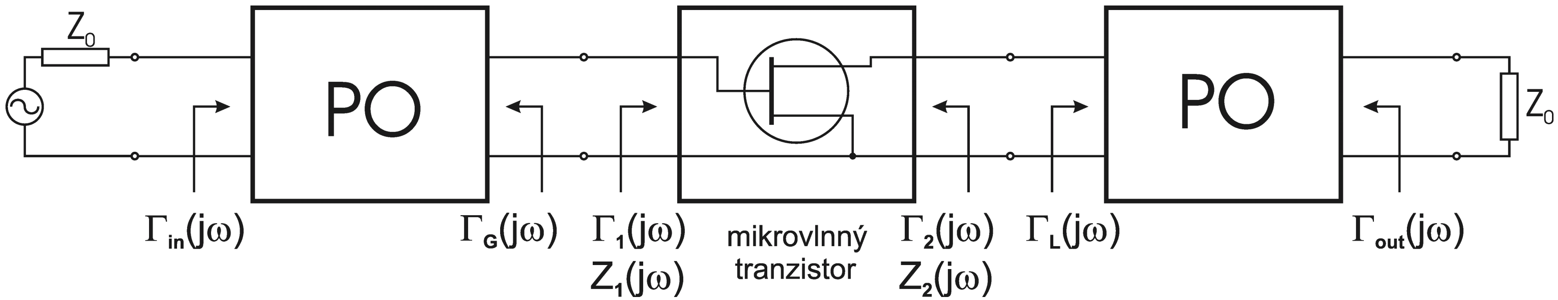


Wide-Band Amplifier Design

- Wide-band amplifiers with loss-less MCs (LC, microstrip, coplanar waveguide, …)
- Typically wideband **LNAs or power amplifiers**
- In the B-wide frequency band only **compromise impedance matching can be reached**

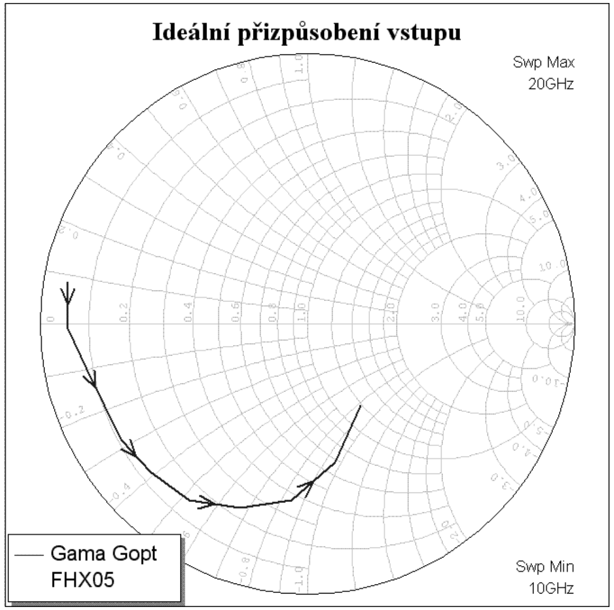
$$\left|\Gamma_{in}(j\omega)\right| \leq const. \leq \Gamma_{in\max} \qquad \left|\Gamma_{out}(j\omega)\right| \leq const. \leq \Gamma_{out\max}$$

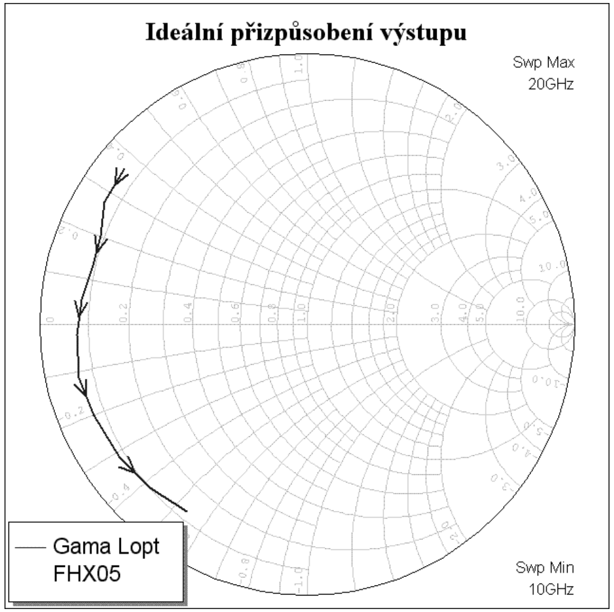
- All design procedures are based on **numerous approximations**, CAD optimization is necessary
- For the given requirements → **NO solution can exist**
- Besides the MC design → **gain equalization** must also be treated
- Wideband **biasing circuits** are necessary
- Substantially more complicated than the narrow-band amplifier design



Wide-Band Amplifier Design

- **AWR-MO:** $\Gamma_{Gopt} = GM1 \qquad \Gamma_{Lopt} = GM2$

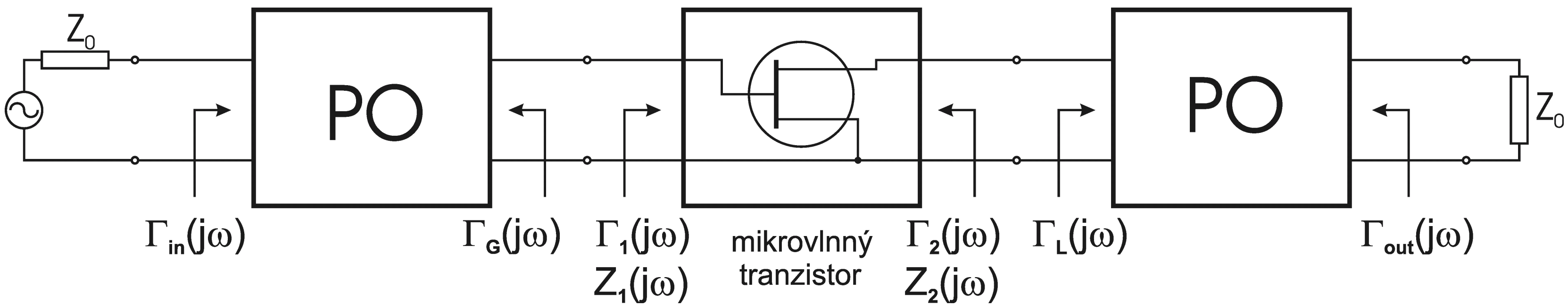




- Frequency plots of Γ_{Gopt} and Γ_{Lopt} run **counter-clockwise**
- Frequency plot of ANY matching circuit run clockwise
- Ideal impedance matching can be reached only at 1 frequency, or at several discrete and distant frequencies
- It is NOT possible to design wide-band amplifiers with ideal impedance matching

