Linear amplifiers

Microwave Electronics

Giovanni Ghione, Marco Pirola
Politecnico di Torino, DET

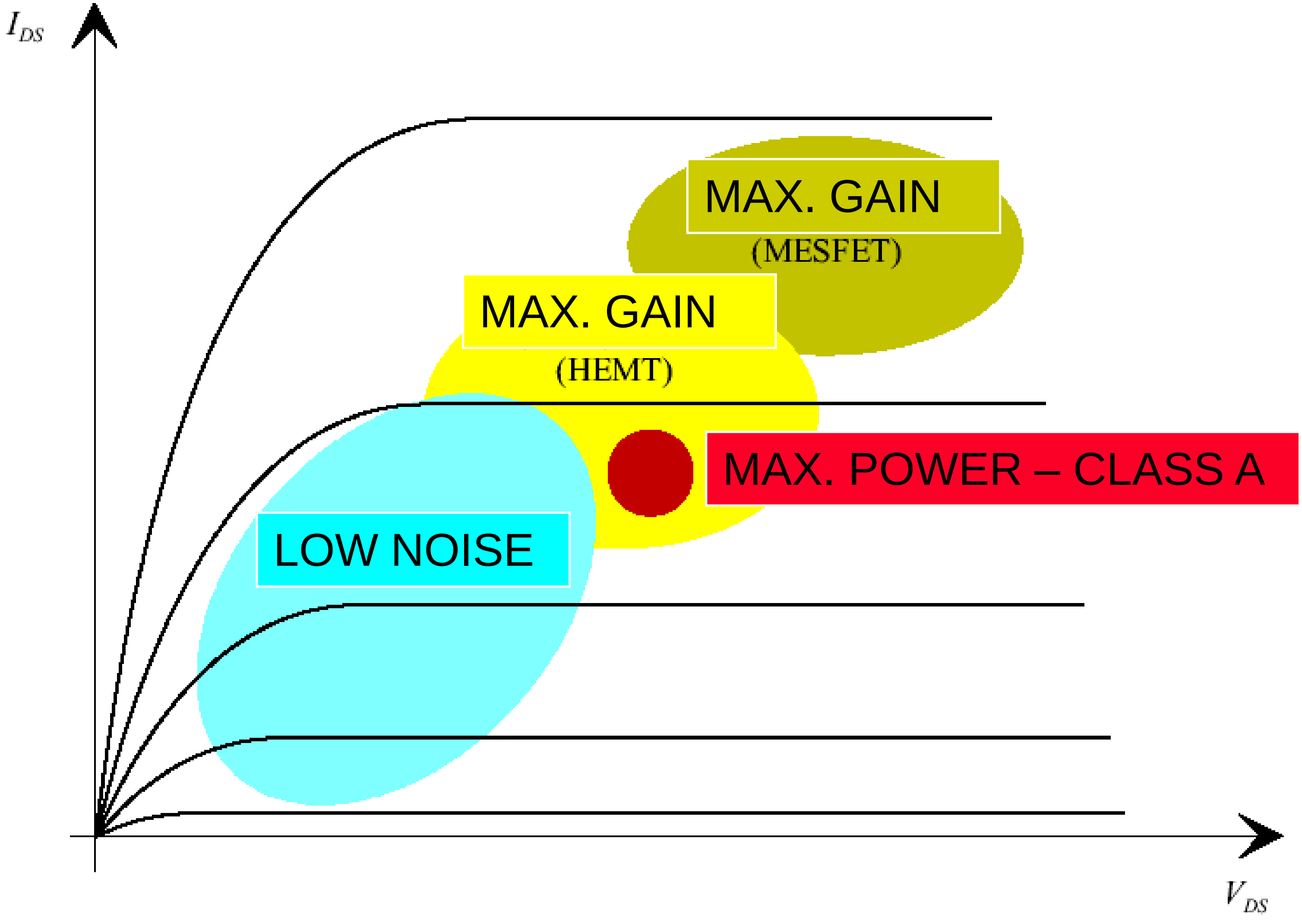
RF amplifier choices

- **Purpose:** max. gain $\leftrightarrow$ LNA $\leftrightarrow$ PA
- **Topology:** single stage $\leftrightarrow$ multistage, open loop $\leftrightarrow$ feedback
- **Technology:** hybrid $\leftrightarrow$ integrated
- **Bandwidth**: Narrowband (<10%) $\leftrightarrow$ wideband (<40%) $\leftrightarrow$ ultrawideband
- **Significant examples**:
  – open-loop, narrowband amplifier, reactively matched
  – wideband balanced amplifier
  – wideband parallel feedback amplifier
  – ultrawideband distributed amplifier
  – narrowband LNA with inductive series feedback

An RF evergreen: open loop, narrowband, maximum gain amplifier design

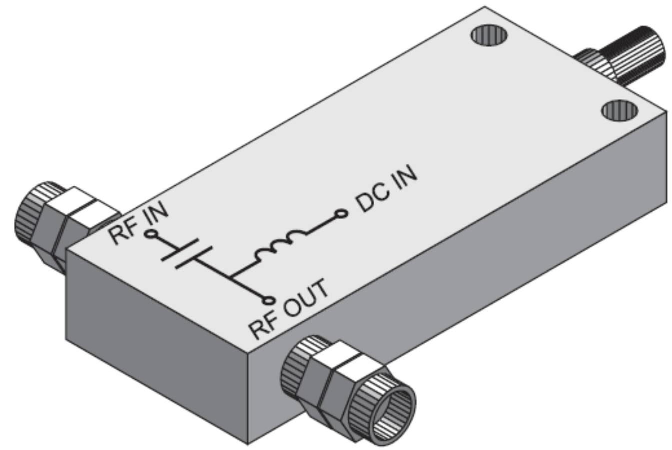
- Narrowband (< 10% f_0) design can be carried out virtually as single frequency design + some further optimization
- The design flow can be summarized as follows:
  - *Choose the active device (e.g. cutoff frequency >2-4 x f_0)*
  - *Choose a suitable bias point and design the bias network*
  - *Make the active device unconditionally stable*
  - *Evaluate the optimum loads for maximum gain*
  - *Design matching networks*
  - *Generate the circuit schematic and layout*
- Broadbanding is possible, but up to a certain point (see later)!

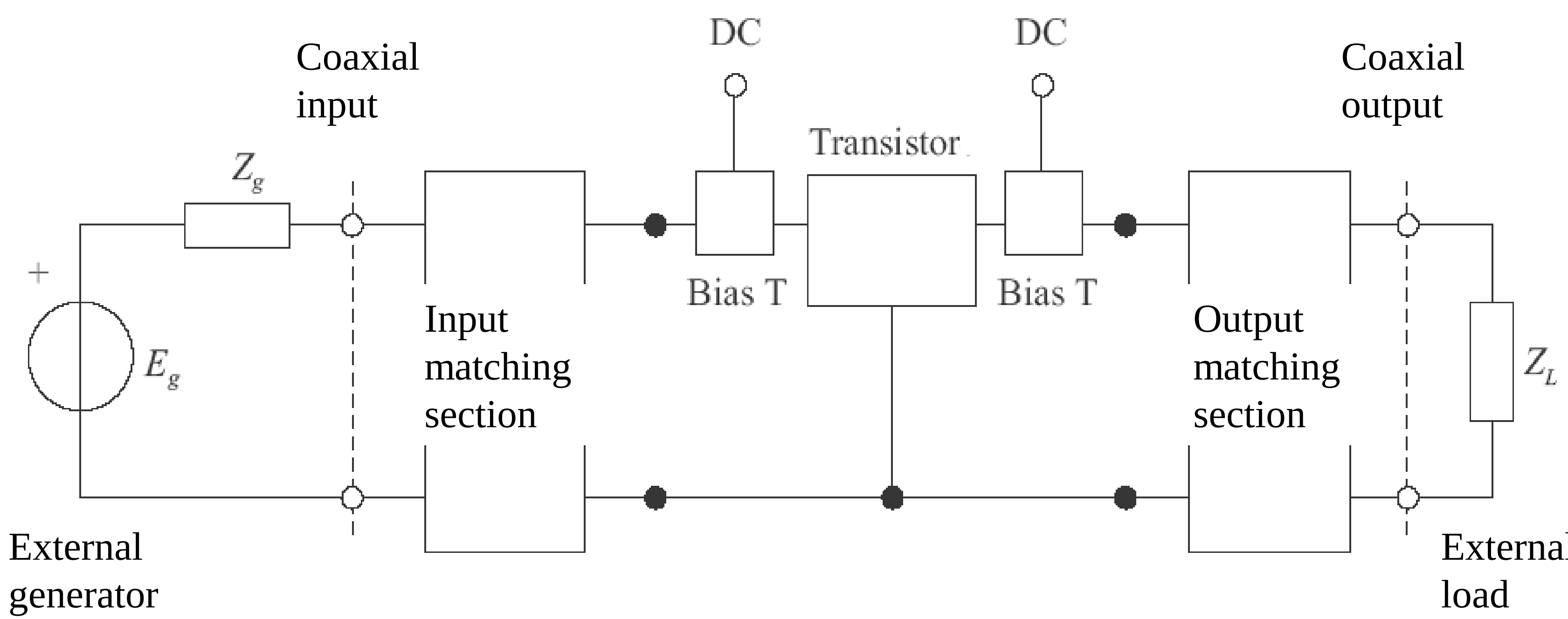
Amplifier “typical” bias points



Device biasing: active device bias Ts

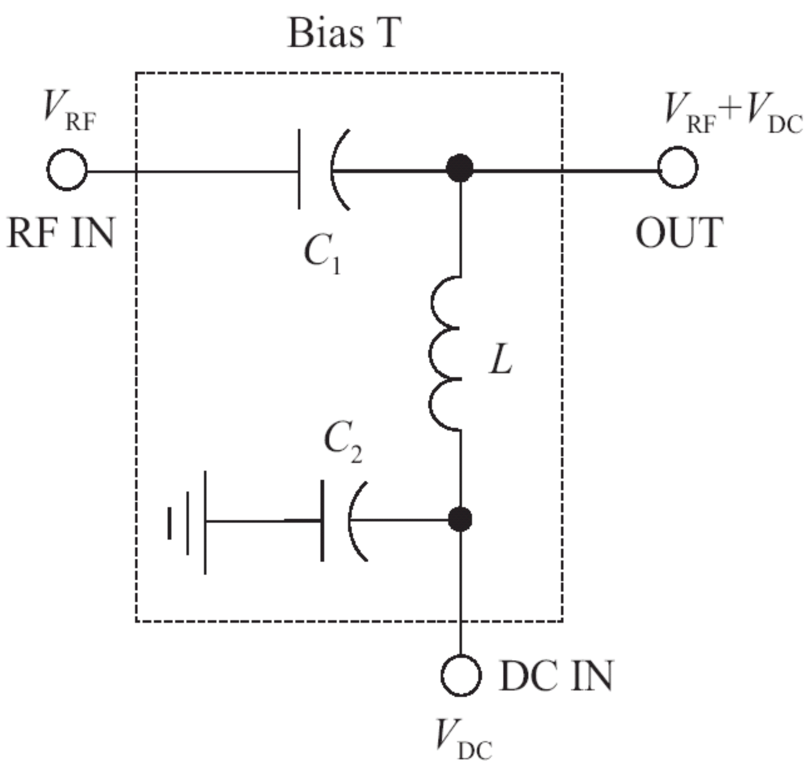
- The bias network has to be decoupled from the RF circuit through proper components, called *Bias Tees (Bias Ts).*
- Bias T wideband design (e.g. for instrumentation) critical

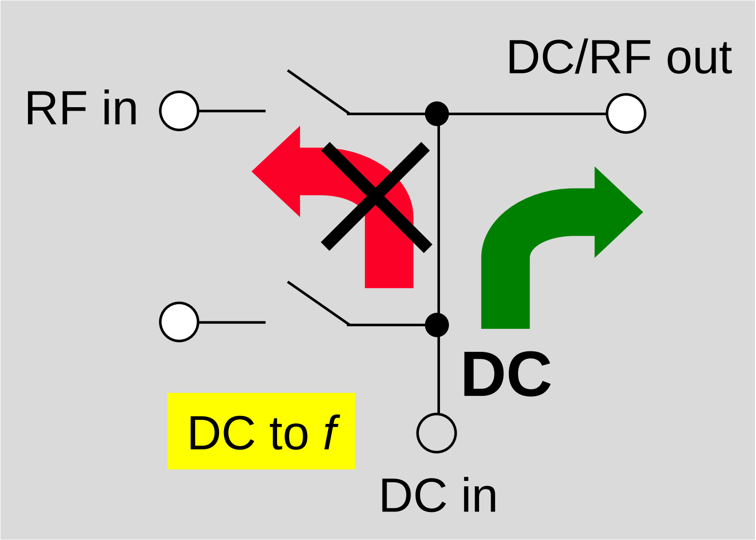


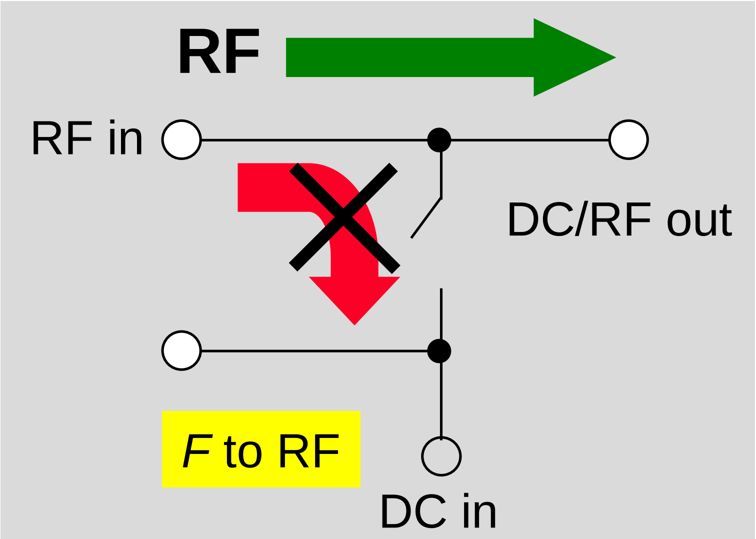


Lumped Bias T

- A tripole with two inputs (RF IN, DC IN) amd one output (RF+DC OUT).
- In practice DC→ from 0 to *f*, RF → from *F* to infinity; if *f* and *F* are low the design requires very high C and L values





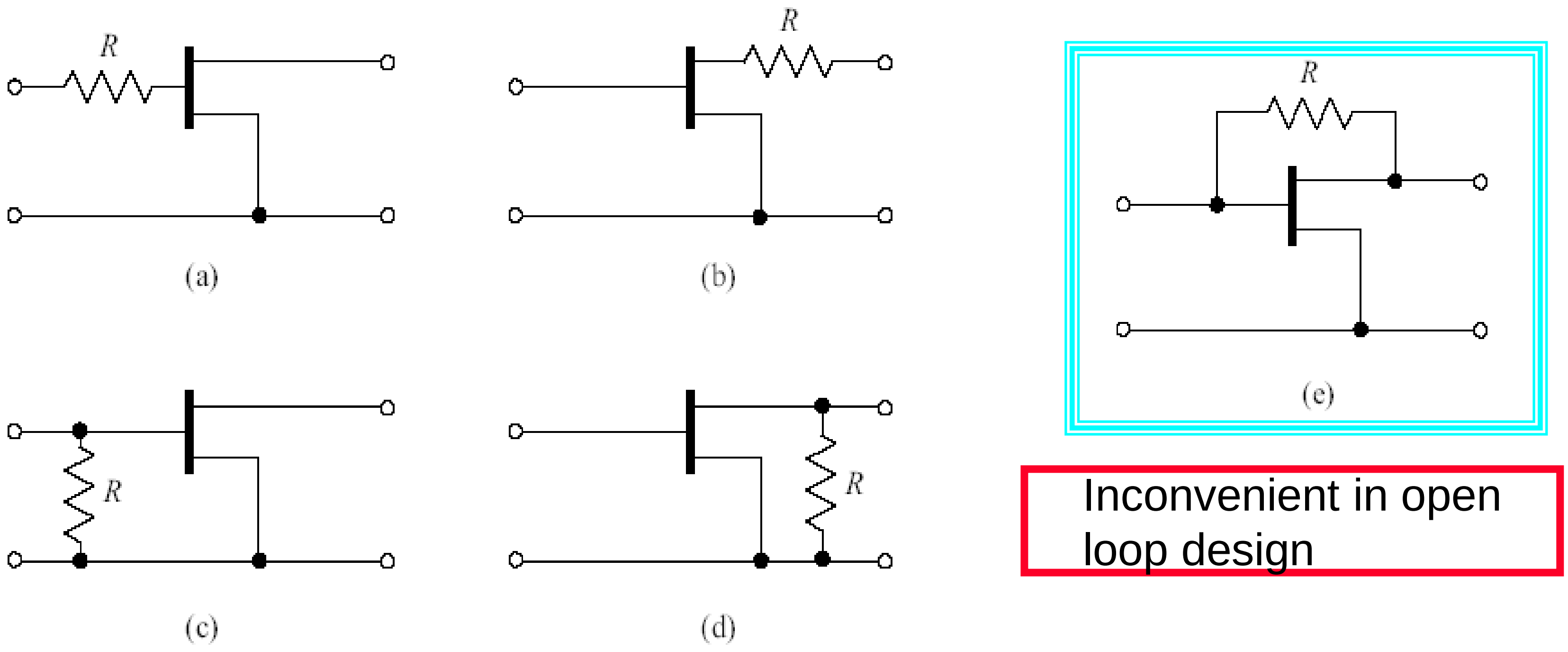


Making active devices unconditionally stable

- RF transistors have a decreasing gain with frequency → they are potentially unstable at least at low frequency, if not in the design bandwidth.
- Stabilization from DC to the design bandwidth is mandatory since there is no control on the loads
- In-band stabilization can be avoided by *proper load and generator design* but is usually strongly advised for easier design, above all in microwave monolithic ICs
- Above the design bandwidth the device becomes asymptotically stable.
- Since the amount of stabilization needed decreases with frequency, proper resistive network with reactive bypasses or blocks have to be used.

