Chapter 4-1

Microwave Amplifiers

Amplification is one of the basic functions in analog electronics. The task of designing a microwave amplifier consists first to ensure that the active device around which design is done will satisfy the circuit requirements, i.e.,

- Meet performance specifications that might otherwise be unattainable,
- Afford improved margins of process tolerance so that yield increases and
- Potentially save on the number of passive/active elements in the circuit, leading to lower cost and smaller size.

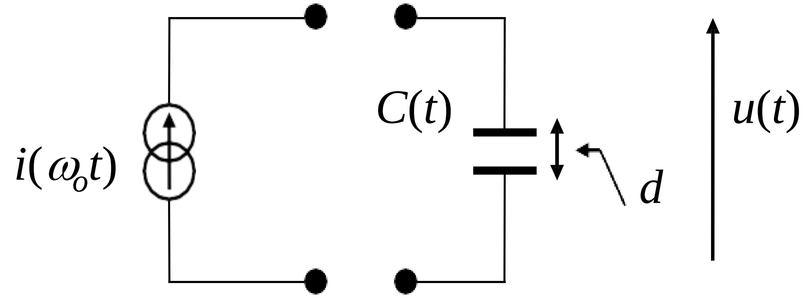
In the microwave area, such circuits can be built around diodes or transistors. Following the power criteria, amplifiers can be classified into: **low-noise amplifiers** (equivalent to a linear circuit), **linear/quasi-linear amplifiers** (designed through a quasi-linear approach) and **power amplifiers** (designed through a nonlinear approach).

PARAMETRIC AMPLIFIERS

We will first emphasize on parametric amplification, which will help us understanding the nonlinear effects in microwave circuits. For clarity, we will focus on one-port reactive components like varactors *that can be modeled by a single nonlinear capacitor*. Reactive device means no active losses will be considered.

DEFINITION OF PARAMETRIC AMPLIFICATION

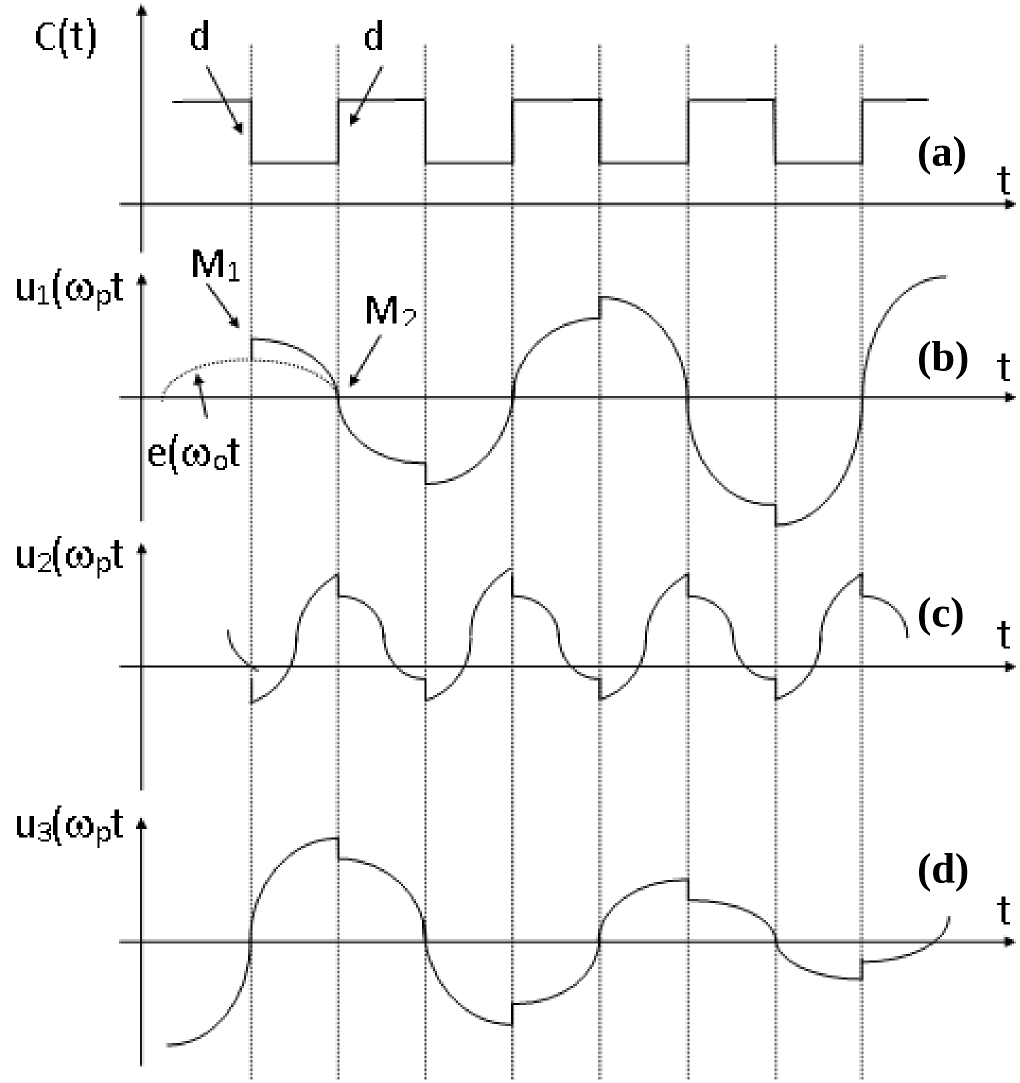
In order to highlight the parametric effect, let us consider the following circuit shown on this Figure where a capacitor C is excited by a current source $i(\omega_o t)$ or an equivalent voltage source $e(\omega_o t)$.