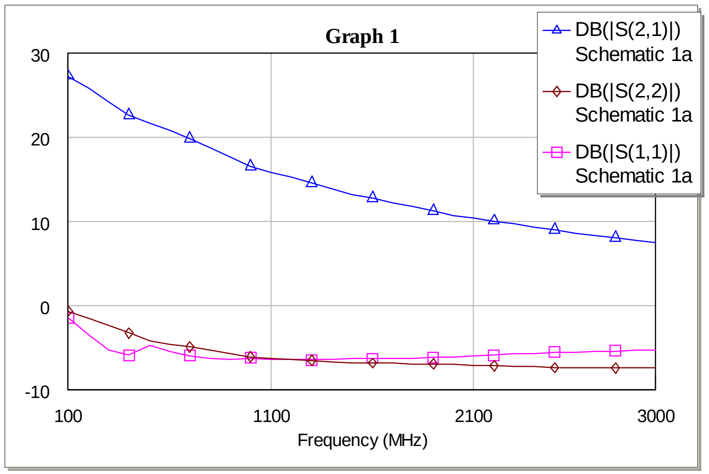
Wide-Band Amplifier Design

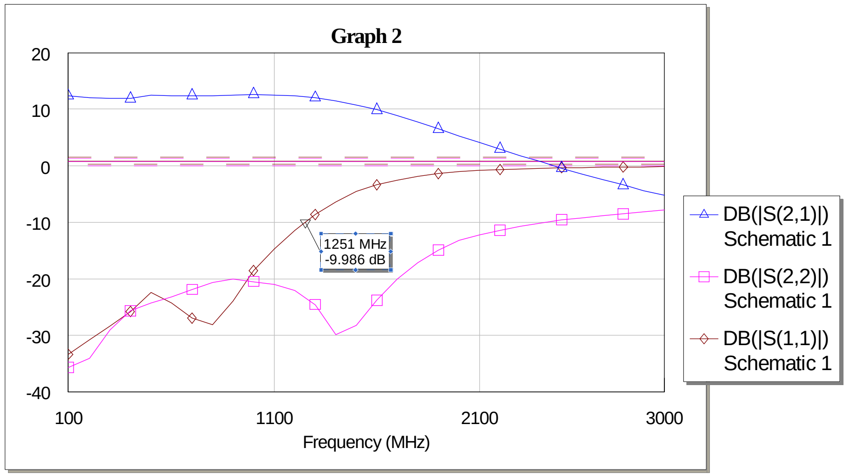
- The problem is enhanced by the fact, that **both input and output MCs must be designed**:
  - The MC_1 transforms Z_0 to $\Gamma_1(j\omega)$ but $\Gamma_1(j\omega) = f(\Gamma_L) = s_{11}(j\omega) + \dfrac{s_{12}(j\omega)s_{21}(j\omega)\Gamma_L(j\omega)}{1 - s_{22}(j\omega)\Gamma_L(j\omega)}$
  - The MC_2 transforms Z_0 to $\Gamma_2(j\omega)$ but $\Gamma_2(j\omega) = f(\Gamma_G) = s_{22}(j\omega) + \dfrac{s_{12}(j\omega)s_{21}(j\omega)\Gamma_G(j\omega)}{1 - s_{11}(j\omega)\Gamma_G(j\omega)}$
- MC_1 and MC_2 must be solved simultaneously → CAD optimization
- Multiple iterations must be performed



Gain Equalization

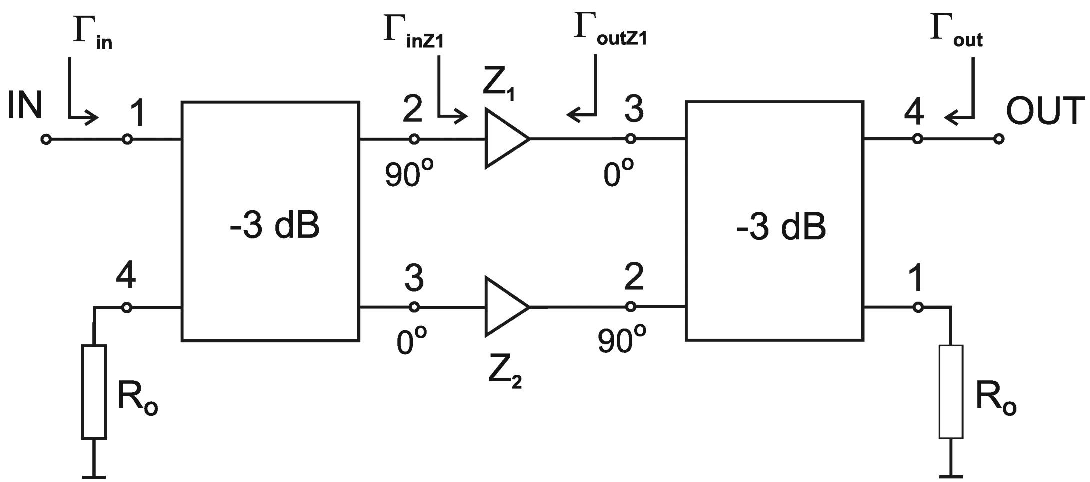
- Transistor gain (available, power) **decreases -6dB/oct**
- Wide-band amplifiers → **constant gain** is required usually
- The drop in gain must be equalized → possible solutions:
  - o **Higher reflections** $\Gamma_{in}(j\omega)$ and $\Gamma_{out}(j\omega)$ at lower frequencies
  - o Passive **equalizing structures** (frequency dependent attenuators)
  - o Special **feedback structures** (cannot be used in case of LNAs and PAs)
- Easier in the balanced structures