Stabilizing active devices

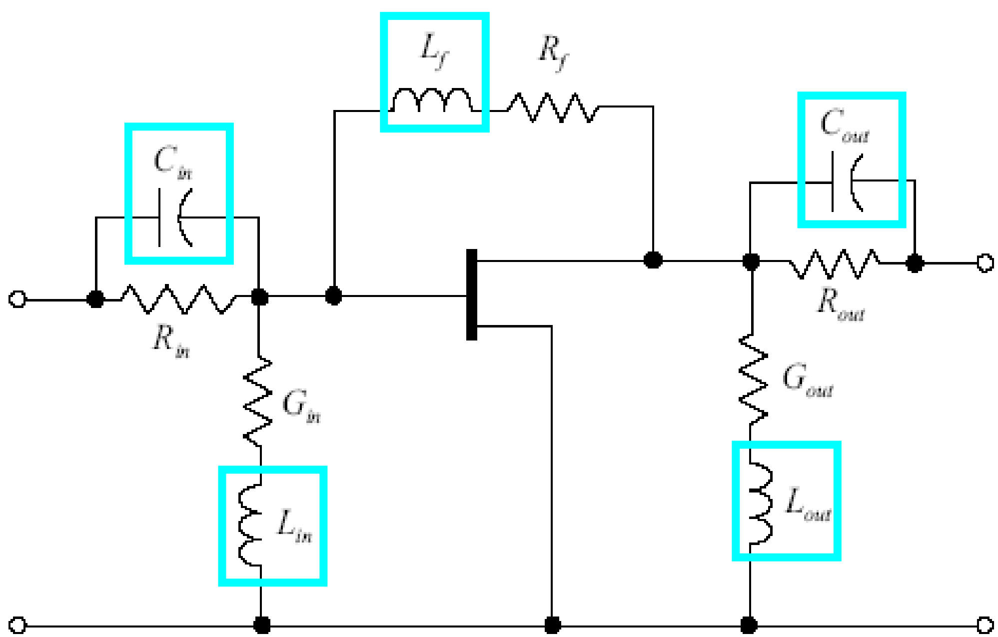
- Stabilization can be obtained through
  - series/parallel input resistors [not for LNA → noise at input!]
  - series/parallel output resistors
  - feedback resistors [non convenient, unless for feedback amplifiers]
- **Mandatory at low frequency**, optional in band



Inconvenient in open loop design

Stabilizing through dissipative elements

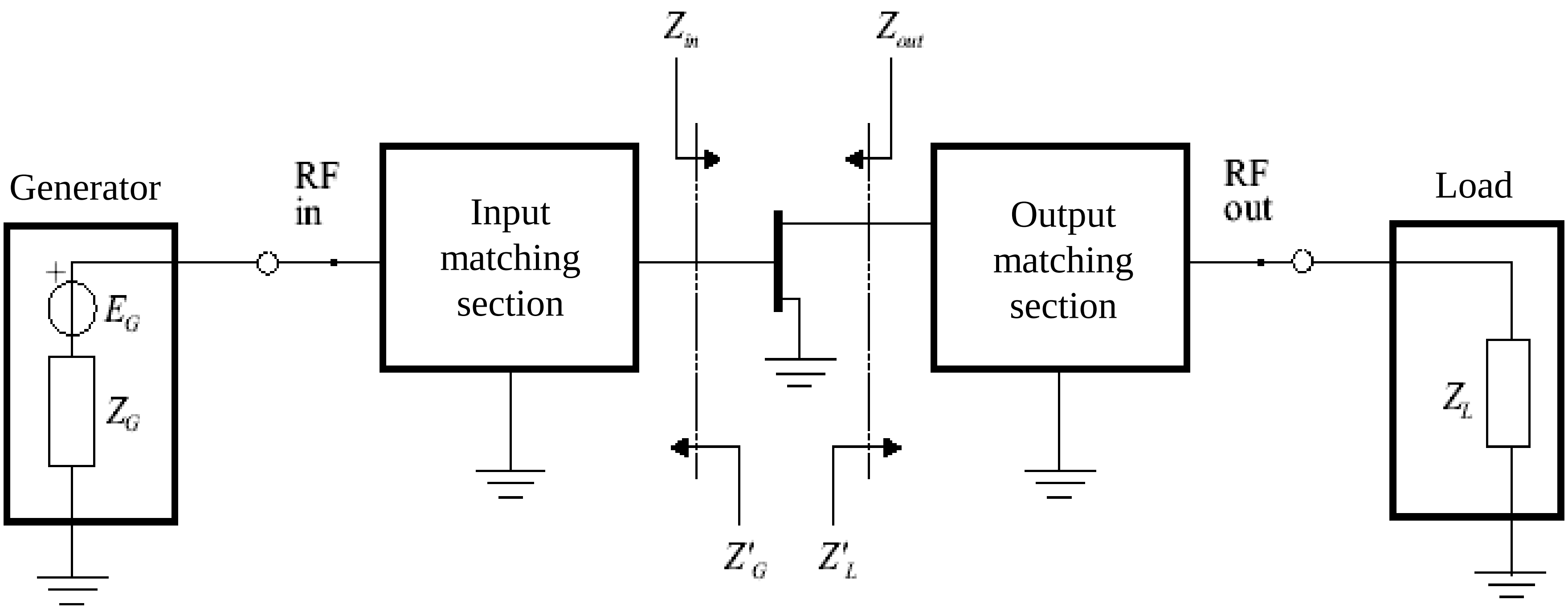
- A combination of reactive and resistive elements can be exploited to obtain the required degree of stabilization at low frequency (strong) and at high frequency (light)
- This can be obtained through numerical optimization of *K*; too strong in-band stabilization compromises gain
- Often just *one* resistor + reactive element is enough (C preferred)



- Sometimes the stabilization networks can be integrated with the Bias Ts
- In the feedback loop also a DC block is needed!

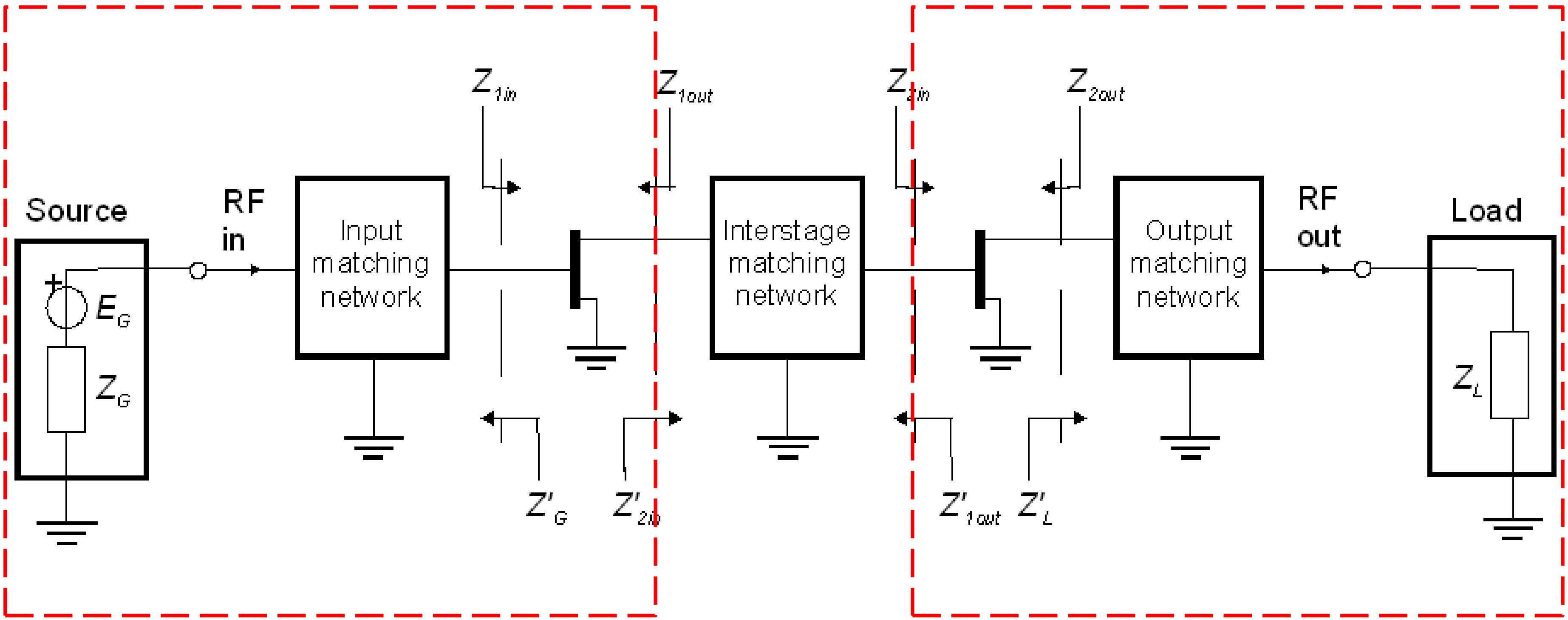
Matching networks

- Needed to transform the load and generator impedances (e.g. 50 Ω) in the device optimum load
- Reactive match: narrowband (why?)
- Resistive or lossy match: more wideband but noisy



Multistage interstage matching

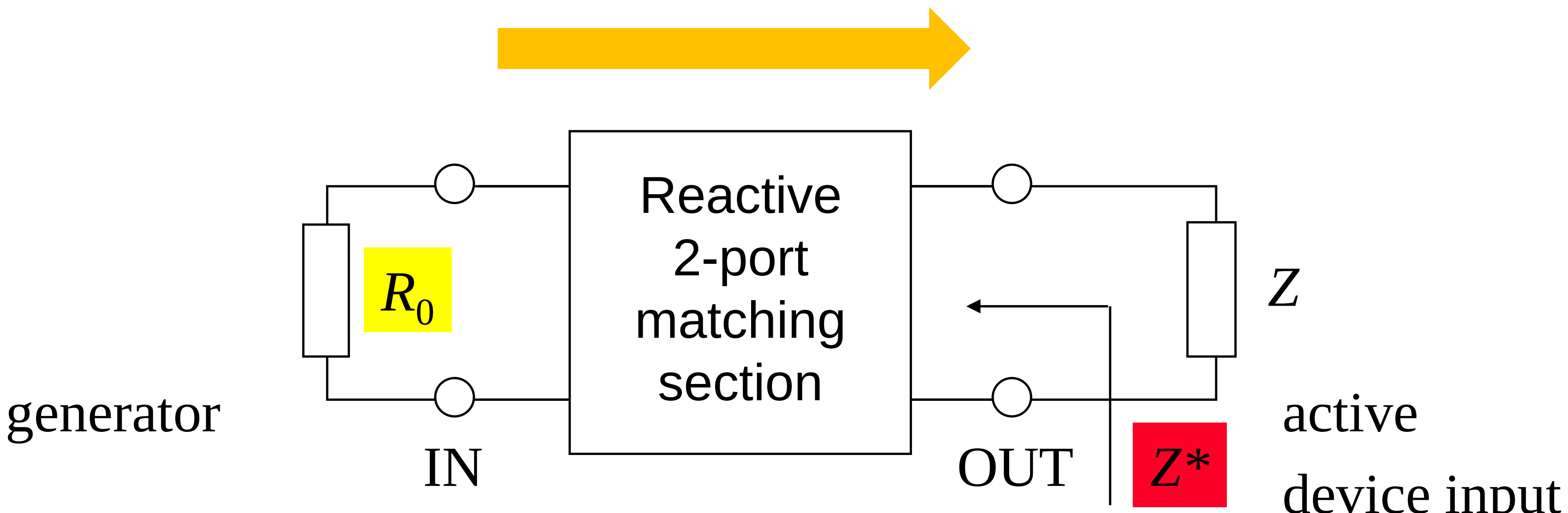
- The interstage matching section transforms into each other the complex impedances allowing for conjugate matching of the e.g. first and second stage
- Typically optimization required!



Two-stage design recipe

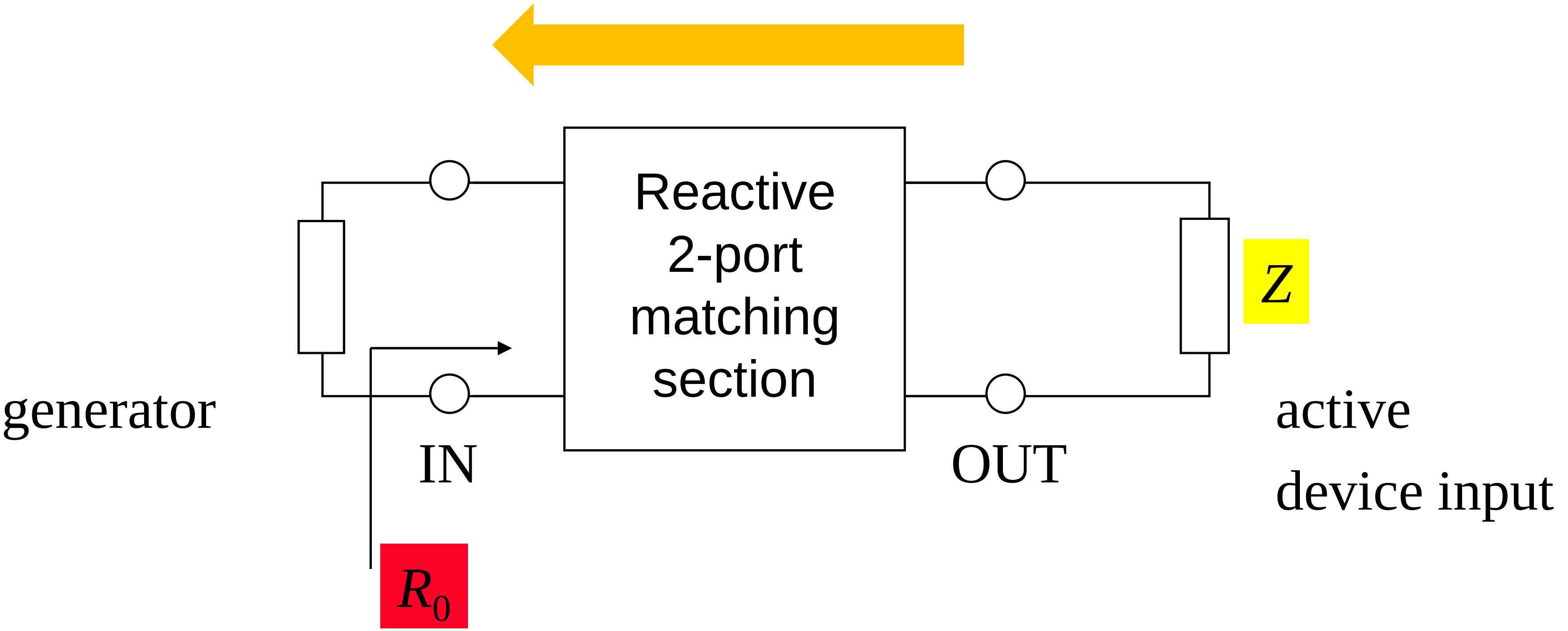
- The optimum load of the two stage is known at a given frequency starting from the S-parameters
- The output stage matching section transforms $Z_L \rightarrow Z'_L$
- The input stage matching section transforms $Z_G \rightarrow Z'_G$
- Since conjugate matching holds in the chain we have: $Z_{2in} = Z^*_{1out}$
- The inter-stage matching stage transforms the complex impedances $Z_{2in} \rightarrow Z'_{2in} = Z^*_{1out}$
- Difficulty: the input and output matching sections design amount to transform a real into a complex impedance, the interstage two complex impedances into each other (see next slides).
- The design can also carried out by optimization.