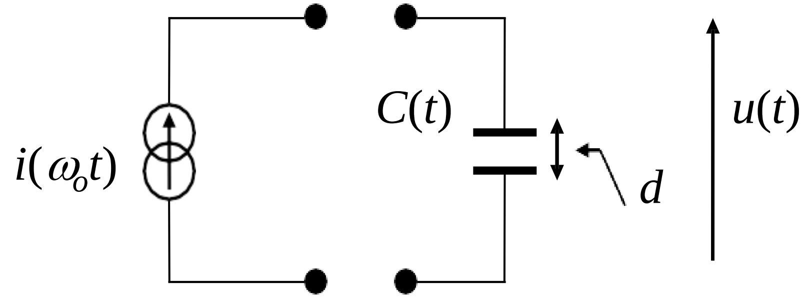


The capacitor has two conducting plates separated by an insulating layer of distance d that can be varied with a **pulsation ω_p**.

Thus, its value will increase and decrease accordingly (**Fig. a).**

Let ω_p be the pulsation of the pump and ω_o the excitation pulsation; we have different scenarios.





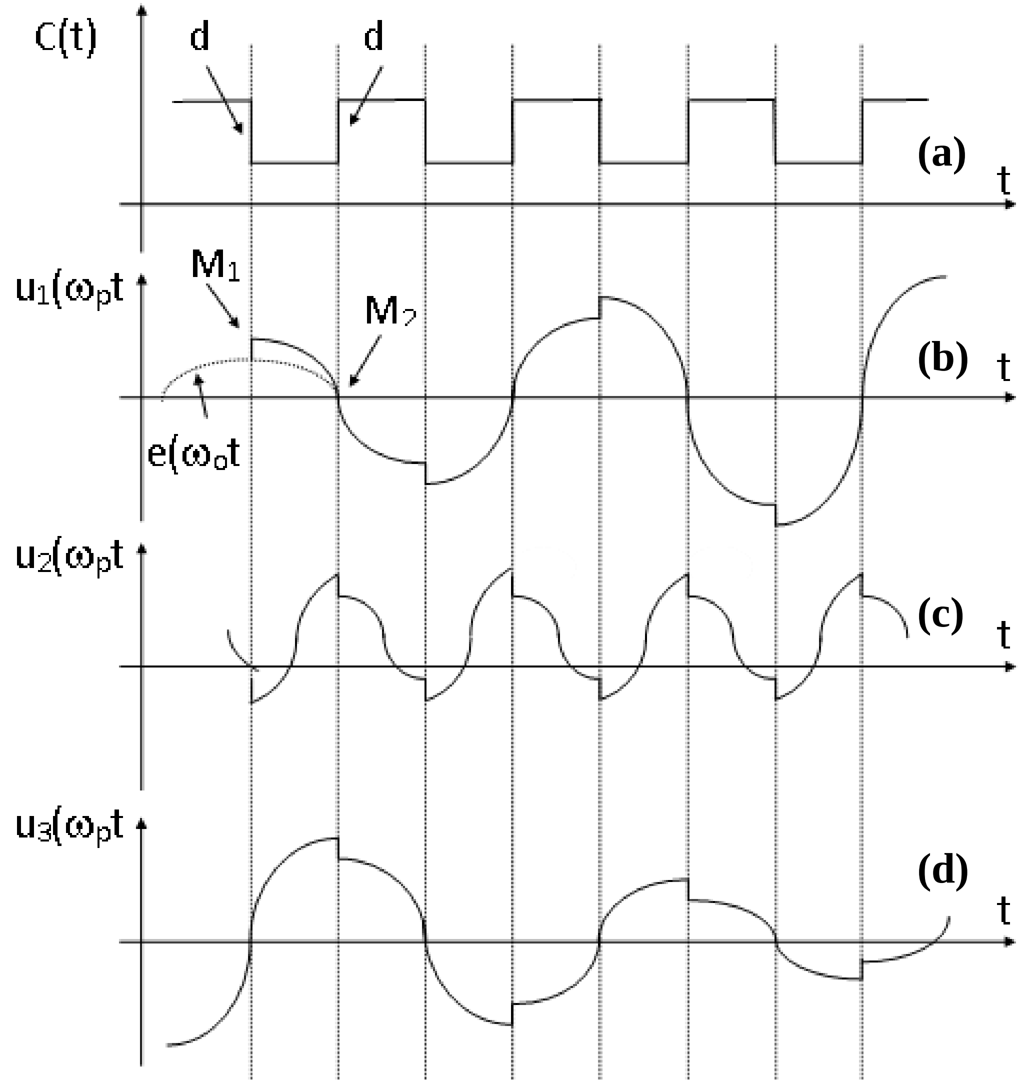
- The two pulsations **are in-phase**: we correspond the capacitance decreasing (larger *d*) with the point M_1 where the applied voltage is max and the capacitance increasing (smaller distance) with the point M_2 where the voltage is zero (turn back to the initial position).
  - → The voltage $u_1(\omega_p t)$ across C will always increase:

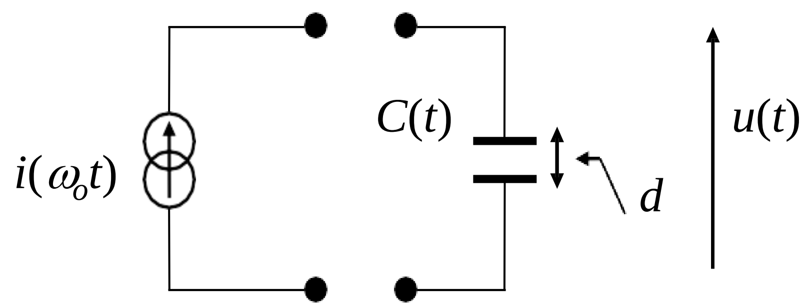
    Parametric amplification **(Fig. b).**

- ω_o and ω_p **are out of phase**:
  - → the amplitude increases and decreases periodically **(Fig. c)**

    Beating,
  - → and, at the limit, could reach

    Parametric attenuation **(Fig. d).**





(a) Variation of the capacitance $C(\omega_p t)$ in function of the distance d between the two planes

(b) The two pulsations ω_p and ω_o are in phase

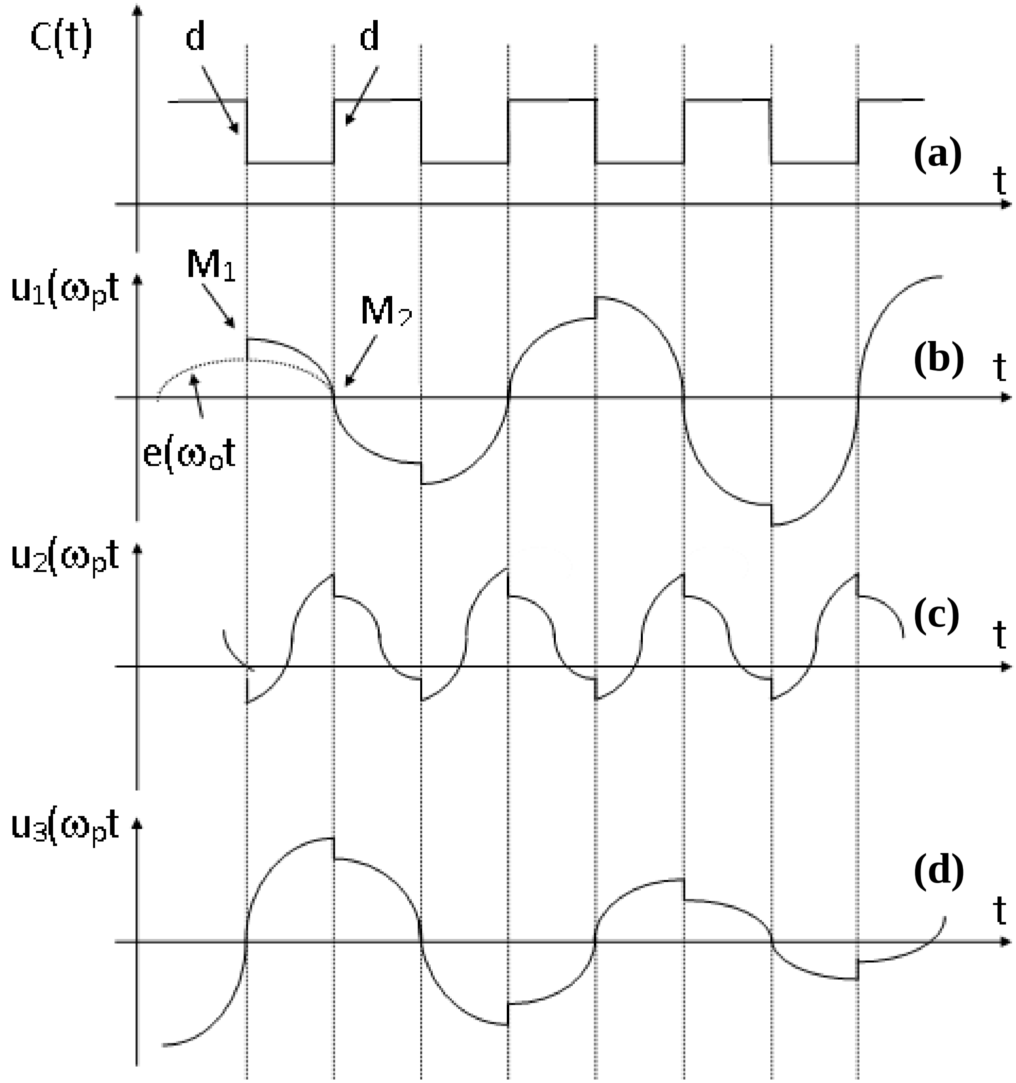
$\rightarrow$ Amplification

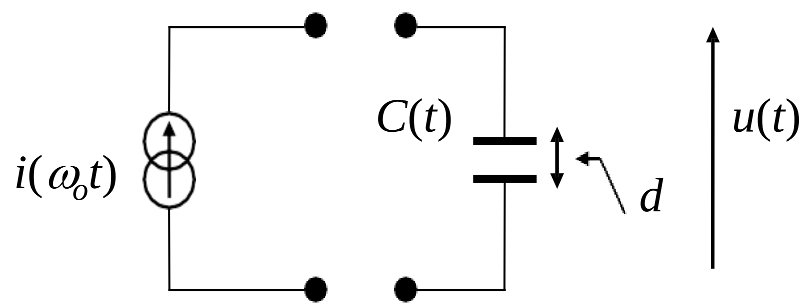
(c) The two pulsations ω_p and ω_o are out of phase

$\rightarrow$ Interference (beating)

(d) The two pulsations ω_p and ω_o are out of phase (limit case)

$\rightarrow$ Attenuation



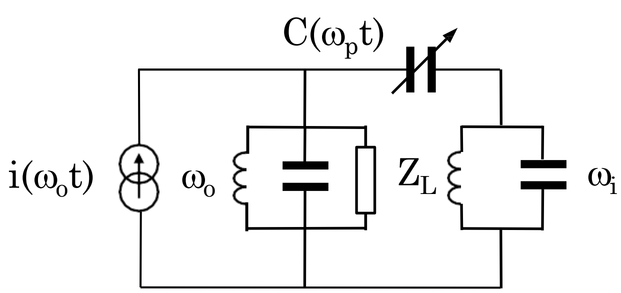