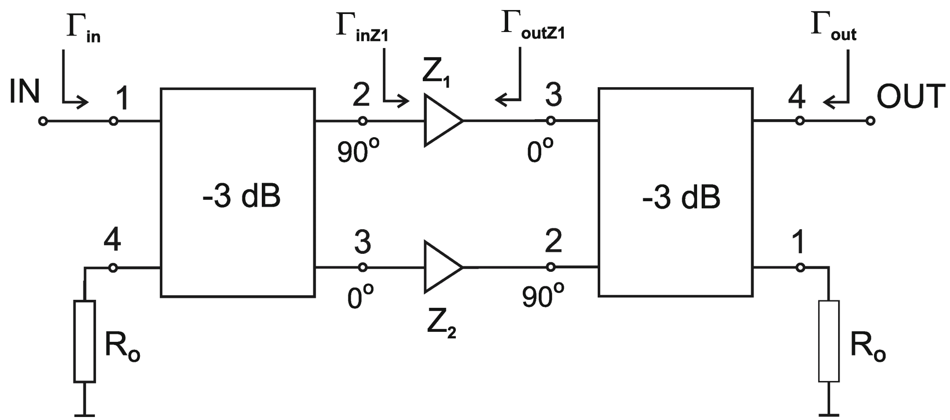
Balanced Amplifiers

- More complex amplifier structure
- Employs 2 identical amplifying stages **operated in parallel**
- Input signal is divided 1:1 by the **90° power divider**
- Quadrature divider or Wilkinson divider with additional λ/4 transmission line can be used
- Dividing loss $L_d = 3dB + L_a$
- Amplified signals are **summed in-phase** with an identical and symetrically connected 90° power splitter/combiner



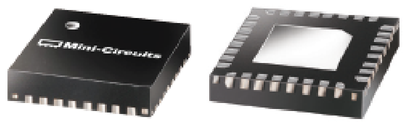
Balanced Amplifiers

- Resulting parameters:
  - **Gain** $G_{resdB} = G_{ZdB} - 2L_{adB} \cong G_{zdB}$
  - **Output power** $P_{resdBm} = P_{ZdBm} + 3dB - L_{adB}$
  - **Input reflections** $\Gamma_{in} = \frac{1}{2}(\Gamma_{inZ1}e^{2j\pi/2} + \Gamma_{inZ2}e^{2j0}) = \frac{1}{2}(-\Gamma_{inZ1} + \Gamma_{inZ2})$
  - **Output reflections** $\Gamma_{out} = \frac{1}{2}(\Gamma_{outZ1}e^{-2j0} + \Gamma_{outZ2}e^{-2j\pi/2}) = \frac{1}{2}(\Gamma_{outZ1} - \Gamma_{outZ2})$
  - If $\Gamma_{inZ1} = \Gamma_{inZ2}$ and $\Gamma_{outZ1} = \Gamma_{outZ2}$, **input and output reflections can be very small**
  - The structure enable reduction of reflections even in case of LNAs or other highly reflecting amplifiers



Balanced Amplifiers

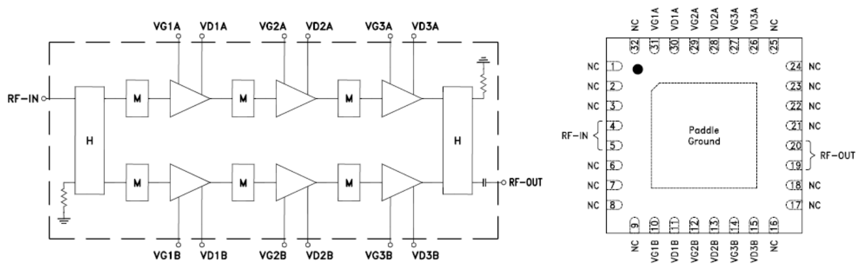
- **Advantages:**
  - Higher output power, low input/output reflections
  - Higher reliability
  - Wideband stability
  - Low sensitivity to parameters of the from outside connected components
  - Highly recommended solution
- **Disadvantages:**
  - Larger-sized
  - Double the DC power consumption
  - Higher price

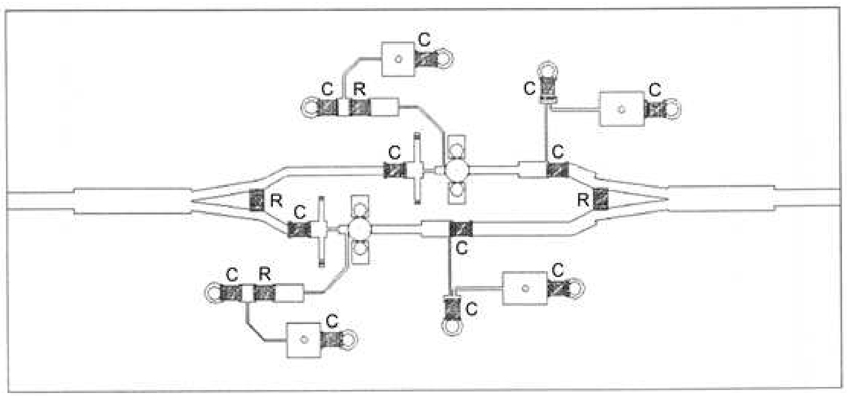


CASE STYLE: DG1677-1
PRICE: $27.95 ea. QTY. (10)

AVM-273HP+

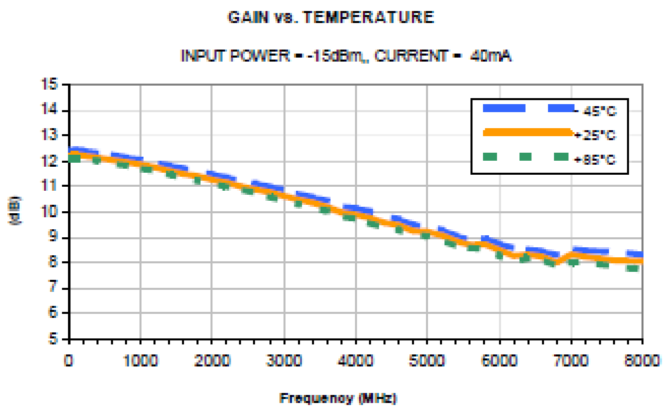
Simplified Schematic and Pad Description