Design of lossless (reactive) matching networks



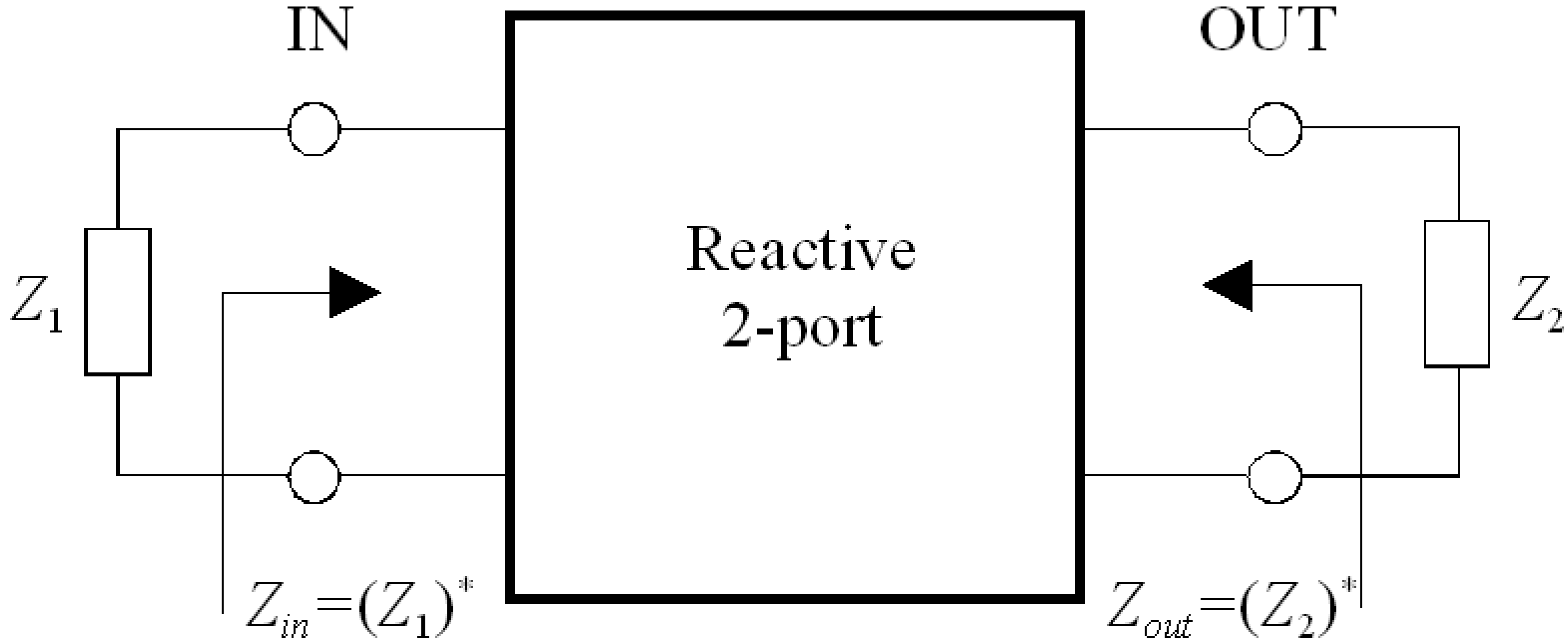
- Lumped- or distributed-parameter implementation

Design of lossless (reactive) matching networks

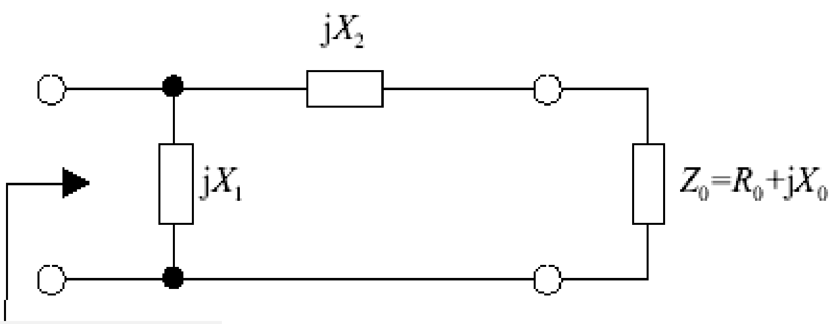


- Lumped- or distributed-parameter implementation

In general…



Example of lumped reactive matching sections



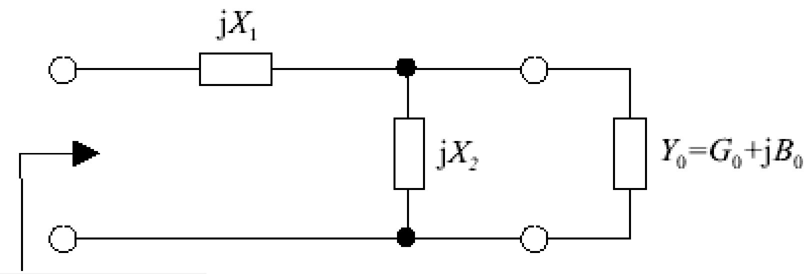
$G_M=1/50$, $B_M=0$

(a)

$$X_1 = \left(-B_M \mp \sqrt{\frac{G_M}{R_0}(1-G_M R_0)}\right)^{-1}$$

$$X_2 = -X_0 \pm \sqrt{\frac{R_0}{G_M}(1-G_M R_0)}.$$

A $\quad G_M R_0 < 1$



$R_M=50$, $X_M=0$

(b)

$$X_1 = X_M \pm \sqrt{\frac{R_M}{G_0}(1-R_M G_0)}$$

$$X_2 = \left(B_0 \mp \sqrt{\frac{G_0}{R_M}(1-R_M G_0)}\right)^{-1}$$

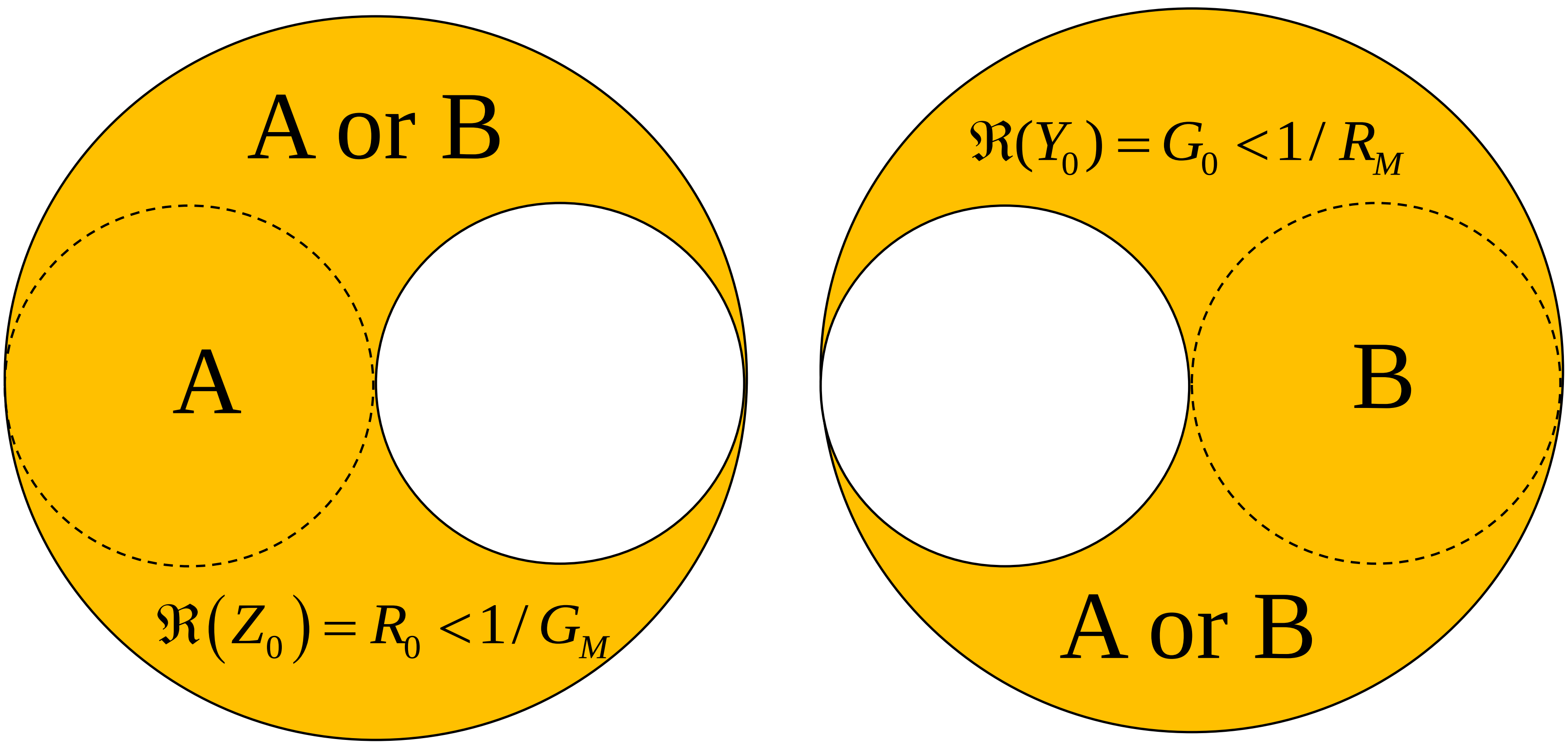
B $\quad R_M G_0 \leq 1$

in practice $R_0 < 50\Omega$

in practice $G_0 < 1/50S$

Solutions A and B on the Smith chart

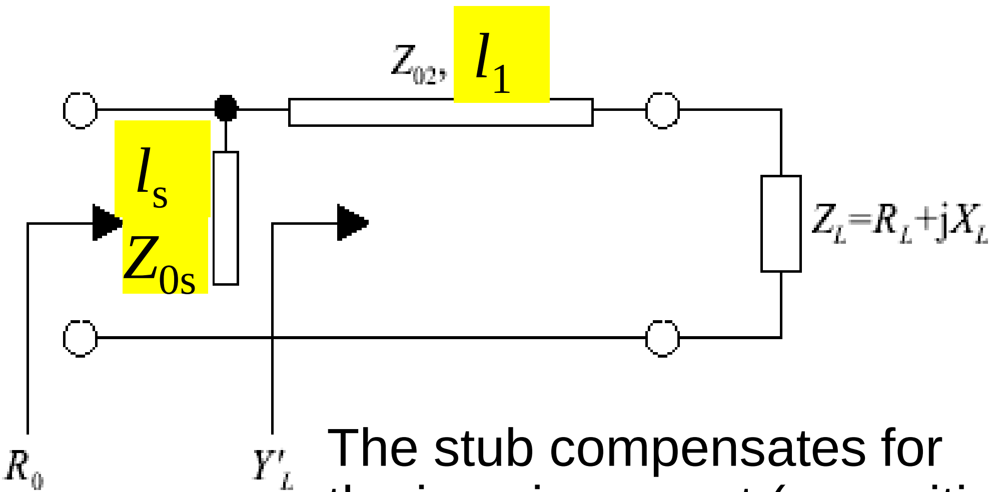
Smith chart Z_0, norm. res. $1/G_M$



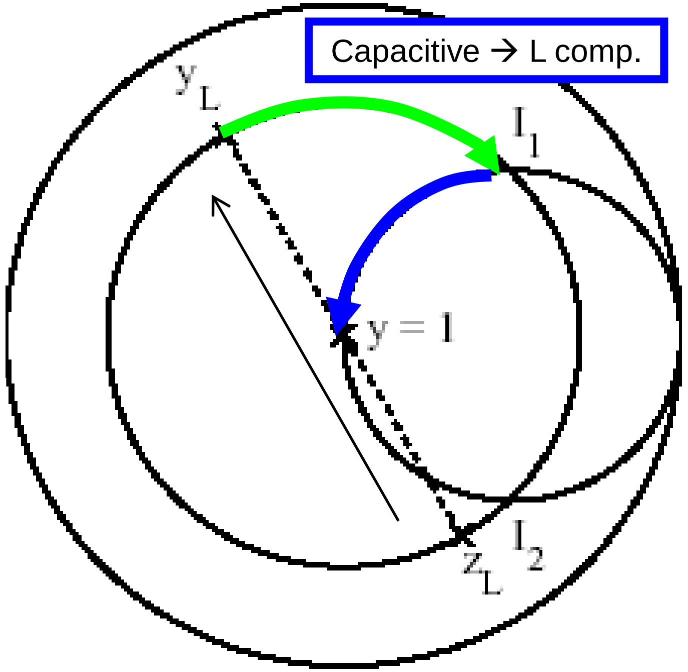
Smith chart Z_0, norm. res. R_M

Line + stub matching network

Typically the characteristic line impedance is given and only the line and stub length should be designed



The stub compensates for the imaginary part (capacitive)

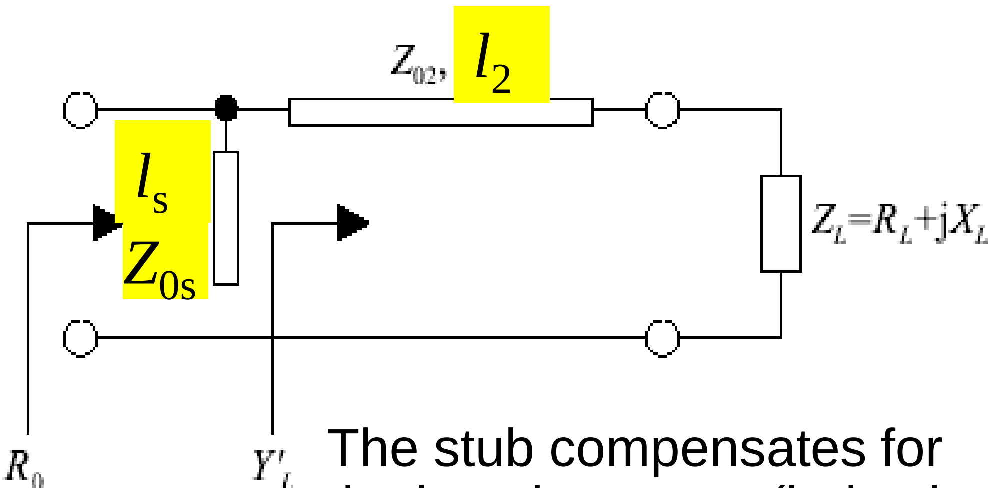


Shorted stub

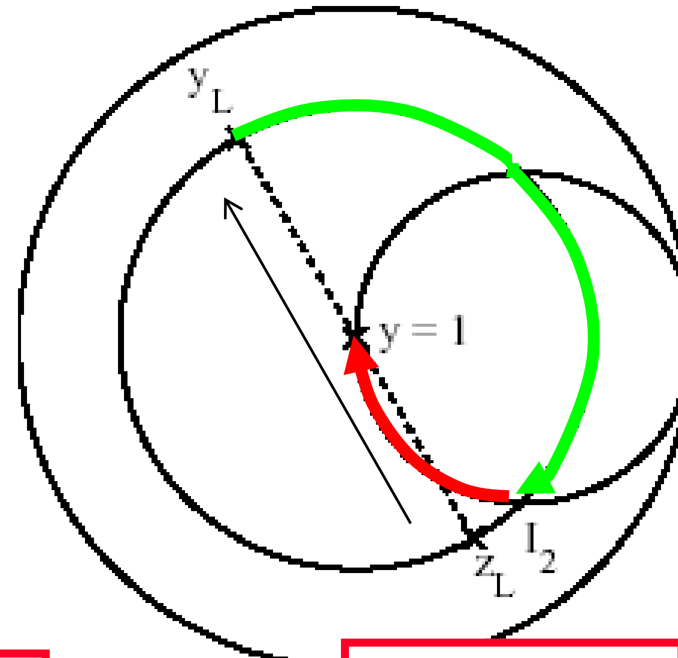
$$Z_{in}(z) = jZ_{0s}\tan(\beta z)$$

Line + stub matching network

Typically the characteristic line impedance is given and only the line and stub length should be designed



The stub compensates for the imaginary part (inductive)