For a general case { $\omega_p \neq 2\ \omega_o$ }, it is impossible to predict the input phase as well as the phase and pulsation conditions { $\omega_p = 2\ \omega_o$ }; so, let us introduce two new frequencies

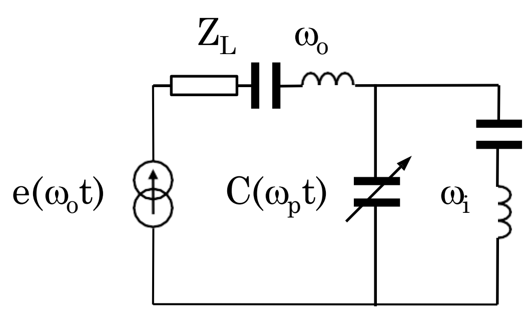
$$\frac{\omega_p}{2} - \left\{\frac{\omega_p}{2} - \omega_o\right\} = \omega_o \qquad\qquad \frac{\omega_p}{2} + \left\{\frac{\omega_p}{2} - \omega_o\right\} = \omega_p - \omega_o = \omega_i$$

We will then have an additional circuit of pulsation ω_i, called "**idler circuit**", allows to avoid the above conditions and to propose two configurations for a parametric amplifier in accordance with the source impedance value:

- excitation by a current source if the impedance is high.
- excitation by a voltage source if the impedance is low.