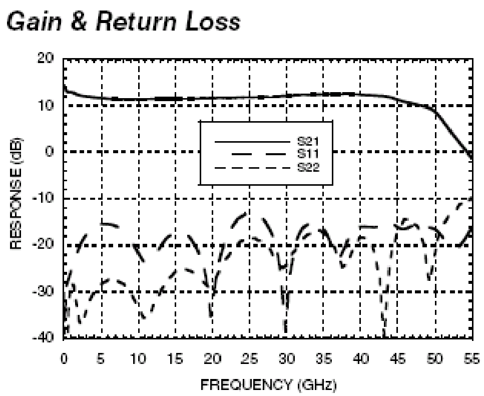


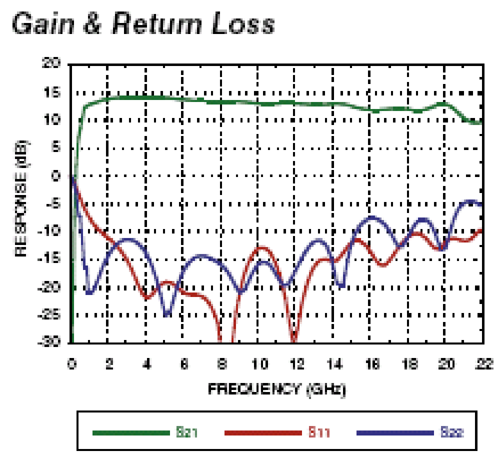


Ultra-Wideband Amplifiers

- Standard wide-band impedance matching provides approx. 1-2 octaves bandwidth
- Even wider amplifiers are required (DC-8GHz, 1 – 20GHz, DC- 35GHz, …)
- For these purposes, special matching techniques must be used:
  - o Lossy impedance matching
  - o Feedback structures
  - o Distributed structures
- Available parameters:
  - o Wide bandwidth
  - o Flat gain, fair input and output reflections, typ. RL≥10dB
  - o But often – higher F



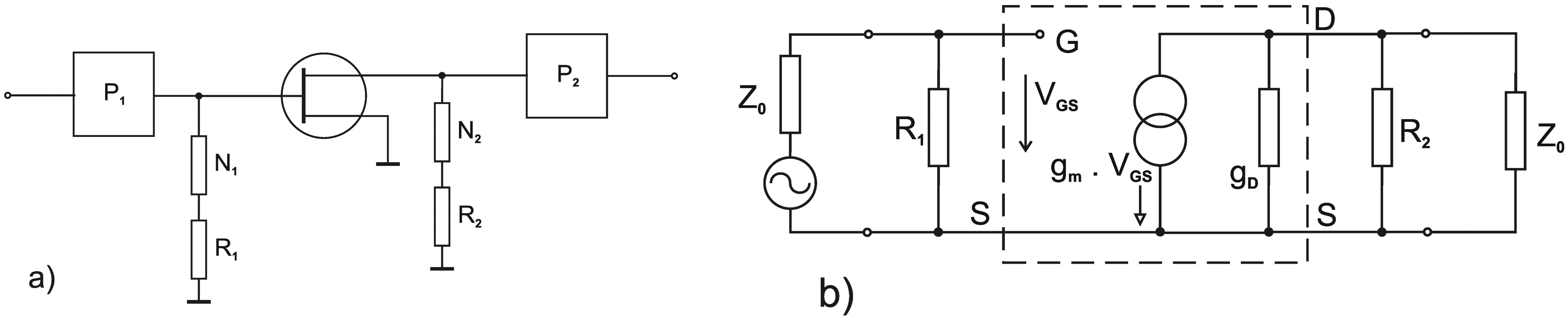




Amplifiers with Resistive Matching

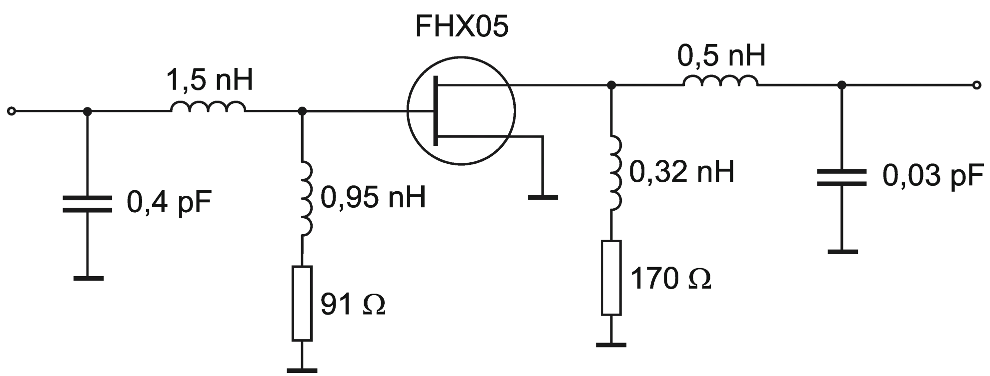
- Employ R_1 and R_2 **resistors in the input and output matching circuits**
- The N_1 and N_2 frequency dependant components (L usually) reduce influences of R_1 and R_2 at higher frequencies →contribute to **gain equalization**
- Additional simple matching circuits P_1 and P_2 improve wide-band impedance matching
- Amplifiers of this type are designed by CAD optimization
- The initial R_1 and R_2 values and gain can be calculated from the LF model:

$$R_1 \cong Z_0 = \frac{1}{G_0} \qquad R_2 \cong \frac{1}{G_0 - g_D} \qquad G \cong \left(\frac{g_m}{2G_0}\right)^2$$