Open stub

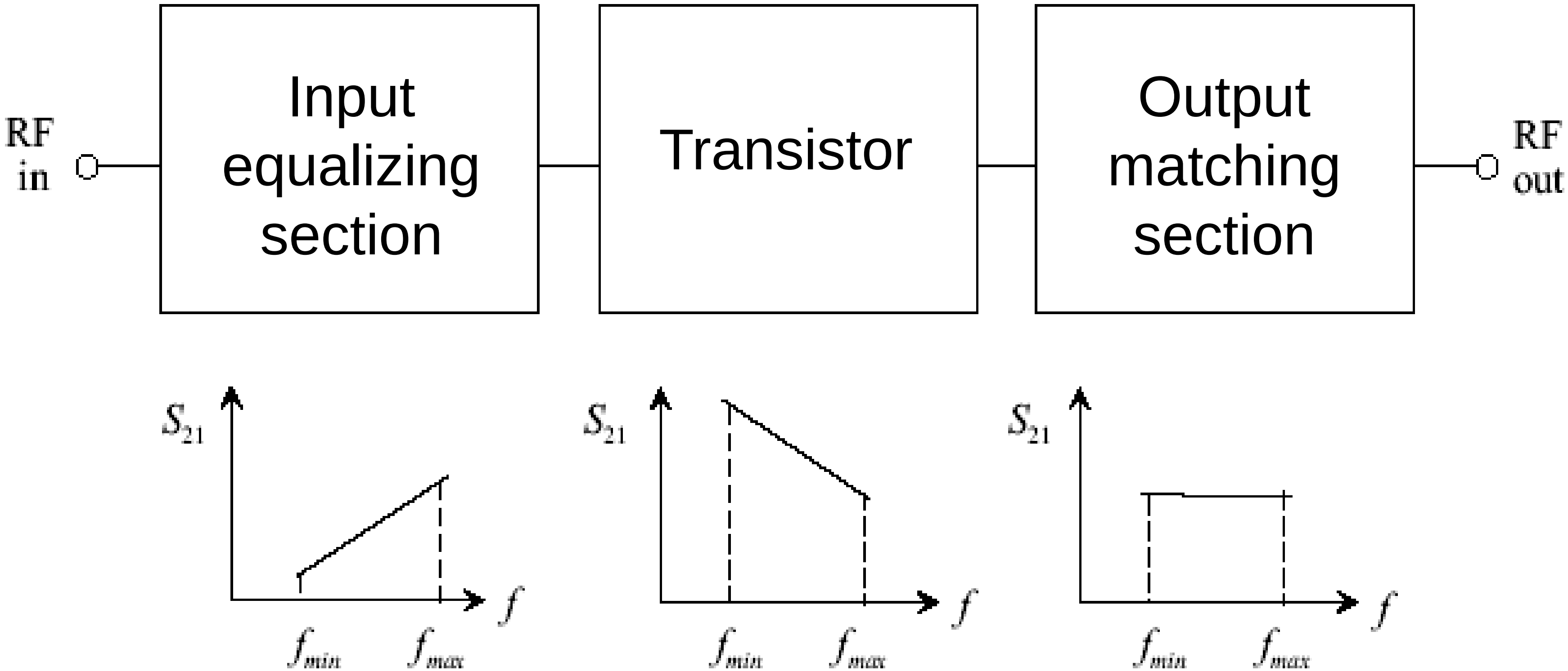
$$Z_{\text{in}}(z) = -jZ_{\infty}\cot(\beta z)$$

Inductive→ C comp.

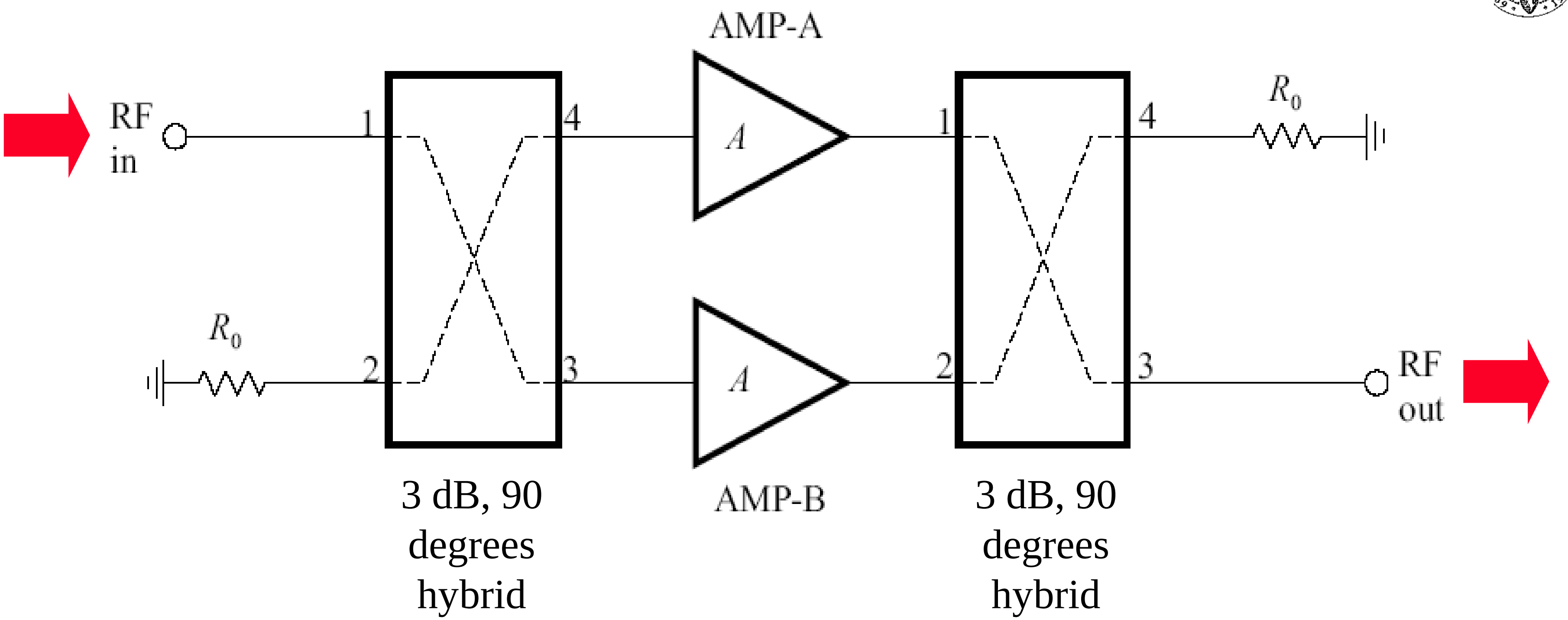
Wideband reactive matching sections for wideband amplifiers

- Transistor inputs (outputs) are RC networks (series or parallel)
- It can be theoretically shown that a lumped or distributed lossless matching section allows for ideal matching in a discrete set of frequencies; however (*Fano's limit*) the average matching *decreases exponentially with bandwidth*, independent of the complexity of the matching stages
- **Moreover, the device MAG decreases with *f* → to have wideband flat gain we need to *mismatch* the transistor (usually at input)**
- If the input network is reactive **mismatch at low frequency** is not compatible with **low input reflection coefficient**, unless some <u>dissipative elements</u> are introduced directly or indirectly in the input matching network

Wideband open-loop amplifiers

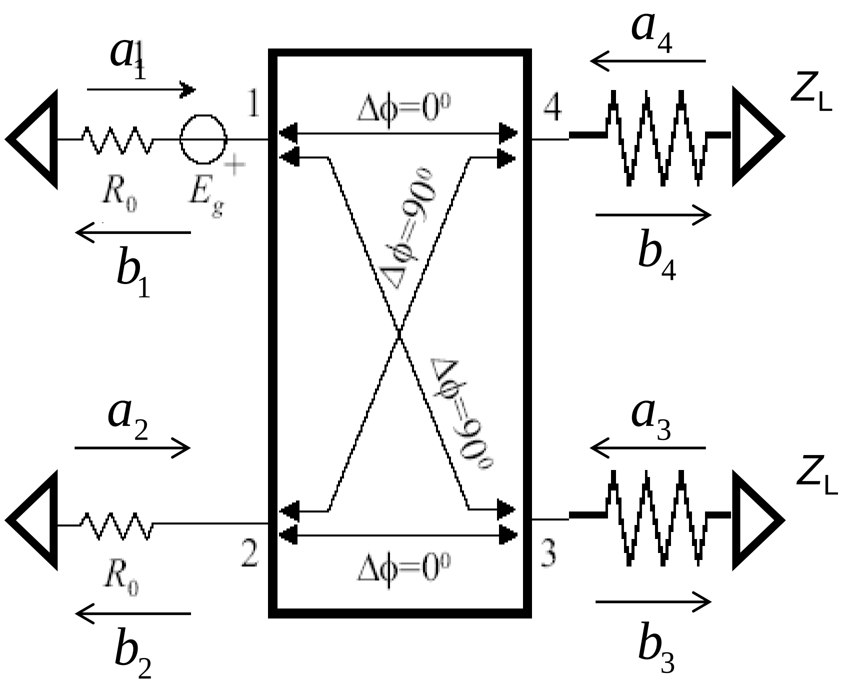


The balanced amplifier



- The BA allows for
  - IN/OUT matching in the coupler bandwidth (1 octave approx.)
  - Gain equal to the single amplifier $|S_{21}|$
  - Saturation power twice as the single amplifier P_{sat}

Balanced amplifier: input power match (center band)



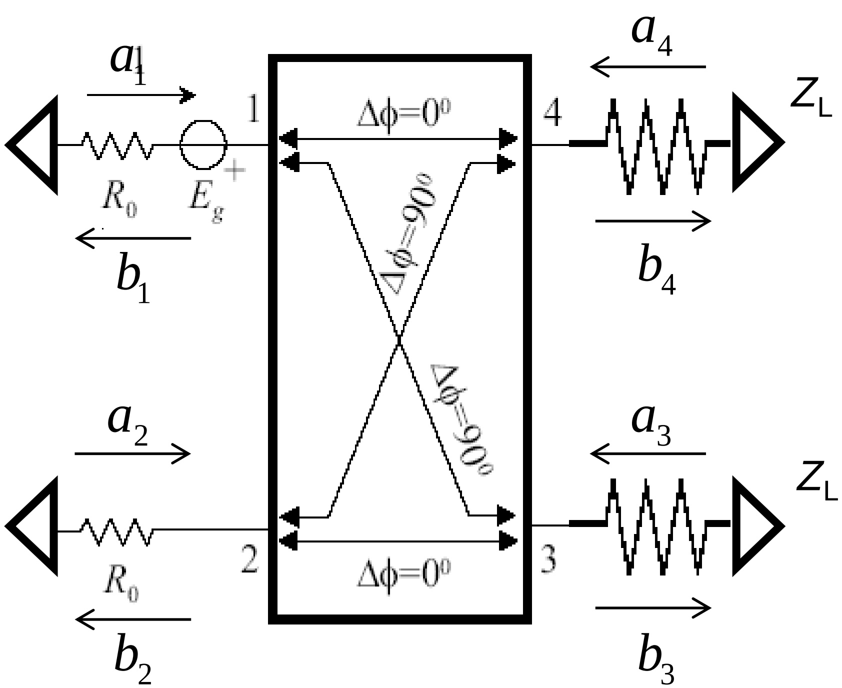
Port 3 and 4 are loaded by the same impedance different from norm. Z

S matrix at centerband

$$\mathbf{b} = \mathbf{S}\mathbf{a}$$

$$\begin{bmatrix} b_1 \\ b_2 \\ b_3 \\ b_4 \end{bmatrix} = \begin{bmatrix} 0 & 0 & -j/\sqrt{2} & 1/\sqrt{2} \\ 0 & 0 & 1/\sqrt{2} & -j/\sqrt{2} \\ -j/\sqrt{2} & 1/\sqrt{2} & 0 & 0 \\ 1/\sqrt{2} & -j/\sqrt{2} & 0 & 0 \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \\ a_3 \\ a_4 \end{bmatrix}$$

Balanced amplifier: input power match (center band)



$$b_1 = \frac{-j}{\sqrt{2}} a_3 + \frac{1}{\sqrt{2}} a_4$$

$$b_2 = \frac{1}{\sqrt{2}} a_3 + \frac{-j}{\sqrt{2}} a_4$$

$$a_2 = 0 \text{ (port loaded on } R_0) \rightarrow$$

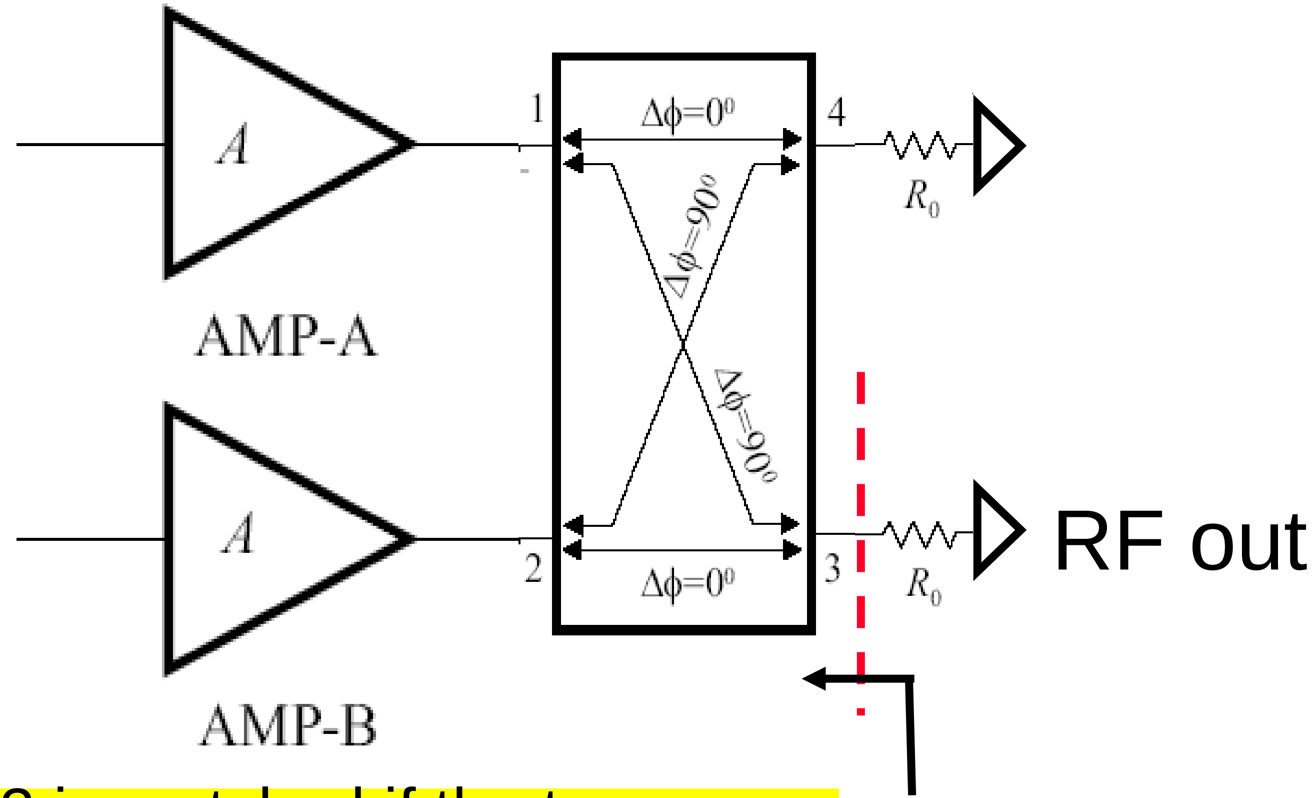
$$b_3 = \frac{-j}{\sqrt{2}} a_1, \quad b_4 = \frac{1}{\sqrt{2}} a_1$$

$$a_3 = \Gamma_L b_3 = \frac{-j\Gamma_L}{\sqrt{2}} a_1, \quad a_4 = \Gamma_L b_4 = \frac{\Gamma_L}{\sqrt{2}} a_1$$

$$\boxed{\begin{aligned} b_1 &= \frac{-j}{\sqrt{2}} \frac{-j\Gamma_L}{\sqrt{2}} a_1 + \frac{1}{\sqrt{2}} \frac{\Gamma_L}{\sqrt{2}} a_1 = 0 \\ b_2 &= \frac{1}{\sqrt{2}} \frac{-j\Gamma_L}{\sqrt{2}} a_1 + \frac{-j}{\sqrt{2}} \frac{\Gamma_L}{\sqrt{2}} a_1 = -j\Gamma_L a_1 \end{aligned}}$$