Parametric amplifier excited by a current source.



Parametric amplifier excited by a voltage source.

PARAMETRIC DEVICES

Many microwave diodes exhibit a nonlinear capacitance: IMPATTs, Varactors, Schottky, … but we will focus on the simplest one, namely the varactor diode. The dynamic excursion of the Q point varies from the breakdown voltage U_a (avalanche voltage) to the barrier voltage Φ so the diode is equivalent to a pure reactance.

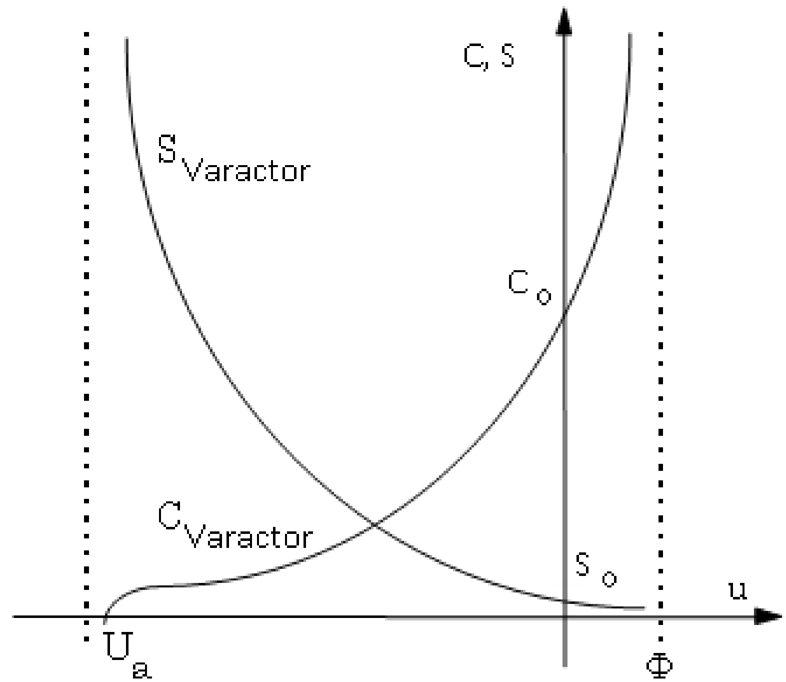
The capacitance variation $C(u(t))$ can be expressed by:

$$C_j(V_j) = C_{jo}\left\{1 - \frac{V_j(t)}{\Phi}\right\}^{-\gamma}$$

but it is **more convenient** to use the elastance $S(u(t))$ as

$$S(u) = \frac{d\,u(t)}{d\,q(t)} = S_o\left\{1 - \frac{u}{\Phi}\right\}^{\gamma}$$

S_o is the value of the elastance at $\{ u = 0 \}$.



more convenient?

PARAMETRIC CIRCUIT ANALYSIS

$C(u)$ is nonlinear, therefore, for each harmonic $k\omega$ of the signal ω, $u(t)$ or $q(t)$ can be written as

$$u(t) = \sum_{k=-\infty}^{\infty} U_k e^{jk\omega t} \qquad U_k = U_{-k}^* \qquad q(t) = \sum_{k=-\infty}^{\infty} Q_k e^{jk\omega t} \qquad Q_k = Q_{-k}^*$$

We must express the Fourier coefficients of the elastance as

$$S(t) = \sum_{k=-\infty}^{\infty} S_k e^{jk\omega t} \qquad S_k = S_{-k}^* \qquad S_k \neq \frac{1}{C_k}$$

Here, we have a large signal ω_p and a smaller one ω_o; thus the large-signal small-signal analysis can be applied. But since the circuit topology is simple, we can utilize simpler approaches to analyze the circuit. Let us express the charge in function of the elastance

$$q(t) = \frac{\Phi}{S_o(1-\gamma)}\left\{1-\left\{1-\frac{u}{\Phi}\right\}^{1-\gamma}\right\}$$

and consider the particular excitations:

$$q(\Phi) = Q_\Phi = \frac{\Phi}{S_o(1-\gamma)} \qquad q(U_a) = Q_a = Q_\Phi\left\{1-\left\{1-\frac{U_a}{\Phi}\right\}^{1-\gamma}\right\} \qquad q(t) = Q_\Phi\left\{1-\left\{1-\frac{u}{\Phi}\right\}^{1-\gamma}\right\}$$