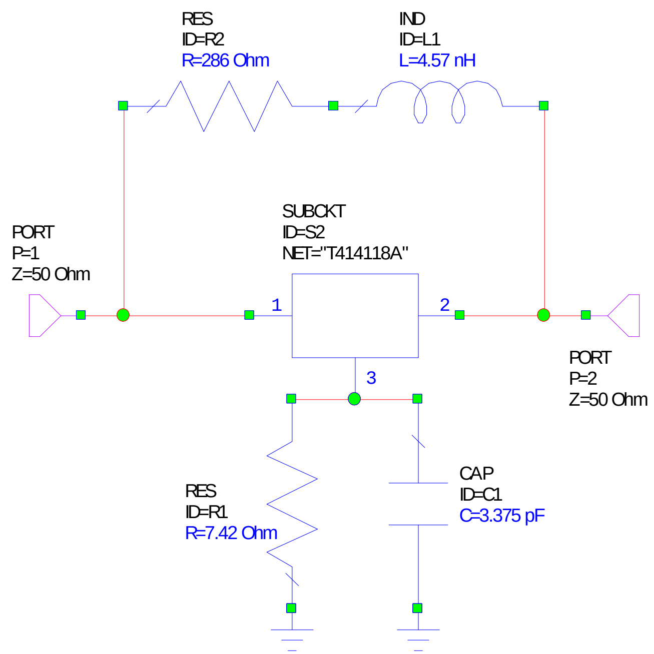
Properties - Example

- **Properties** of amplifiers with resistive matching:
  - Simple structure
  - Low frequency is limited only by the blocking capacitors used
  - But high noise figure, low output power and PAE, strong dependence of G on g_m
- **Example:**
  - Design DC – 6 GHz, HEMT, AWR optimization
  - G=6,2dB @ 1GHz, G=8,9dB @ 6GHz
  - $RL_{in} \geq 10$dB, $RL_{out} \geq 10$dB @DC – 6GHz



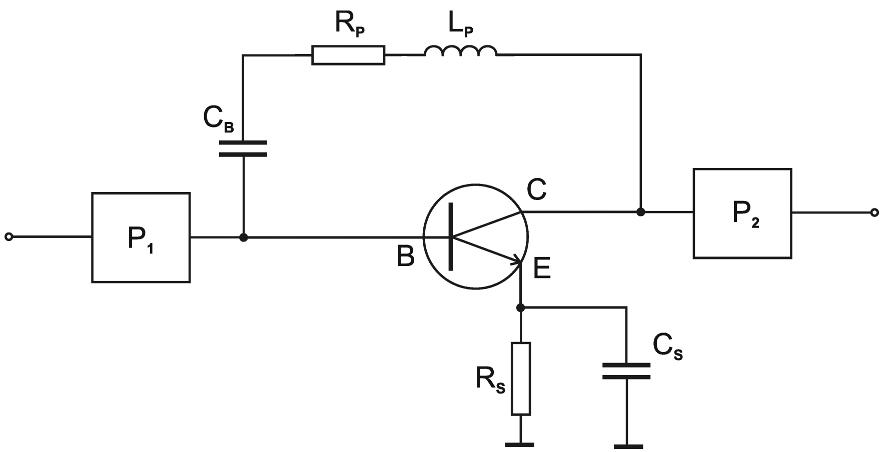
Feedback Amplifiers

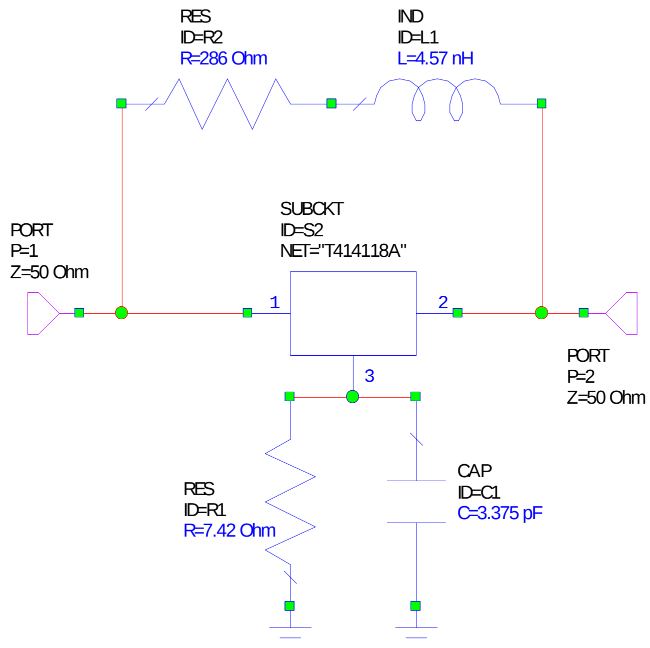
- Employ **resistive feedback with additional frequency dependent components**
    - Series feedback → BJT
    - Parallel feedback → FET
    - Combination of both → BJT
- Additional frequency dependent components are used for **gain equalization**
- Feedback structures:
    - Ensure wide-band input and output matching
    - Ensure flat gain
    - Improve stability
    - But raise F
    - MMIC – the most frequent amplifier type
- Design → CAD optimization



Feedback Amplifiers - BJT

- BJT → both **series and parallel negative resistive feedbacks used**
- C_s parallel to R_s reduces influence of the series feedback at higher frequencies
- L_p in series with R_p reduces influence of the parallel feedback at higher frequencies
- P_1, P_2 → simple additional matching circuits – optional
- Common values:
  - o $R_S=10\Omega$
  - o $R_p=250\Omega$
- Upper frequency → HBT MMIC DC-8GHz



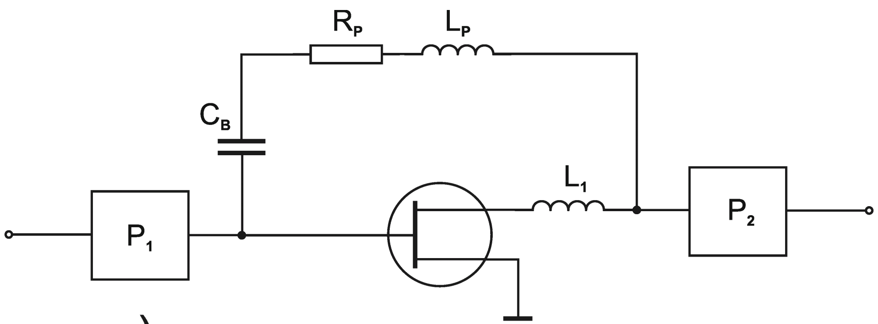


Feedback Amplifiers - FET

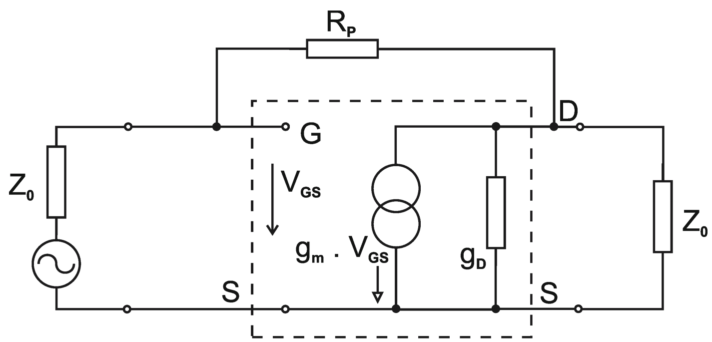
- FETs → **only parallel feedback used**
- Initial values for the CAD optimization can be derived from the LF model

$$R_p \approx \frac{g_m - g_d}{G_0 (G_0 + g_d)} \qquad G \approx \left(\frac{G_0 - g_m}{G_0 + g_D} \right)^2$$