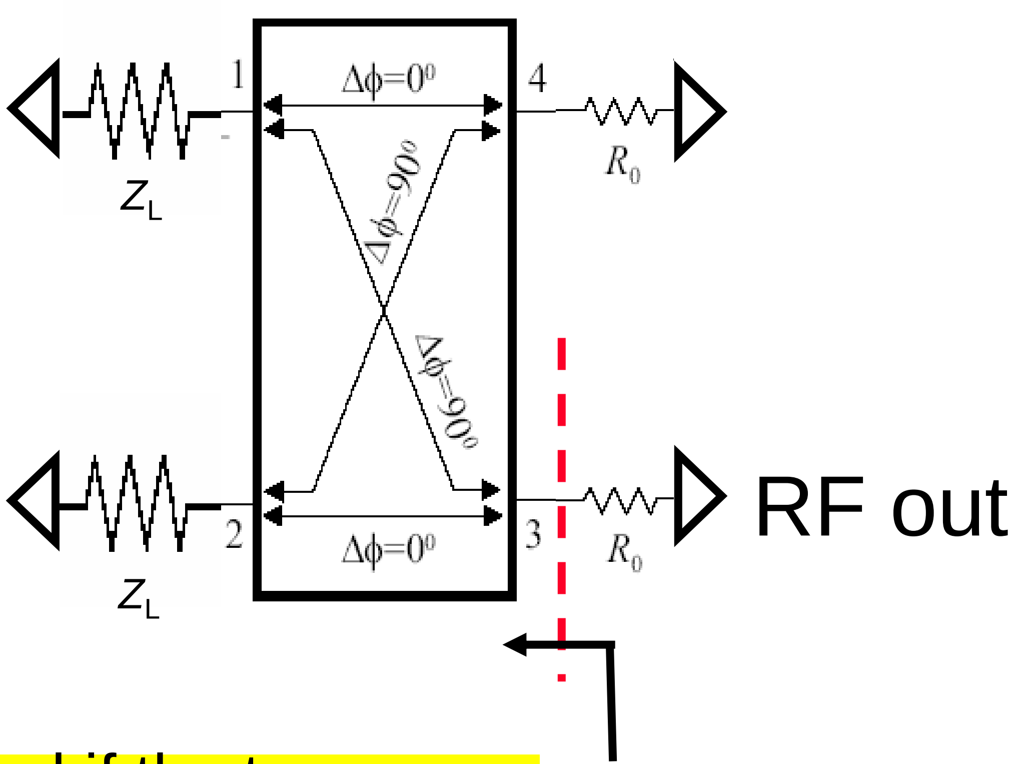
Port 1 is matched if the two amplifiers have the same input impedance
The reflected power is dissipated in the resistor at port 2

Balanced amplifier: output power match (center band)



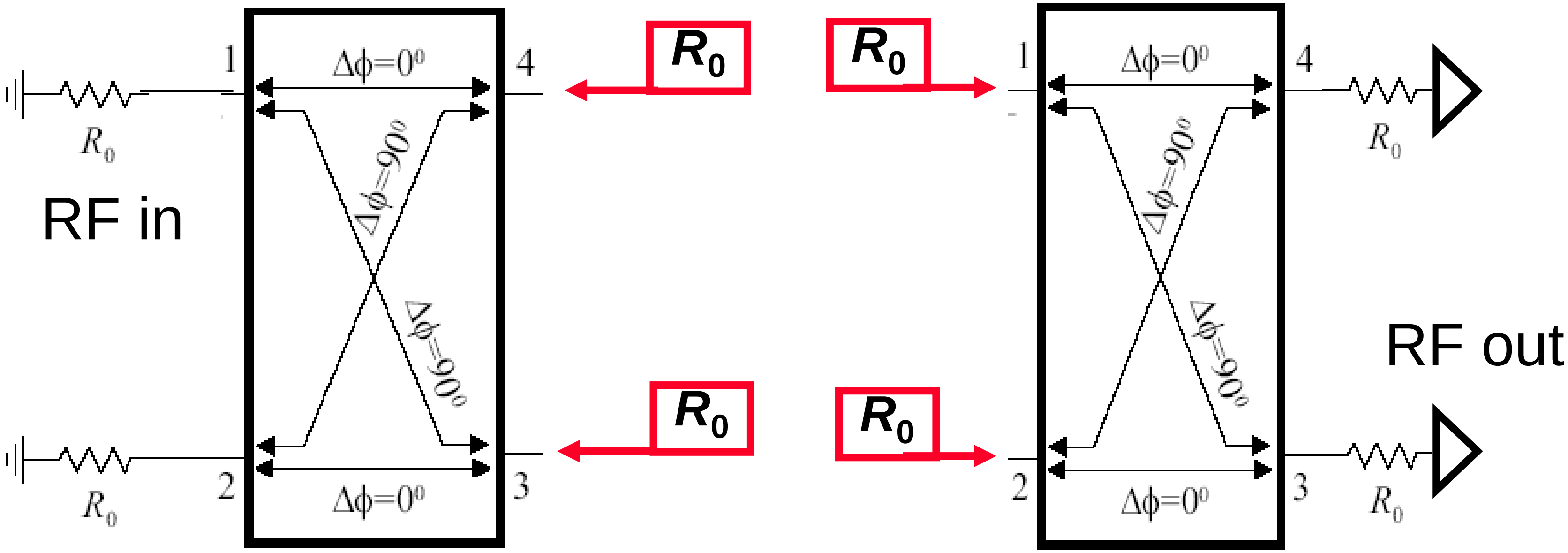
Port 3 is matched if the two amplifiers have the same output impedance
The reflected power is dissipated in the resistor at port 4

Balanced amplifier: output power match (center band)



Port 3 is matched if the two amplifiers have the same output impedance
The reflected power is dissipated in the resistor at port 4

Balanced amplifier: device input and output matching



Ports 1 and 2 are matched, $a_1 = a_2 = 0$

$$b_3 = \frac{-j}{\sqrt{2}} a_1 + \frac{1}{\sqrt{2}} a_2 = 0$$

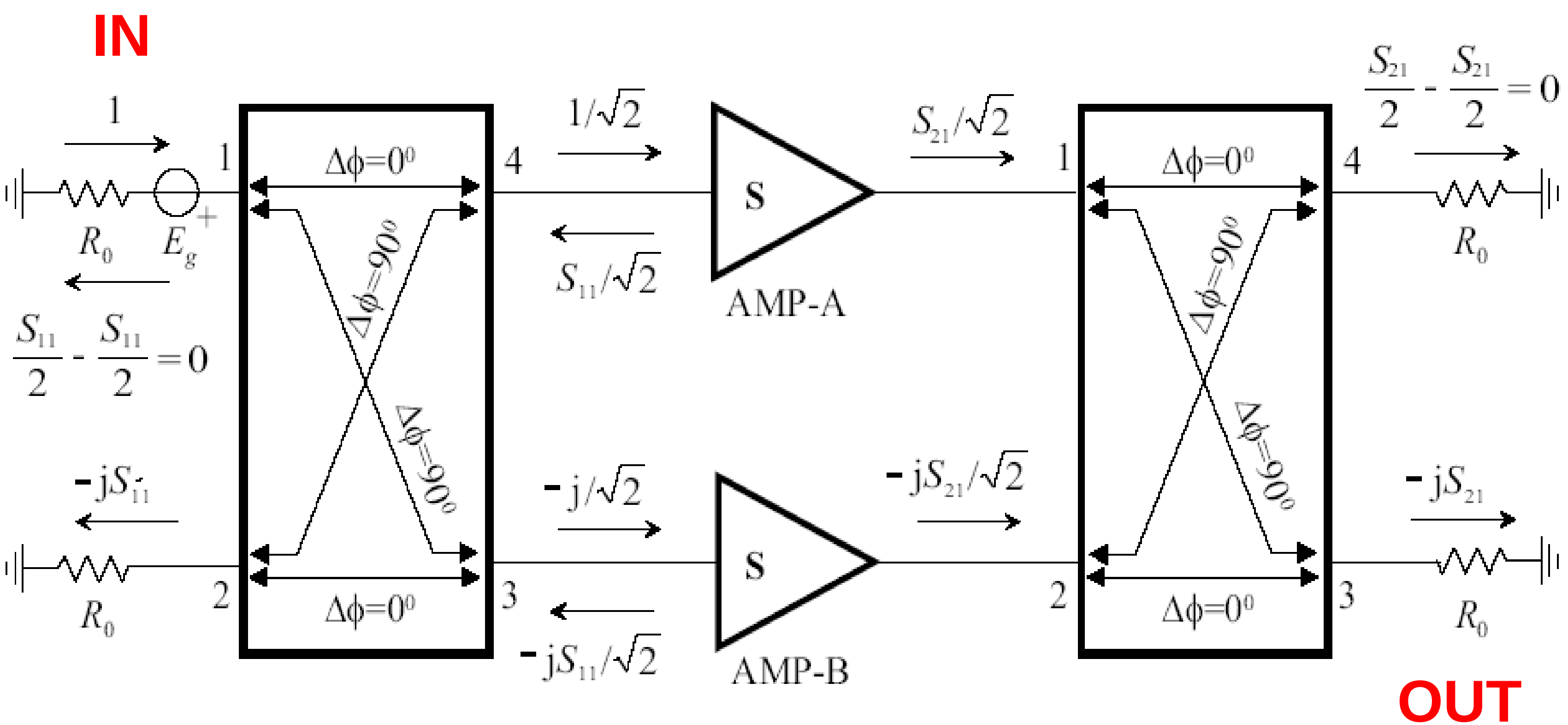
$$b_4 = \frac{1}{\sqrt{2}} a_1 - \frac{j}{\sqrt{2}} a_2 = 0$$

also ports 3 and 4 have $Z_{in} = R_0$

At the input and output ports the two AMPS A B are closed on the normalization resistance

The AMPS A B is described by the S parameters on R_0

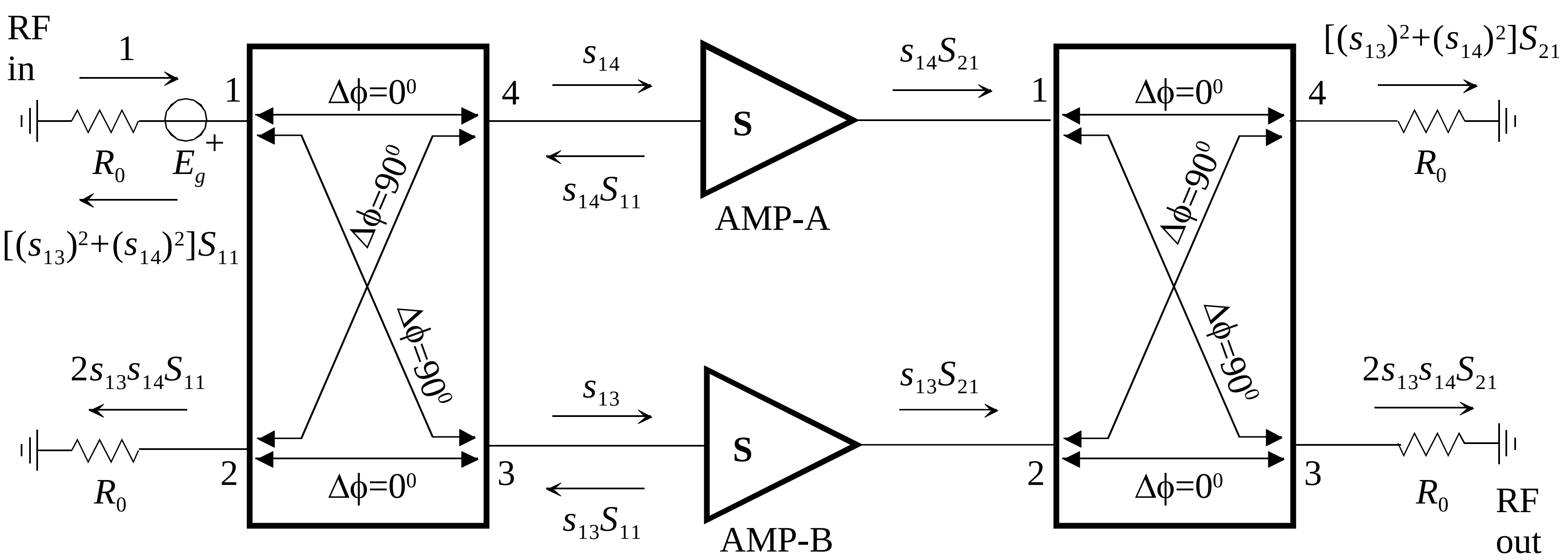
Balanced amplifier operation (centerband)



Amplification →$|S_{21}|$, gain $|S_{21}|^2$

Balanced amplifier operation (out of centerband)