Thus, we have, with the condition $\gamma \neq 1$

$$\left\{1-\frac{u}{\Phi}\right\}^{\gamma} = \left\{1-\frac{q}{Q_\Phi}\right\}^{\frac{\gamma}{1-\gamma}}$$

Moreover, using the maximum value of the elastance

$$S_{\max} = S(U_a) = S_o\left\{1-\frac{U_a}{\Phi}\right\}^{\gamma}$$

we can deduce the relationship between elastance and voltage

$$S(u) = S_{\max}\left\{\frac{u-\Phi}{U_a-\Phi}\right\}^{\gamma}$$

or elastance and charge $\quad S(q) = S_{\max}\left\{\frac{q-Q_\Phi}{Q_a-Q_\Phi}\right\}^{\frac{\gamma}{1-\gamma}}$

For abrupt junction ($\gamma = ½$): $\quad S(u) = S_{\max}\sqrt{\frac{u-\Phi}{U_a-\Phi}} \quad \rightarrow \quad$ Nonlinear

$$S(q) = S_{\max}\frac{q-Q_\Phi}{Q_a-Q_\Phi} \quad \rightarrow \quad \text{Linear}$$

Voltage pump is nonlinear while current pump is linear:

a user can decide on either linear or nonlinear pumping.

For gradual junction ($\gamma = 1/3$):

$$S(u) = S_{\max} \sqrt[3]{\frac{u - \Phi}{U_a - \Phi}} \quad \rightarrow \quad \text{Nonlinear}$$

$$S(q) = S_{\max} \sqrt[2]{\frac{q - Q_\Phi}{Q_a - Q_\Phi}} \quad \rightarrow \quad \text{Nonlinear}$$

Both sources are nonlinear.

The final step is to obtain the S_k coefficients. As these coefficients are functions of the two pulsations ω_p and ω_o (one with large and one with small magnitude), we will use the large-signal / small-signal analysis approach.

I - Large signal analysis: Excitation by one signal (the pump)

In the case of a current pumping, the charge could be developed in a series expansion as

$$q(t) = Q_{dc} + 2Q_{acp}\cos(\omega_p t) = Q_{dc} + Q_{acp}e^{j\omega_p t} + Q_{acp}e^{-j\omega_p t}$$

$$S(q) = S_{\max}\left\{\frac{Q_{dc}-Q_{\Phi}}{Q_a-Q_{\Phi}}\right\}^{\frac{\gamma}{1-\gamma}}\left\{1+\frac{2Q_{acp}}{Q_{dc}-Q_{\Phi}}\cos(\omega_p t)\right\}^{\frac{\gamma}{1-\gamma}} = S(Q_{dc})\{1+2\tau_q\cos(\omega_p t)\}^{\frac{\gamma}{1-\gamma}}$$

where τ_q is the current modulation factor (i.e., when the junction is excited by a current).

A similar form can be also found for a voltage excitation

$$S(u) = S_{\max}\left\{\frac{U_{dc}-\Phi}{U_a-\Phi}\right\}^{\gamma}\left\{1+\frac{2U_{acp}}{U_{dc}-\Phi}\cos(\omega_p t)\right\}^{\gamma} = S(U_{dc})\{1+2\tau_u\cos(\omega_p t)\}^{\gamma}$$

where τ_u is the voltage modulation factor (i.e., when the junction is excited by a voltage).

I - Large signal analysis: Excitation by one signal (the pump)

$$S(q)=S_{\max}\left\{\frac{Q_{dc}-Q_{\Phi}}{Q_a-Q_{\Phi}}\right\}^{\frac{\gamma}{1-\gamma}}\left\{1+\frac{2Q_{acp}}{Q_{dc}-Q_{\Phi}}\cos(\omega_p t)\right\}^{\frac{\gamma}{1-\gamma}}=S(Q_{dc})\{1+2\tau_q\cos(\omega_p t)\}^{\frac{\gamma}{1-\gamma}}$$

$$S(u)=S_{\max}\left\{\frac{U_{dc}-\Phi}{U_a-\Phi}\right\}^{\gamma}\left\{1+\frac{2U_{acp}}{U_{dc}-\Phi}\cos(\omega_p t)\right\}^{\gamma}=S(U_{dc})\{1+2\tau_u\cos(\omega_p t)\}^{\gamma}$$