- Common values: $g_d = 0{,}0025$ $g_m = 0{,}007$ lead to $R_p = 150\ \boldsymbol{\Omega}$ $G = 6{,}9\ dB$



a)

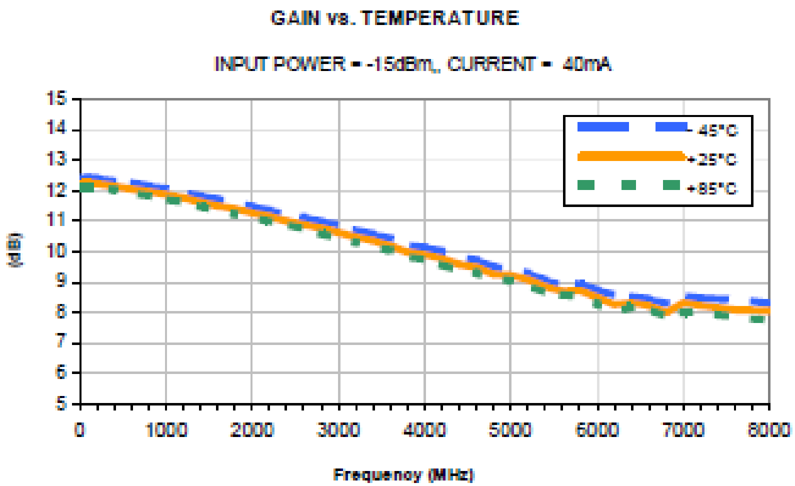


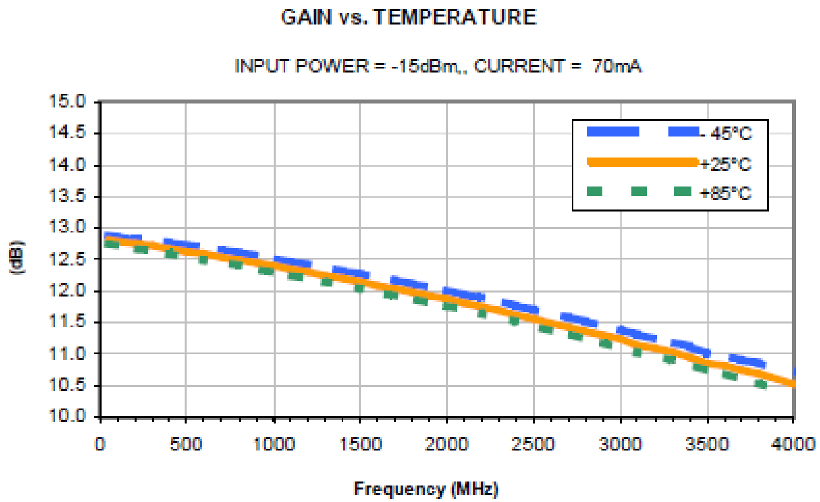
b)

Example 1

- Several Mini-Circuits wide-band MMIC amplifiers:

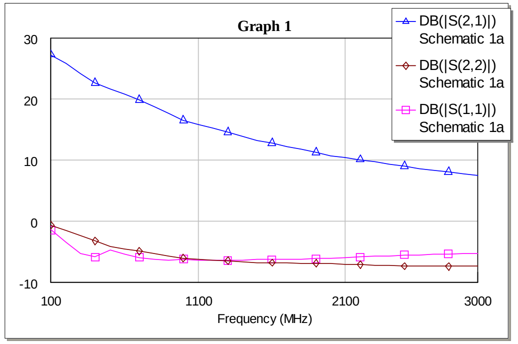
Type	band B $[GHz]$	gain G $[dB]$	P_{-1dB} $[dBm]$	F $[dB]$	V_N $[V]$	I_N $[mA]$
ERA-1	$DC \div 8$	11,4	11,7	5,3	3,6	40
ERA-2	$DC \div 6$	15,2	12,8	4,7	3,6	40
ERA-3	$DC \div 3$	21,3	12,1	3,8	3,5	35
ERA-4	$DC \div 4$	13,5	17,0	5,5	5,0	65
ERA-5	$DC \div 4$	18,6	18,4	4,5	4,9	65
ERA-6	$DC \div 4$	11,4	18,5	8,4	5,5	70