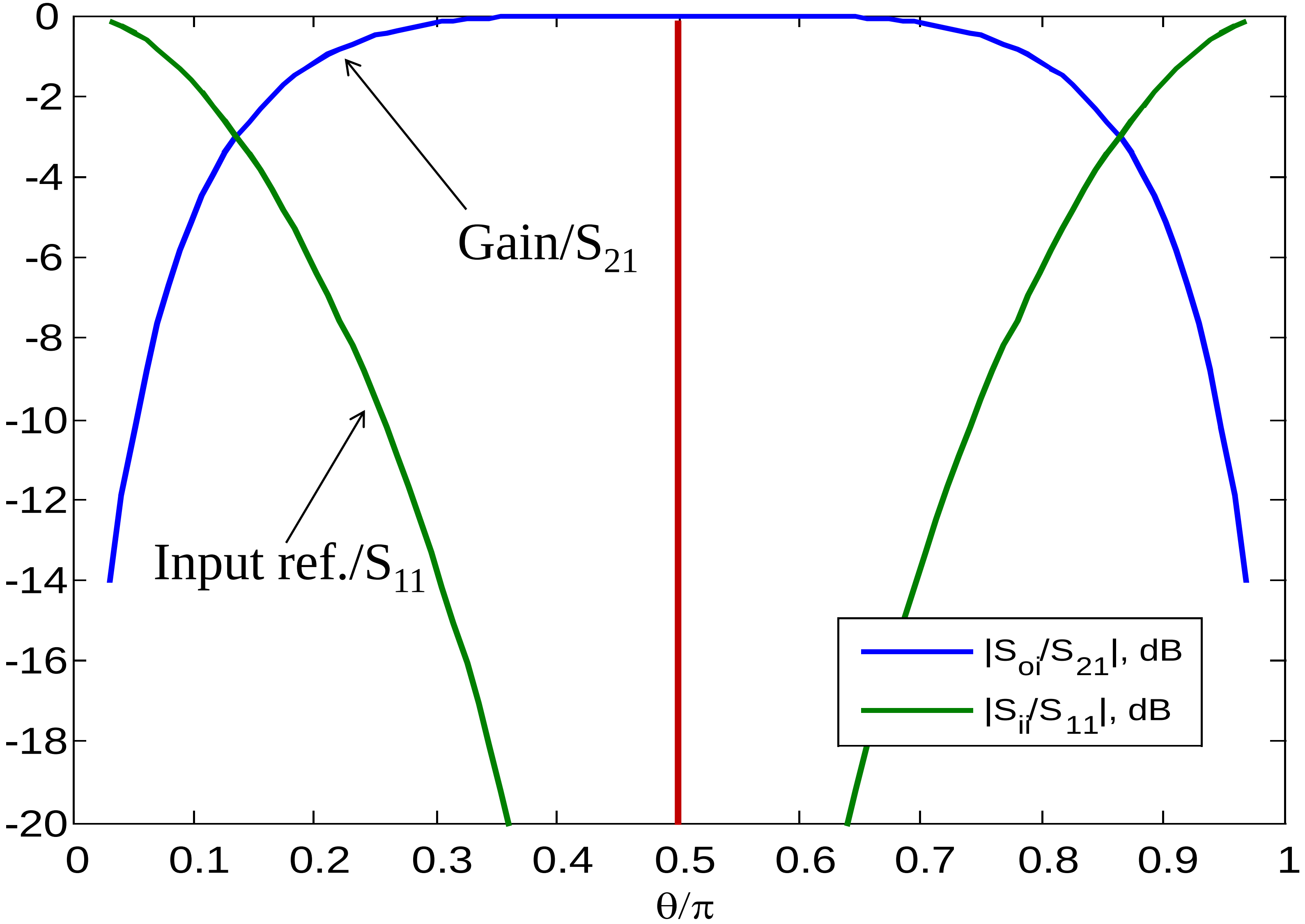
IN



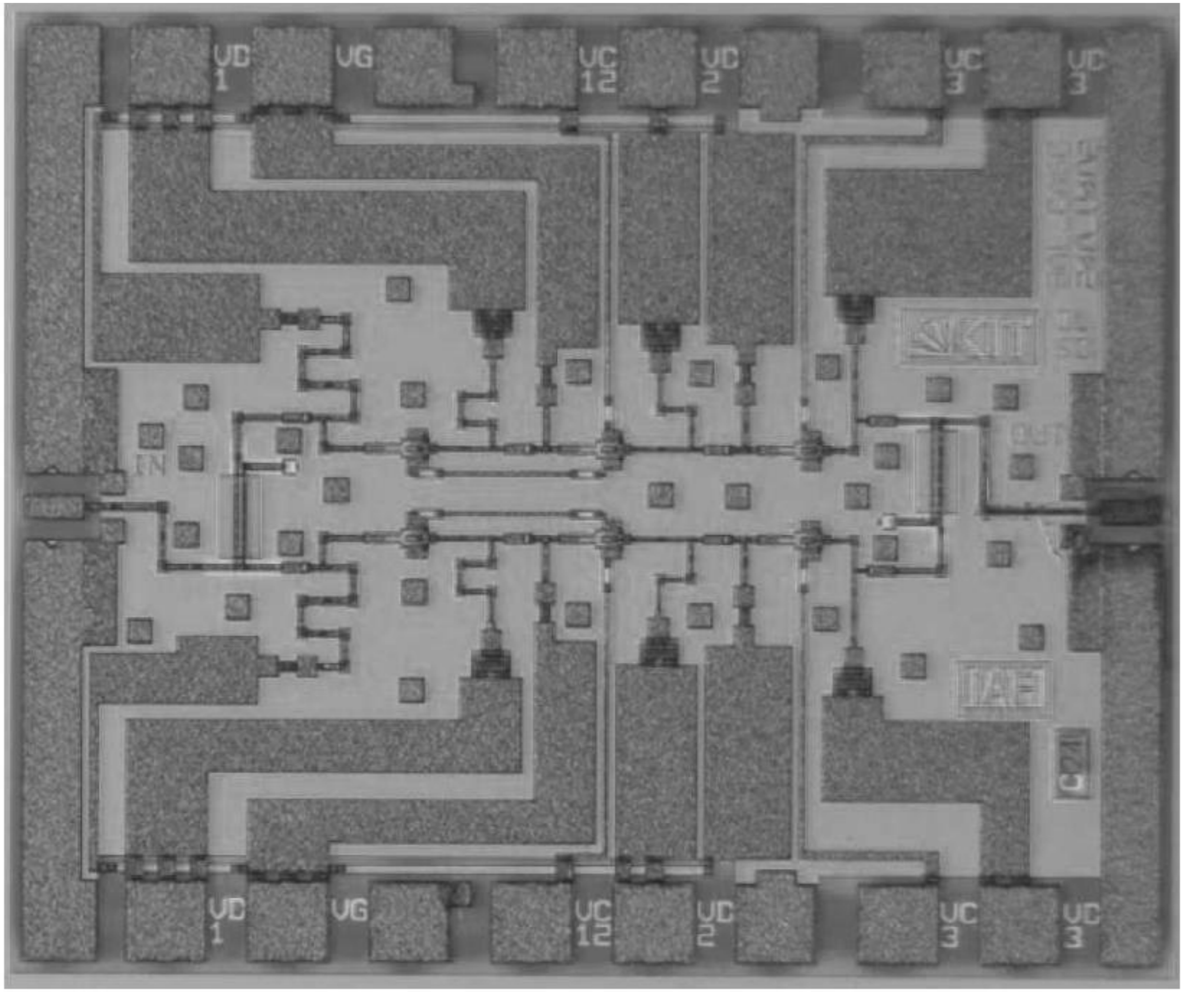
OUT

Amplification → $|2s_{13}s_{14}S_{21}|$**, gain** $|2s_{13}s_{14}S_{21}|^2$

Balanced amplifier operation (out of centerband)



Mm-wave MMIC balanced amplifier



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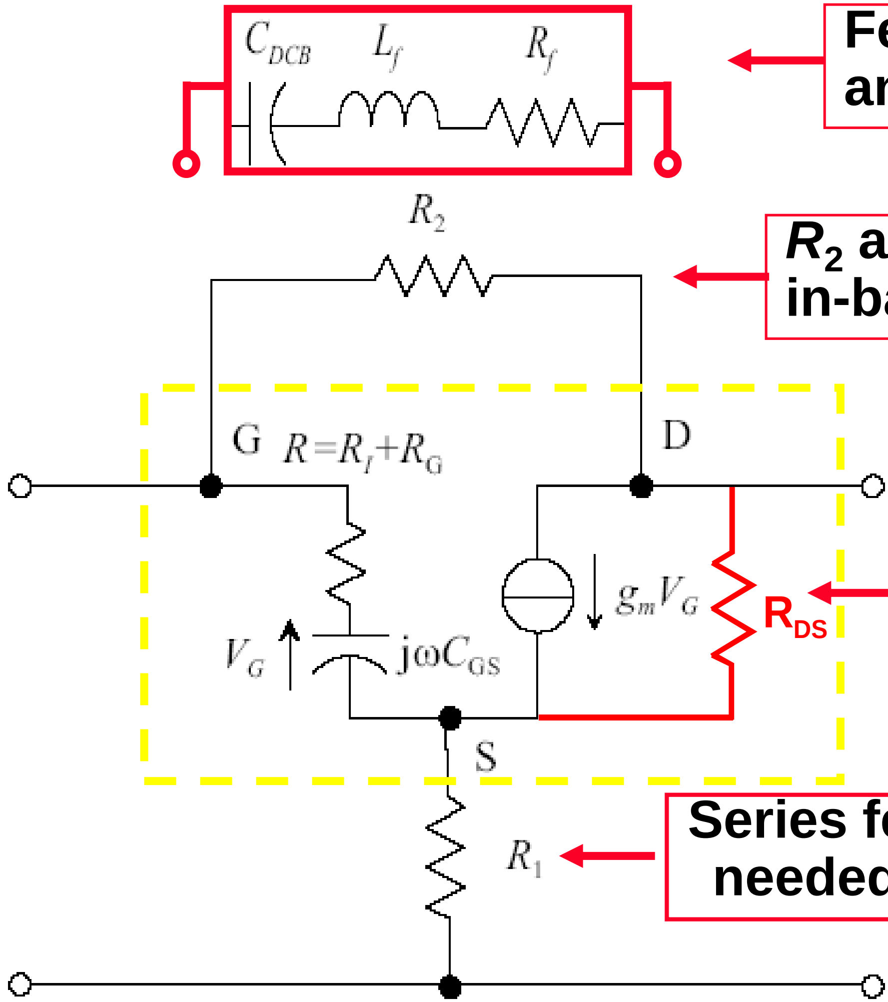
RF parallel feedback amplifiers

- As at low frequency, resistive parallel feedback is used to obtain flat gain from low frequency up to a high-frequency cutoff, however feedback also improves input and output matching
- Parallel feedback requires high (enough) open-loop gain vs. gain with feedback
- In the example we consider in the low-frequency limit a device with *mixed resistive series-parallel feedback*; reactive elements are added to the feedback path to increase the gain flatness ("inductive peaking")
- Feedback common also in narrowband MMIC design → gain stabilization vs. active element fluctuations.

RF feedback amplifier design



Feedback with DC block and inductance

R_2 also provides in-band stabilization

R_{DS} will be neglected, but becomes low is the device periphery in increased to increase g_m !

Series feedback R_1 only needed to input match

Low-frequency analysis

- The scattering parameters of the feedback amplifier can be shown to be:

$$S_{11}(0) = \frac{\boxed{\left(R_2R_1 - R_0^2\right) g_m + R_2}}{\left(2R_0R_1 + R_2R_1 + R_0^2\right) g_m + 2R_0 + R_2}$$

$$S_{12}(0) = 2R_0 \frac{1 + R_1 g_m}{\left(2R_0R_1 + R_2R_1 + R_0^2\right) g_m + 2R_0 + R_2}$$

$$S_{21}(0) = -2R_0 \frac{\left(R_2 - R_1\right) g_m - 1}{\left(2R_0R_1 + R_2R_1 + R_0^2\right) g_m + 2R_0 + R_2}$$

$$S_{22}(0) = \frac{\boxed{\left(R_2R_1 - R_0^2\right) g_m + R_2}}{\left(2R_0R_1 + R_2R_1 + R_0^2\right) g_m + 2R_0 + R_2}.$$

- Input and output matching can be simultaneously obtained by proper dimensioning of the feedback resistances!

Matching condition and minimum open-loop gain

- Matching implies:

$$R_1 = \frac{R_0^2}{R_2} - \frac{1}{g_m} \Rightarrow \boxed{\begin{aligned} S_{11f} &= S_{22f} = 0 \\ S_{12f} &= \frac{R_0}{R_0 + R_2} \\ S_{21f} &= \frac{R_0 - R_2}{R_0} \end{aligned}}$$

- The parallel feedback resistance is obtained from feedback gain:

$$R_2 = R_0\left(1 + \left|S_{21f}(0)\right|\right)$$

- A minimum g_m is needed:

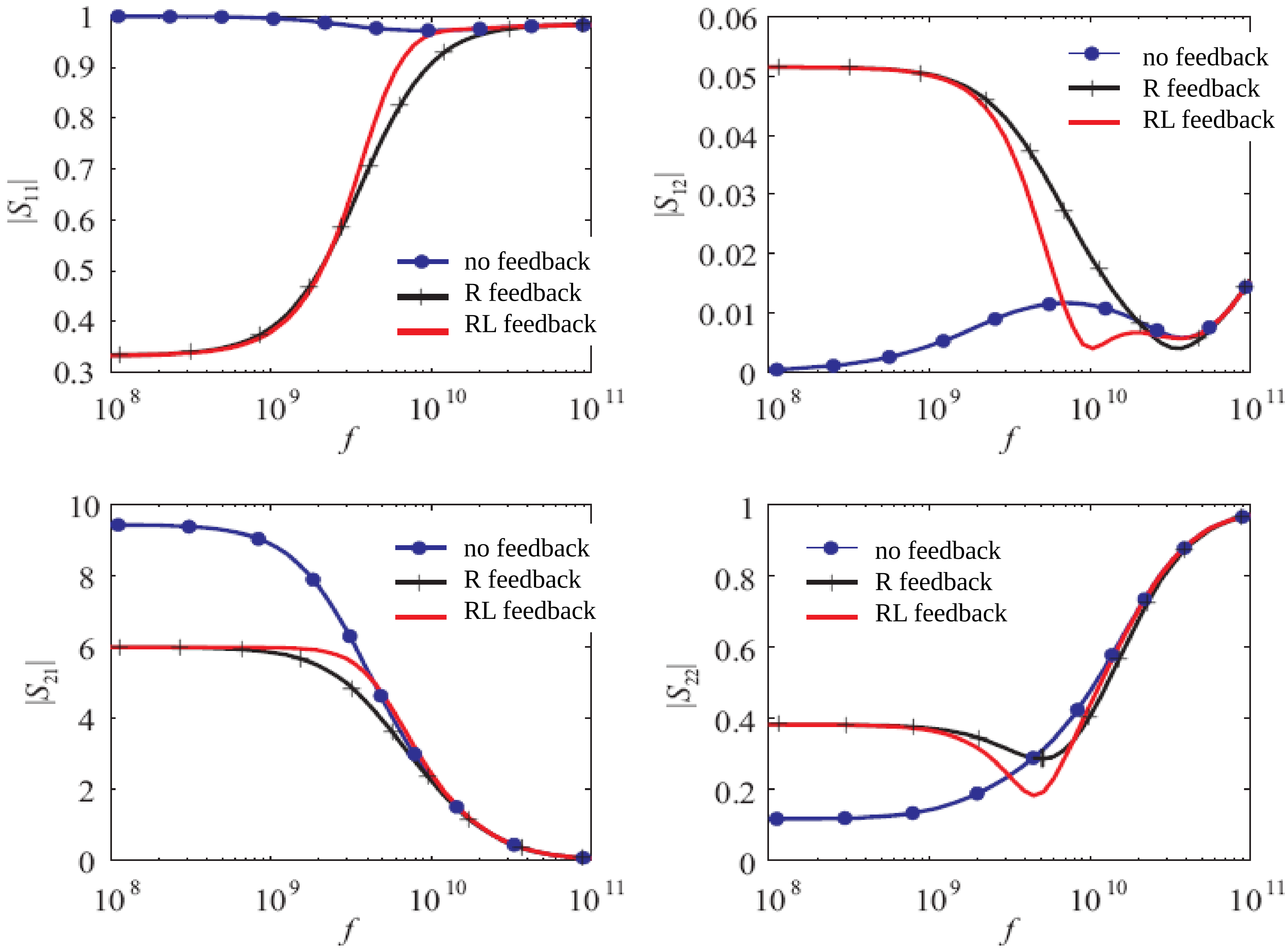
$$g_m \geq R_2 / R_0^2$$

- corresponding to a condition on open-loop gain:

$$\left|S_{21}\right| \geq 2\left(1 + \left|S_{21f}(0)\right|\right)$$

- The transconductance can be increased by scaling but with a decrease of R_{DS}….
- In MMIC design the minimum g_m is used to avoid series feedback

S parameters with & without feedback



Ultrawideband distributed amplifiers (DAMPs)

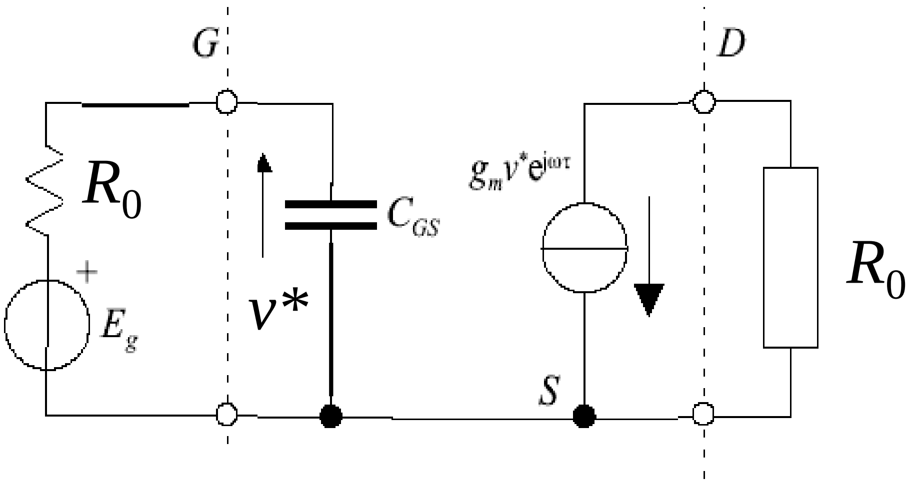
- DAMPs allow to (theoretically) increase the gain-bandwidth product (GBP) of conventional amplifiers through a distributed structure made of two transmission lines having the same phase velocitiy:

$$f_T = A_V(0) f_\beta = \frac{g_m}{2\pi C_{GS}} \qquad \text{(why?)}$$

Low-frequency gain ↑ ↑ 3dB bandwidth

- In practice the bandwidth is limited by velocity mismatch and by losses; transmission lines are implemented through a discrete-cell approach
- Practical DAMPS have 5-10 cells with low gain (10-15 dB) but spectacular bandwidths (e.g. 100 KHz - 60 GHz)