Note that the above equations have the same format (according to α { α = γ or α = (γ /1−γ) }):

$$m(t)=\{1+2\tau\cos(\omega_p t)\}^{\alpha}=\sum_{k=-\infty}^{\infty}M_k e^{jk\omega_p t}$$

$$\omega_p t=\theta-\pi$$

$$x=\{1+4\tau\}^{-1/2}$$

$$M_k=\left\{\begin{array}{c}\dfrac{S_k}{S(Q_{dc})}\\ \text{or}\\ \dfrac{S_k}{S(U_{dc})}\end{array}\right\}=\frac{1}{\pi}\int_0^{\pi}\{1+2\tau\cos(\omega_p t)\}^{\alpha}\cos(k\omega_p t)\,d(\omega_p t)$$

I - Large signal analysis: Excitation by one signal (the pump)

$$m(t) = \{1 + 2\tau\cos(\omega_p t)\}^{\alpha} = \sum_{k=-\infty}^{\infty} M_k e^{j k \omega_p t}$$

$$M_k = \left\{ \begin{matrix} \dfrac{S_k}{S(Q_{dc})} \\ \text{or} \\ \dfrac{S_k}{S(U_{dc})} \end{matrix} \right\} = \frac{1}{\pi}\int_0^{\pi} \{1 + 2\tau\cos(\omega_p t)\}^{\alpha} \cos(k\omega_p t)\, d(\omega_p t)$$

The values of the integrals (deduced from the Legendre's polynoms) were calculated by Lowan

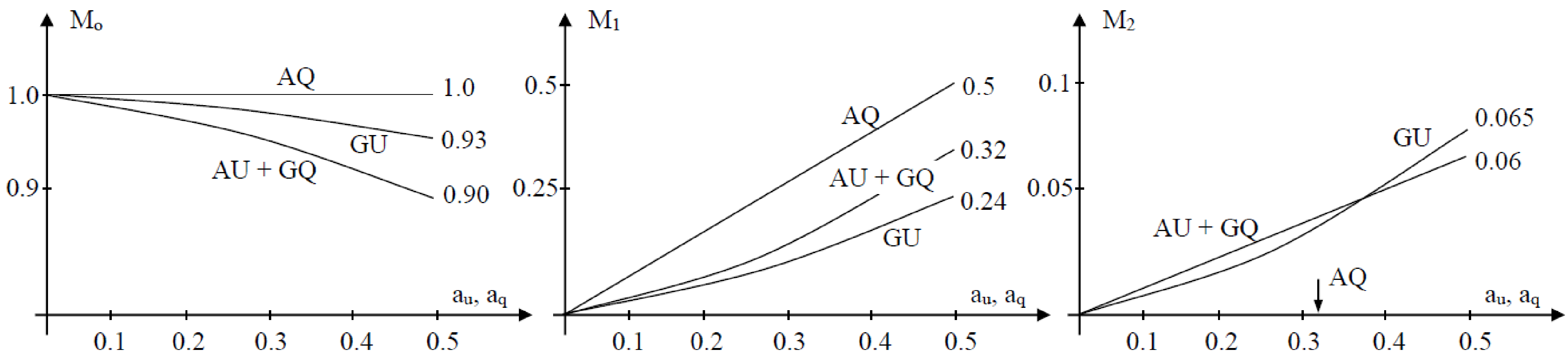
$$\int_0^{\pi} \left\{x + \sqrt{x^2 - 1}\cos\theta\right\}^{\alpha} \cos(k\theta)\, d\theta = \frac{\pi P_{\alpha}^{k}(x)}{(\alpha+1)\cdots(\alpha+k)}$$

$$M_k = \frac{(-1)^k (1 - 4a^2)^{\frac{\alpha}{2}}\, P_{\alpha}^{k}(1 - 4a^2)^{-\frac{1}{2}}}{(\alpha+1)\cdots(\alpha+k)}$$

S-parameter ratios (in %) for a = 0.5

$$M_k = \frac{(-1)^k \left(1-4a^2\right)^{\frac{\alpha}{2}} P_\alpha^k \left(1-4a^2\right)^{-\frac{1}{2}}}{(\alpha+1)\cdots(\alpha+k)}$$

Junction	Pump	S_2/S_1	S_3/S_1	S_4/S_1	S_5/S_1	S_6/S_1	S_7/S_1	S_8/S_1	S_9/S_1	S_{10}/S_1
G	U	28.6	14.3	8.8	6.7	4.5	3.4	2.8	2.3	1.9
G	Q	20.0	8.6	4.8	3.1	2.0	1.5	1.2	0.9	0.8
A	U	20.0	8.6	4.8	3.1	2.0	1.5	1.2	0.9	0.8
A	Q	-	-	-	-	-	-	-	-	-



Let $\boldsymbol{P_{m.n}}$ be the **average power at frequency $\boldsymbol{f_{m,n}}$**. It is noted positive if it is generated from a source and negative if it is dissipated in the circuit.