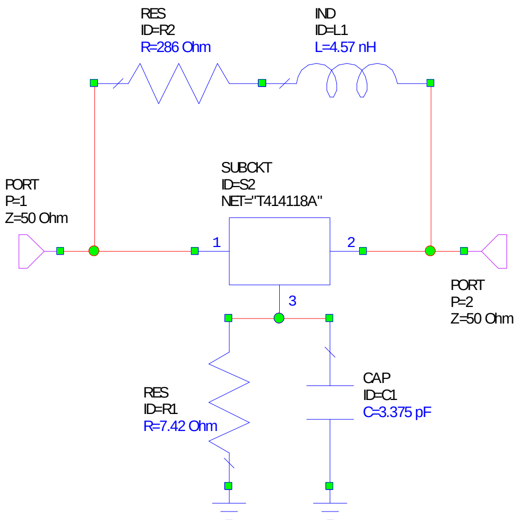


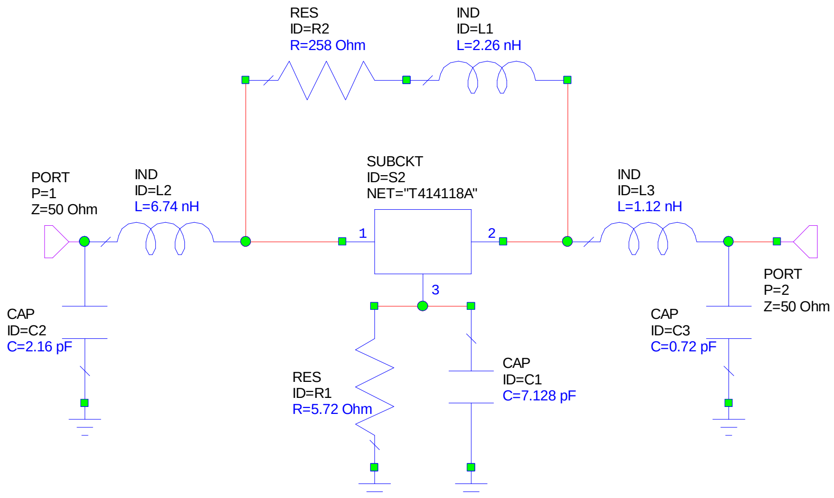


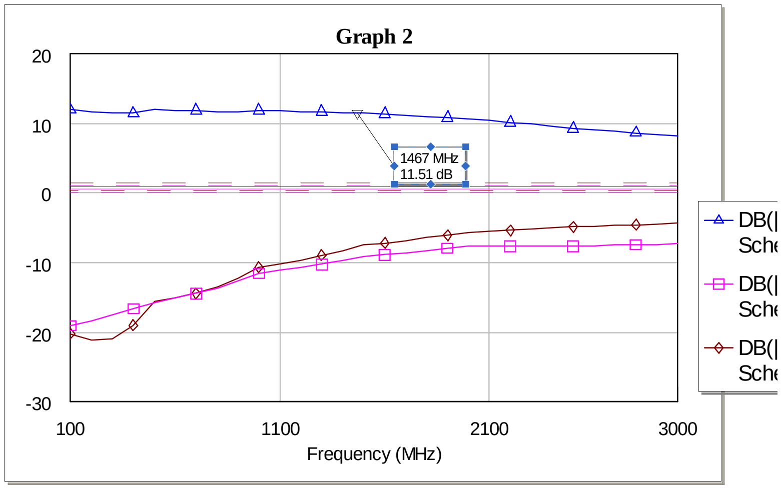
Example 2

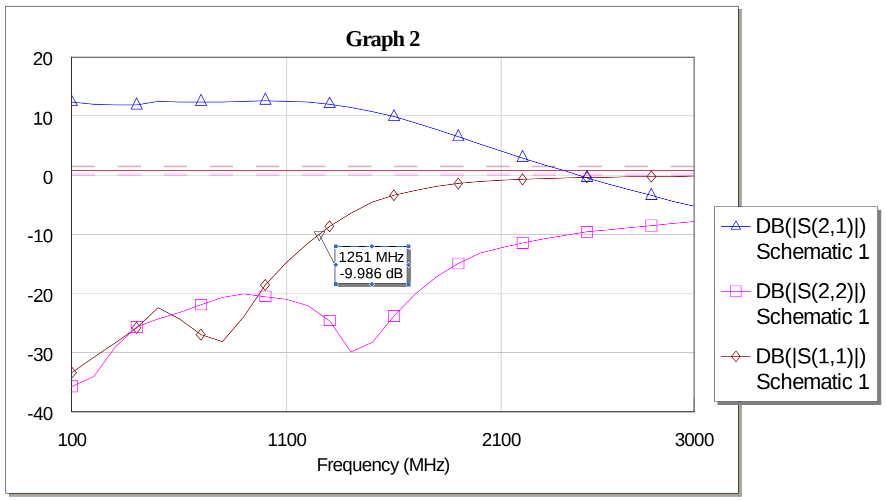
- **Individual task No.6**
- BJT AT-41411, 8V/10mA