Gain Bandwidth Product



$$v^* = \frac{E_g}{1 + j\omega C_{GS} R_0}$$

$$V_L = -g_m v^* R_0 = -\frac{g_m R_0 E_g}{1 + j\omega C_{GS} R_0}$$

$$\frac{V_L}{E_g} = A_V(f) = -\frac{R_0 g_m}{1 + j\omega C_{GS} R_0}$$

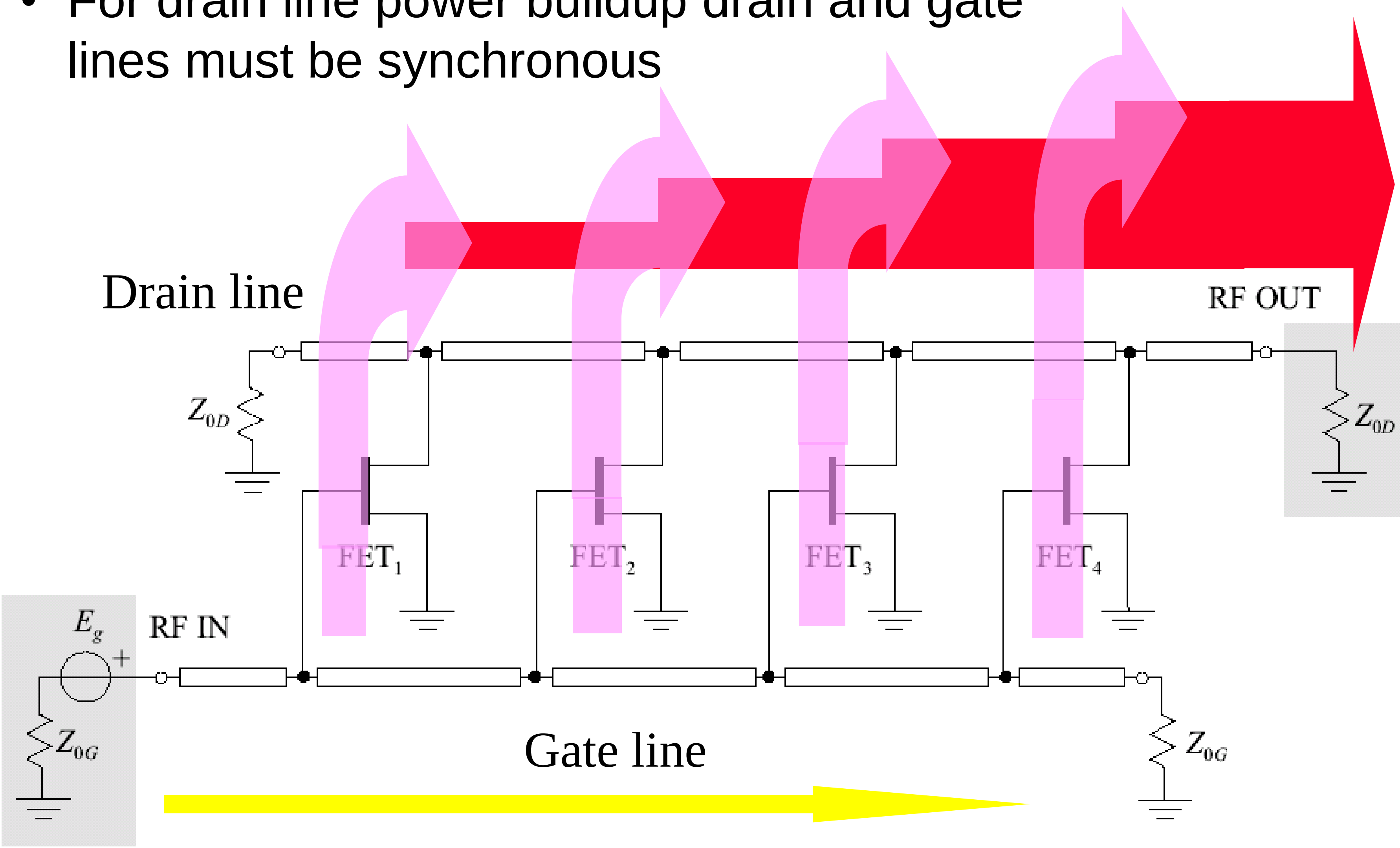
$$|A_V(0)| = R_0 g_m, \quad f_\beta = \frac{1}{2\pi C_{GS} R_0}$$

$$\boxed{|A_V(0)|\, f_\beta = \frac{g_m}{2\pi C_{GS}} = f_T}$$

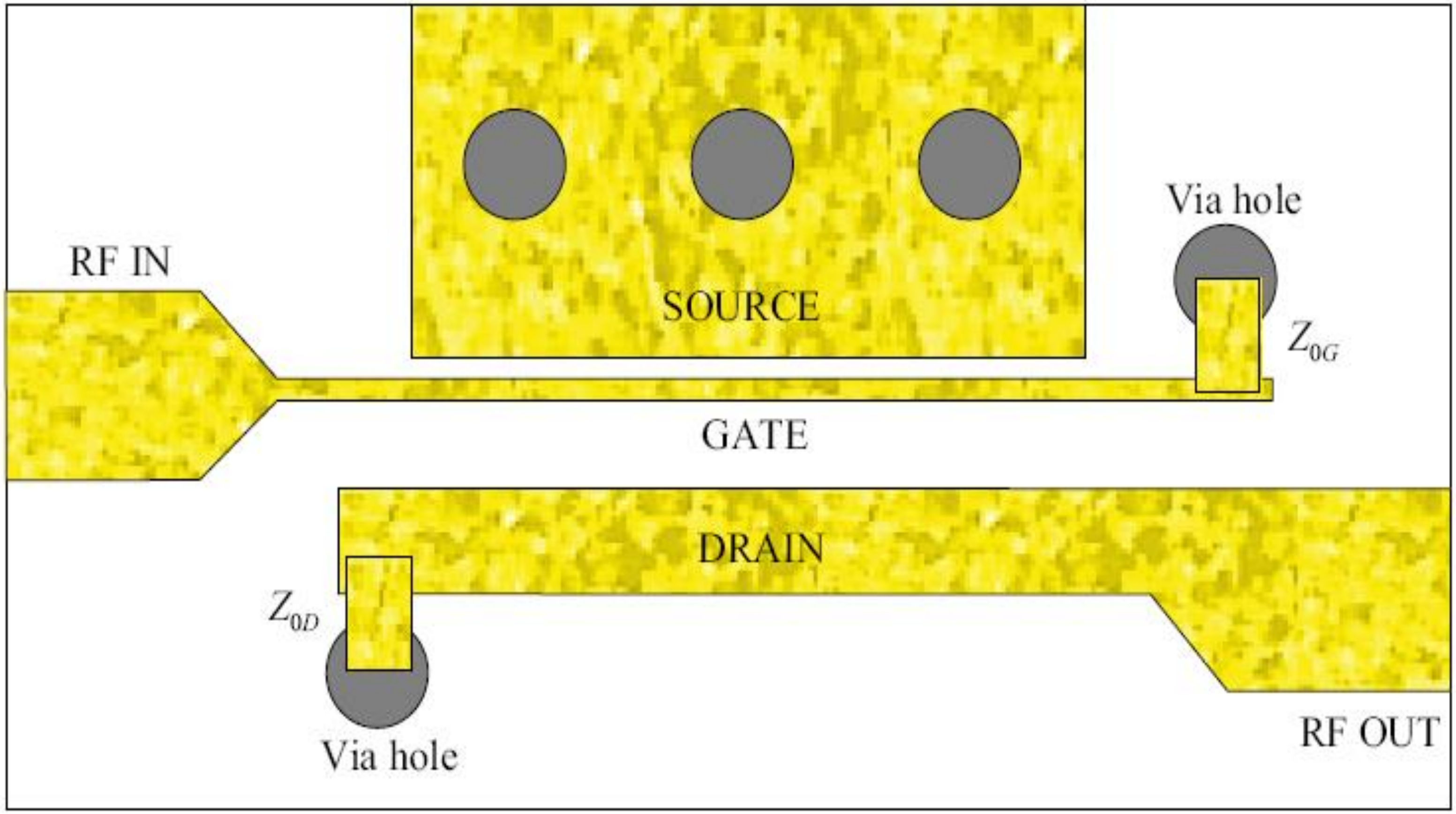
- Same generator and load resistance

The DAMP concept

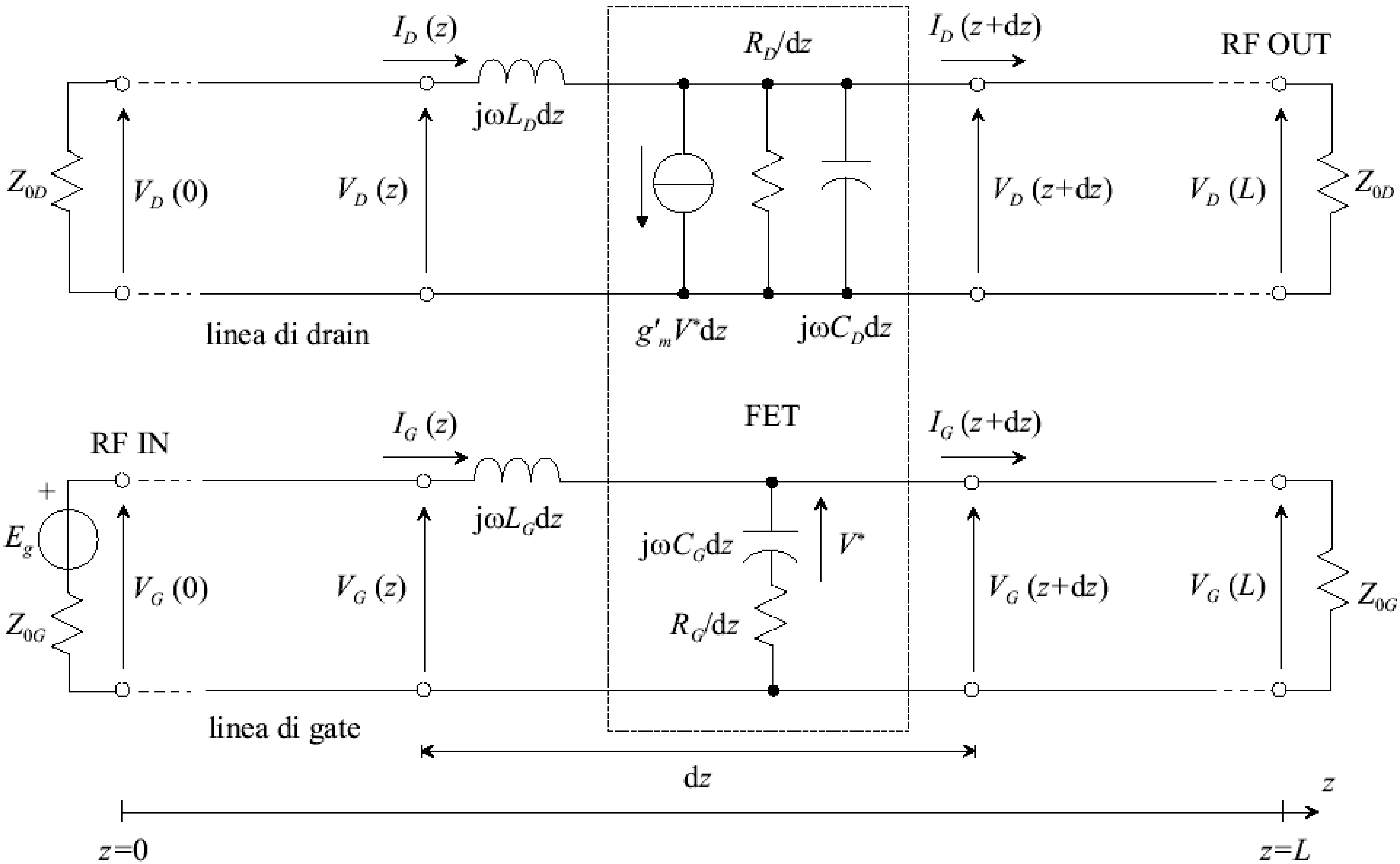
- For drain line power buildup drain and gate lines must be synchronous



The continuous DAMP concept



The continuous DAMP model



Gate and drain lines

- In general those are dispersive, however a frequency range exists were they behave as lossy nondispersive lines:

$$Z_{0G} \approx \sqrt{\frac{L_G}{C_G}}$$

$$Z_{0D} \approx \sqrt{\frac{L_D}{C_D}}$$

$$\gamma_G \approx \frac{\omega^2 R_G C_G \sqrt{L_G C_G}}{2} + \mathrm{j}\omega\sqrt{L_G C_G} = \alpha_G + \mathrm{j}\beta_G$$

$$\gamma_D \approx \boxed{\frac{1}{2}\frac{\sqrt{L_D C_D}}{R_D C_D} +} \mathrm{j}\omega\sqrt{L_D C_D} = \alpha_D + \mathrm{j}\beta_D$$

DAMP frequency response

- The frequency response is of a sinc type:

$$\boxed{A_V = -\frac{g_m Z_{0D}}{2\left(1+\mathrm{j}\omega\tau_G\right)}\,\mathrm{e}^{-\left(\frac{\gamma_D+\gamma_G}{2}\right)L}\,\frac{\sinh\left[\left(\frac{\gamma_D-\gamma_G}{2}\right)L\right]}{\left(\frac{\gamma_D-\gamma_G}{2}\right)L}}$$

- **Neglecting losses, infinite bandwdith for <u>synchronous gate and drain lines</u>:**

$$|A_V| = \frac{g_m Z_{0D}}{2}\frac{\sin\left[\left(\frac{\beta_D-\beta_G}{2}\right)L\right]}{\left(\frac{\beta_D-\beta_G}{2}\right)L} \;\Longrightarrow\; \boxed{|A_V| = \frac{Z_{0D}\,g_m}{2}}$$

$$\boxed{L_D C_D = L_G C_G}$$

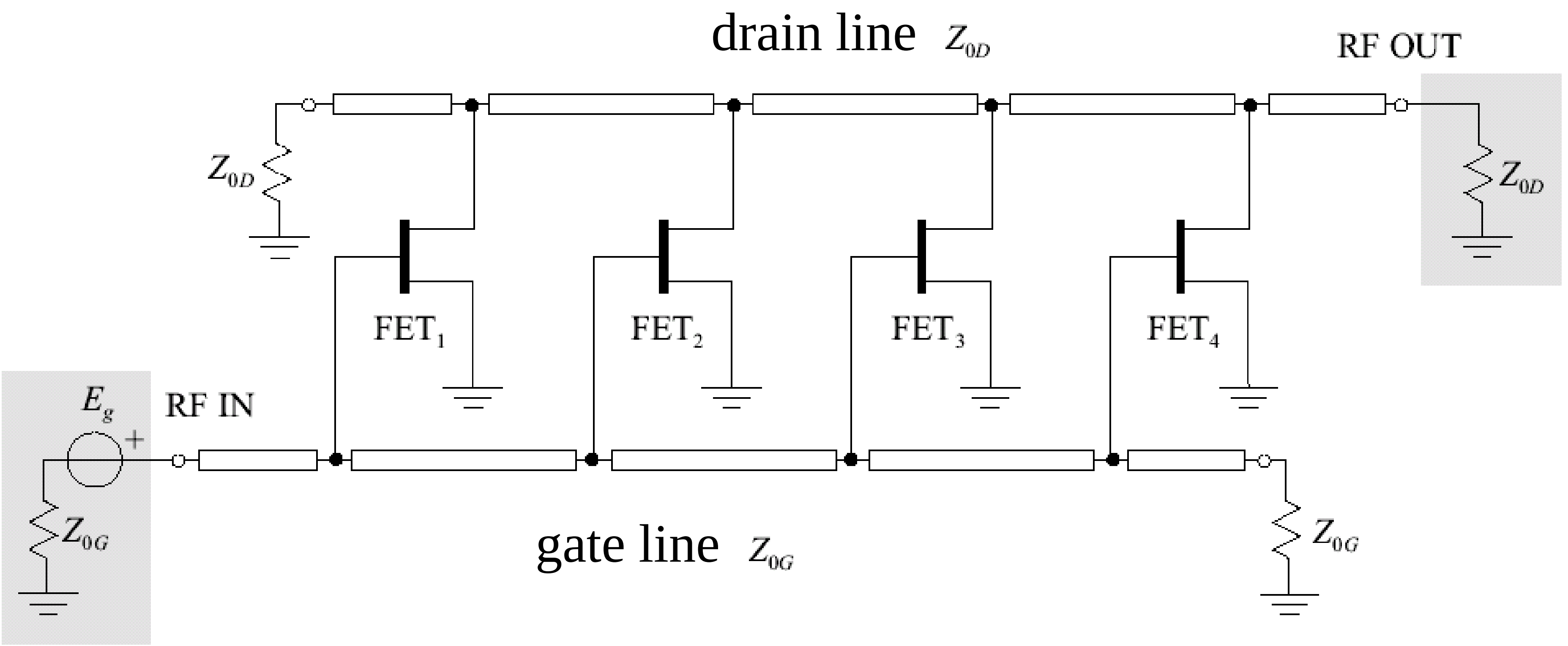
…too good to be true…

- In practice the frequency response is limited by velocity mismatch and losses:

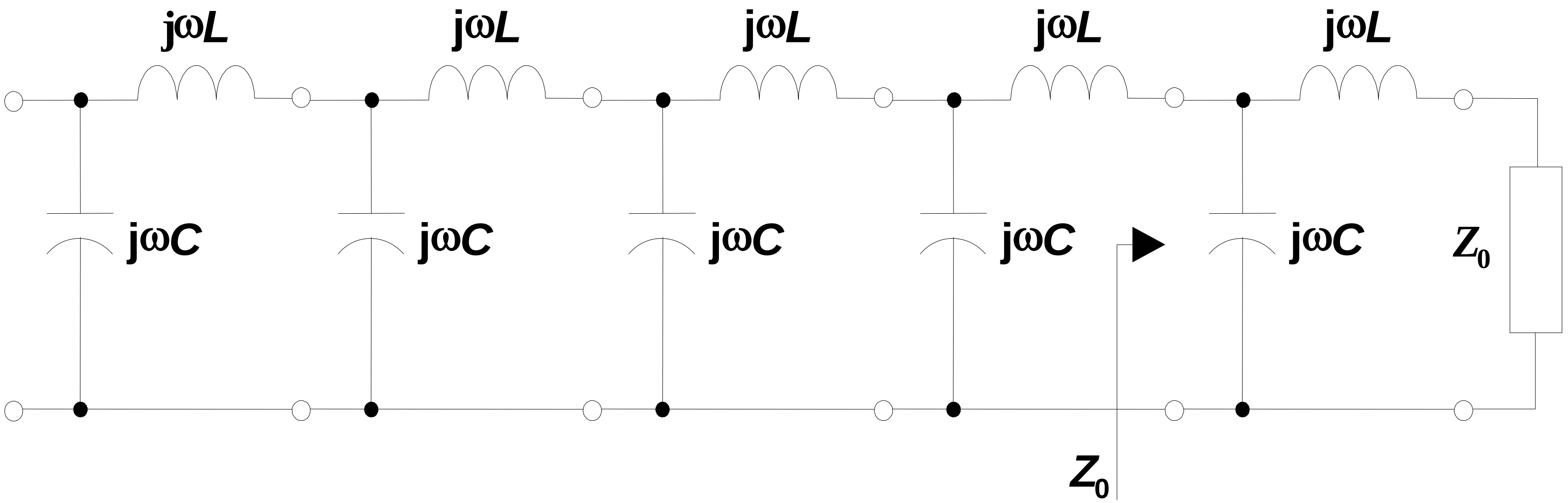
$$L_{opt} = \frac{\log\left(\alpha_D/\alpha_G\right)}{\alpha_D - \alpha_G}$$

- Moreover the continuous amplifier is unfeasible from a technological standpoint → losses are huge, the discrete-cell structure allows an easier velocity matching through lumped or distributed elements
- Bandwidth is limited by losses and by the bandwidth of the delay elements (“artificial lines”).