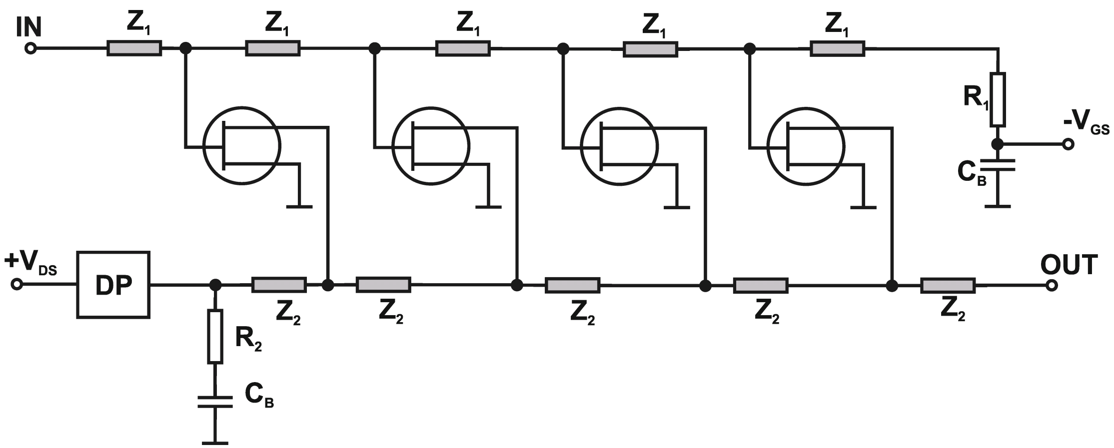






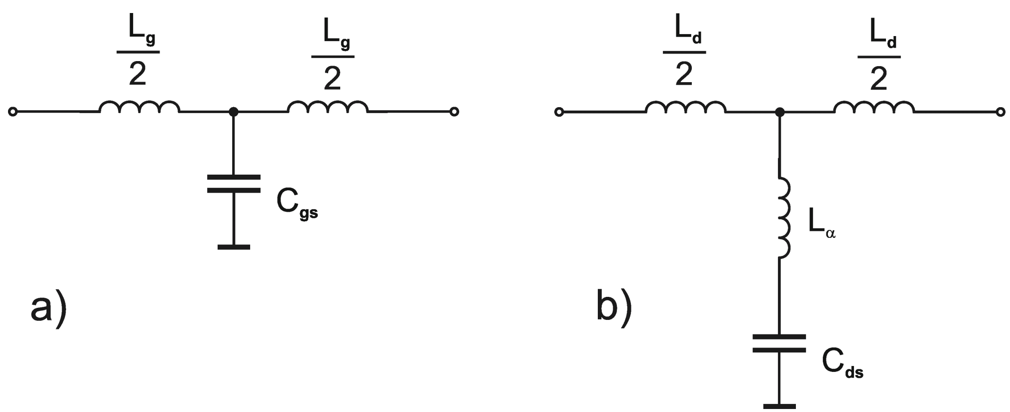
Distributed Amplifiers

- Truly **wide-band components/structures:**
  - Ideal resistors
  - Matched transmission lines
- Distributed amplifiers employ wideband capabilities of the **matched transmission lines**
- The **input transmission line** distributes the input RF power to FETs
- The rest of the input wave is absorbed in the R_1 termination
- The **output transmission line** collects in phase the amplified RF powers from the FET outputs



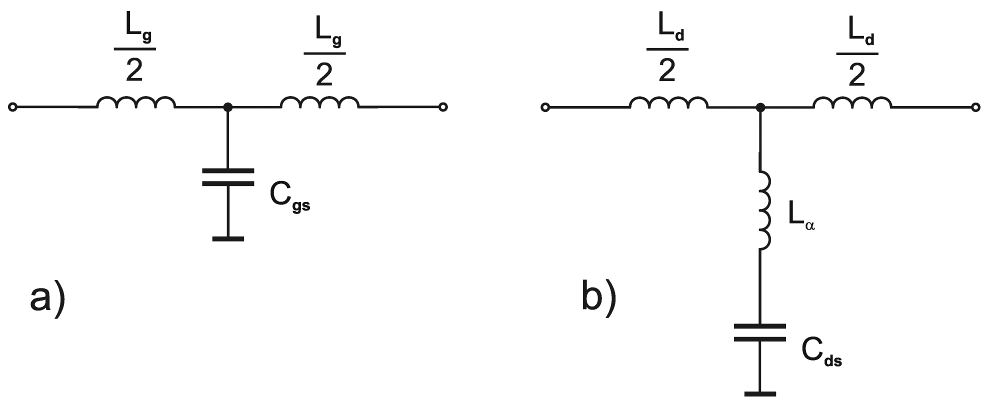
Artificial Transmission Lines

- Extremely wide-band amplifiers → can be obtained by realizing **artificial transmission lines and by incorporating FET parasitic reactances into them:**
  - Artificial transmission lines are formed by the equivalent **LC structures**
  - The input transmission line is formed by a cascade of external series inductances $L_g/2$ and FET **input parasitic capacitances C_{gs}**
  - The output transmission line is formed by a cascade of external series inductances $L_d/2$ and FET **output parasitic capacitances C_{ds}**
- **Problem:** $C_{gs} >> C_{ds}$, while impedances $Z_{in} = Z_{out} = Z_0 = 50\Omega$ and phase velocities in both transmission lines must be the same
- **Solution:** Compensating inductance L_α



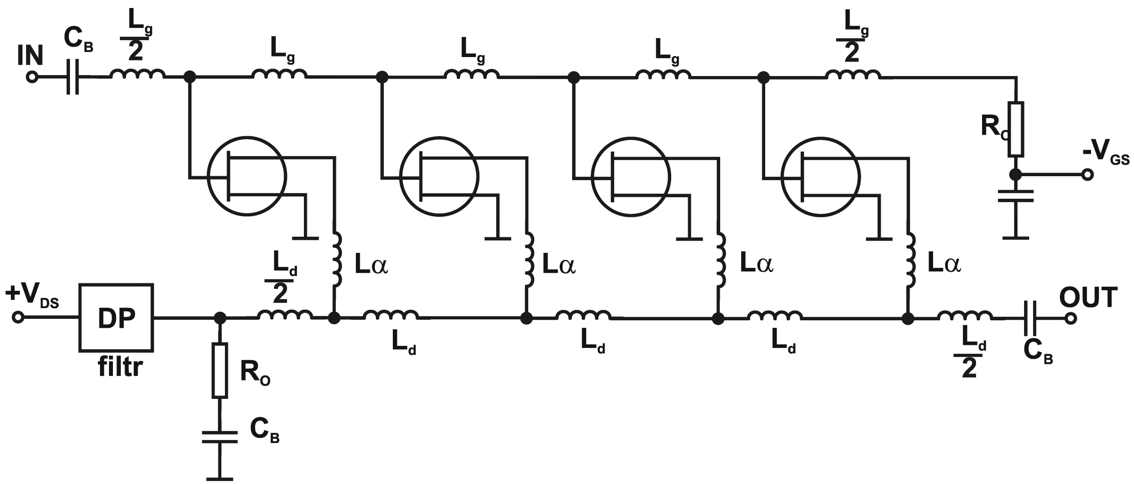
Artificial Transmission Lines

- Artificial transmission lines:
  - Enable realization of the distributed amplifiers (DAs)
  - Are not ideal transmission lines → only equivalent circuits
  - Behave like **low-pass filters and limit the DA upper frequency** $f_m = \frac{1}{\pi\sqrt{LC}}$
- DA gain:
  - Depends upon number of FET sections $G = \frac{n^2 g_m^2 Z_0^2}{4}$
  - The formula considers the loss-less transmission lines
  - In practice – both lines are lossy, input power drops alongside the cascade
- Practical No. of FET sections → **4 - 5**



Practical DA Circuit

- Practical DA circuit:
  - Four FET sections
  - **External inductances L_g, L_α, L_d** → their parasitic capacitances limit the DA upper frequency
  - **Gate biasing** (zero current) – through the R_0 termination
  - **Drain biasing** (high current) – through the low-pass filter
  - The LP filter limits the lower operating frequency
- Realization → MMIC (with external powering circuits)



Example

- HITTITE (www.hittite.com) :
  - Amplifiers – Wideband (Distributed)
  - Tens of models, also LNAs and PAs
  - Frequency ranges: DC-20GHz, 2-22 GHz, DC-35GHz, 0,5-65 GHz
  - Usually chips - but also in the LP5 package (DC-20GHz)