The discrete DAMP



Artificial lines (Z_0 iterative impedance)



$$\boxed{Z_0 = \frac{1}{j\omega C + \dfrac{1}{j\omega L + Z_0}}}$$

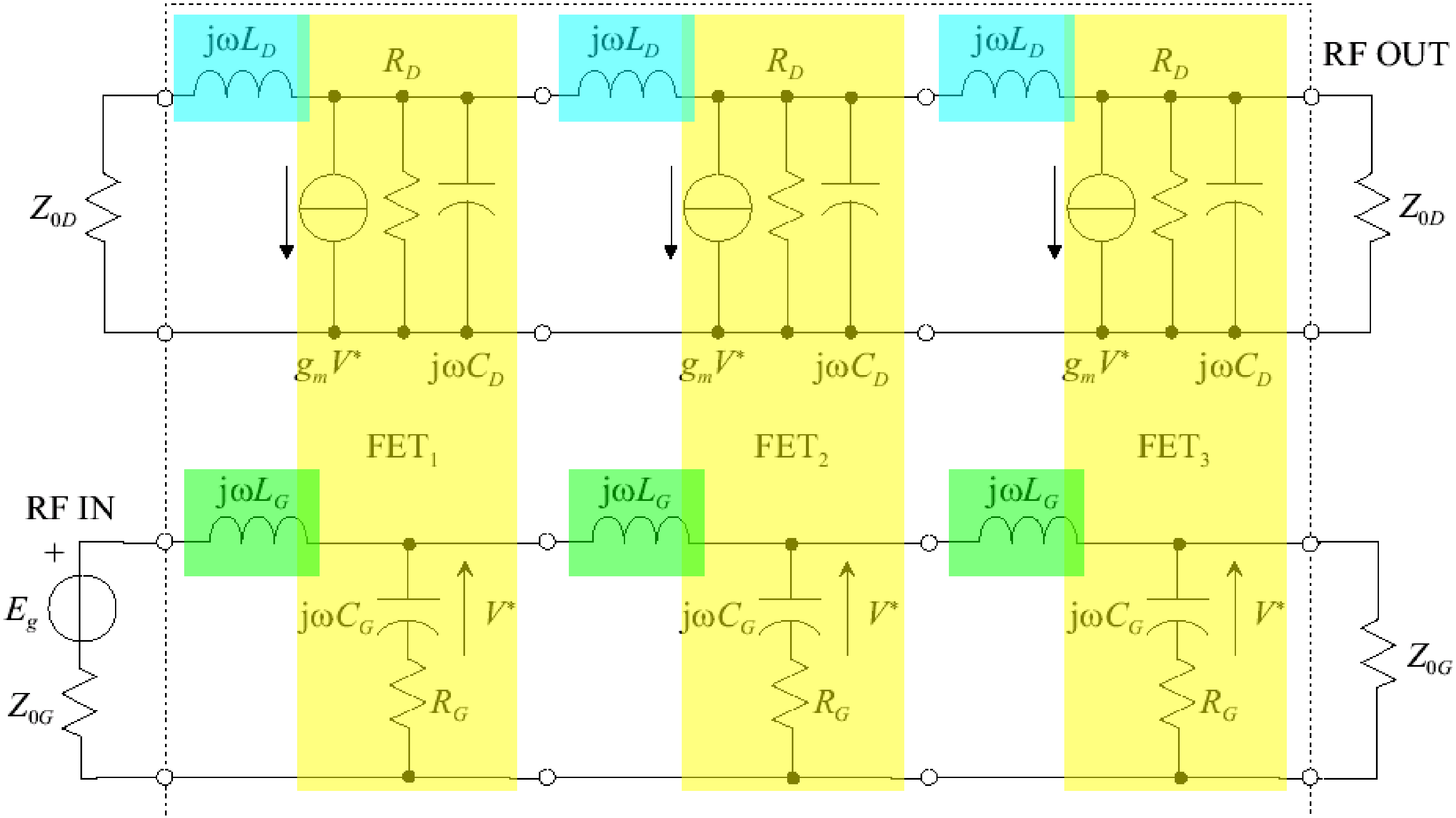
$$\boxed{\begin{gathered} Z_0^{\;2} + j\omega L Z_0 - \frac{L}{C} = 0 \\ Z_0 = -j\omega \frac{L}{2} + \sqrt{\frac{-\omega^2 L^2}{4} + \frac{L}{C}} \end{gathered}}$$

$$\boxed{f_C = \frac{1}{\pi Z_0(0) C}}$$

$$\boxed{A_V(0) = \frac{Z_0(0)}{2} n g_m}$$

$$\boxed{f_T = \frac{n g_m}{2\pi C}}$$

Discrete cell DAMP model



Discrete cell DAMP

- In a discrete cell DAMP the ideal frequency response suggest that the gain bandwidth product is:

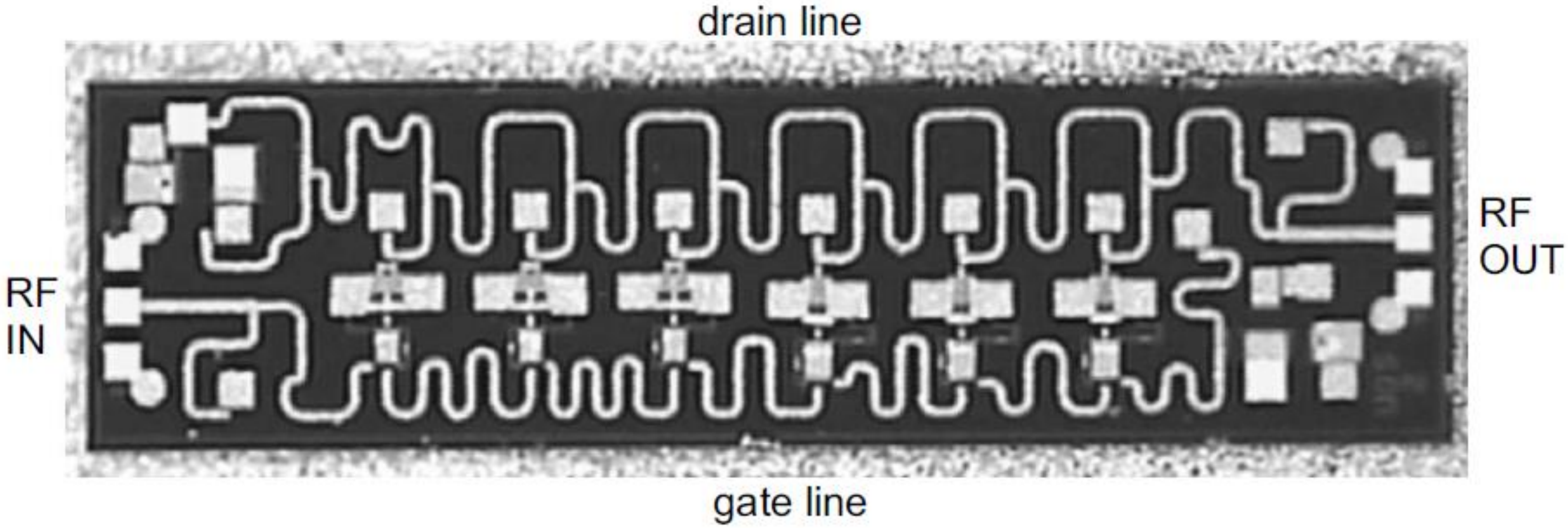
PBG (f_T) of n cells ~ n x PBG (f_T) of one cell

- Optimum cell number → losses → typically 10-15 max in integrated implementations

$$n_{ott} = \frac{\log(A_D/A_G)}{A_D - A_G}$$

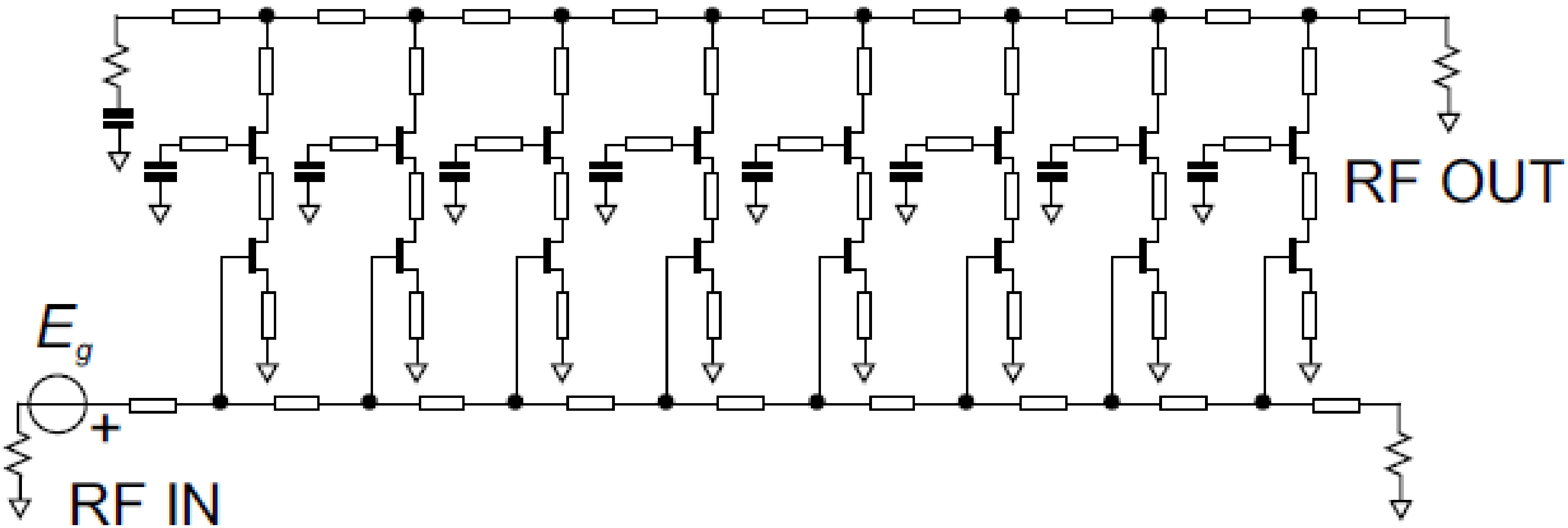
- A_D and A_G are the total attenuations of the gate and drain lines.

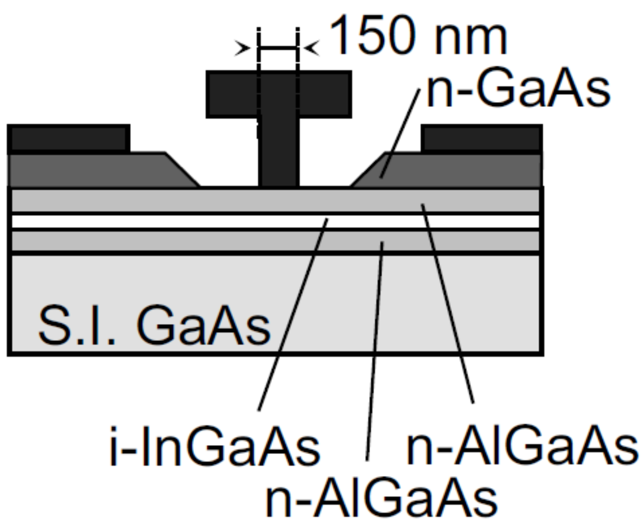
Example of integrated DAMP



The amplifier has 6 stages and includes drain and gate delay lines having different length to achieve synchronous coupling. The chip technology is power PHEMT and the chip size is 3.75 × 1 mm2. The average gain is 7 dB up to around 15 GHz and the DAMP was designed as a 10 Gbps electro-optic modulator driver. Copyright 2005 GAAS Association. All rights reserved. Reprinted, with permission, from J. Shohat, I. D. Robertson, and S. J. Nightingale, “High efficiency 10 Gb/s optical modulator driver amplifier using a power pHEMT technology,” in European Gallium Arsenide and Other Semiconductor Application Symposium, GAAS 2005, Oct. 2005, pp. 129–132.

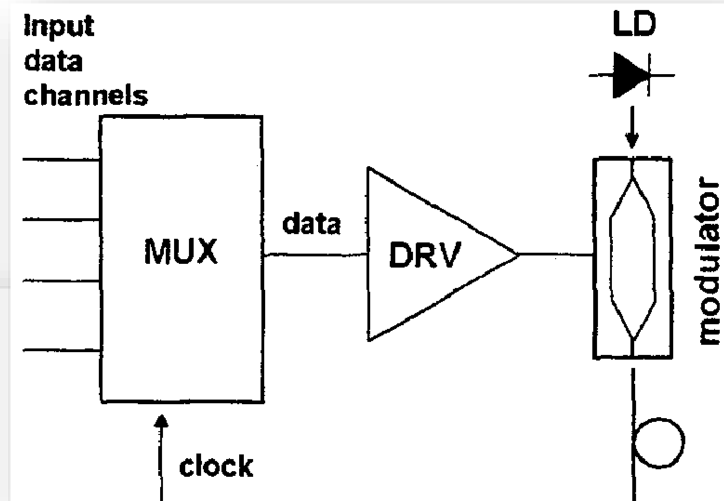
Fujitsu cascode DAMP for optical communication systems - I

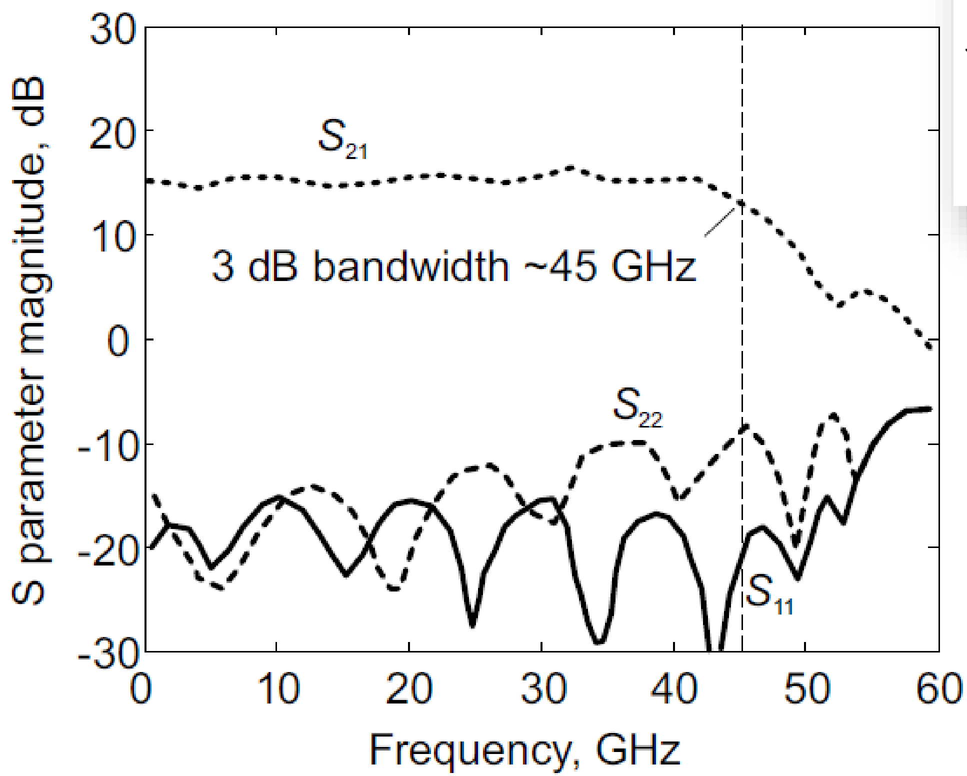




- PHEMT InGaAs/GaAs Fujitsu
- 40 Gbps, 6 Vpp

Fujitsu cascode DAMP for optical communication systems - II





H. Shigematsu, M. Sato, T. Hirose, and Y. Watanabe, “A 54-GHz distributed amplifier with 6-VPP output for a 40-Gb/s LiNbO3 modulator driver,” IEEE Journal of Solid-State Circuits, vol. 37, no. 9, pp. 1100–1105, Sep. 2002. Copyright 2002 IEEE, with permission.