Following the principle of energy conservation and multiplying each power by a number equals to unity, we get:

$$\sum_m \sum_n P_{m,n} \left\{ \frac{m\omega_o + n\omega_p}{m\omega_o + n\omega_p} \right\} = 0 \quad \rightarrow \quad \omega_o \sum_m \sum_n \frac{m\, P_{m,n}}{\omega_{m,n}} + \omega_p \sum_m \sum_n \frac{n\, P_{m,n}}{\omega_{m,n}} = 0$$

Therefore, with

$$u(t) = \sum_m \sum_n U_{m,n}\, e^{j\omega_{m,n} t} \qquad\qquad q(t) = \sum_m \sum_n Q_{m,n}\, e^{j\omega_{m,n} t}$$

the powers are equal to

$$P_{m,n} = 2\,\mathrm{Re}\left\{ U_{m,n}\, I^*{}_{m,n} \right\} = -2\,\omega_{m,n}\, \mathrm{Im}\left\{ U_{m,n}\, Q^*{}_{m,n} \right\}$$

where **Re** and **Im** represent real part and imaginary part, respectively.

Let us introduce normalized powers:

$$\frac{P_{m,n}}{\omega_{m,n}} = -2\ \mathrm{Im}\left\{U_{m,n}\ Q^*_{m,n}\right\}$$

We can note that they are not dependent of ω_o and ω_p because the $U_{m,n}$ and $Q_{m,n}$ coefficients are function only of the transfer diode curve $C(t)$.

This equation can be then reformatted in order to show relationship between all different powers.

These relations are called the “Manley-Rowe" relations:

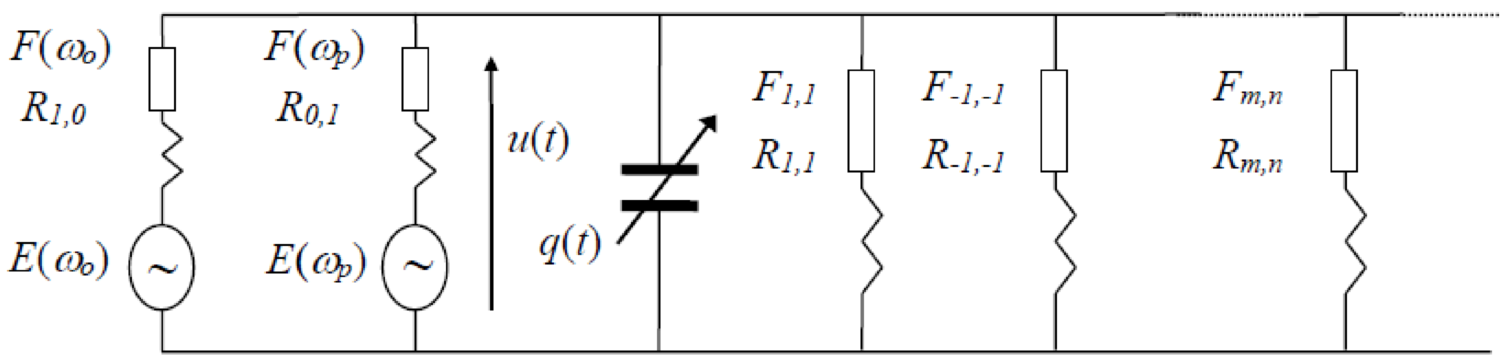
$$\left.\begin{aligned} \sum_{m=1}^{\infty}\sum_{n=-\infty}^{\infty}\frac{mP_{m,n}}{m\,f_o+n\,f_p} &= 0 \\ \sum_{m=-\infty}^{\infty}\sum_{n=1}^{\infty}\frac{nP_{m,n}}{m\,f_o+n\,f_p} &= 0 \end{aligned}\right\}$$

Summation boundaries are chosen so that each power $P_{m.n}$ is taken into account **only once**.

$$\left.\begin{aligned}\sum_{m=1}^{\infty}\sum_{n=-\infty}^{\infty}\frac{mP_{m,n}}{m f_o + n f_p} = 0\\ \sum_{m=-\infty}^{\infty}\sum_{n=1}^{\infty}\frac{nP_{m,n}}{m f_o + n f_p} = 0\end{aligned}\right\}$$

Note that these relations show that the source powers are converted into output powers at frequency $f_{m,n}$ but they did not show **how** the undesirable powers are dissipated.

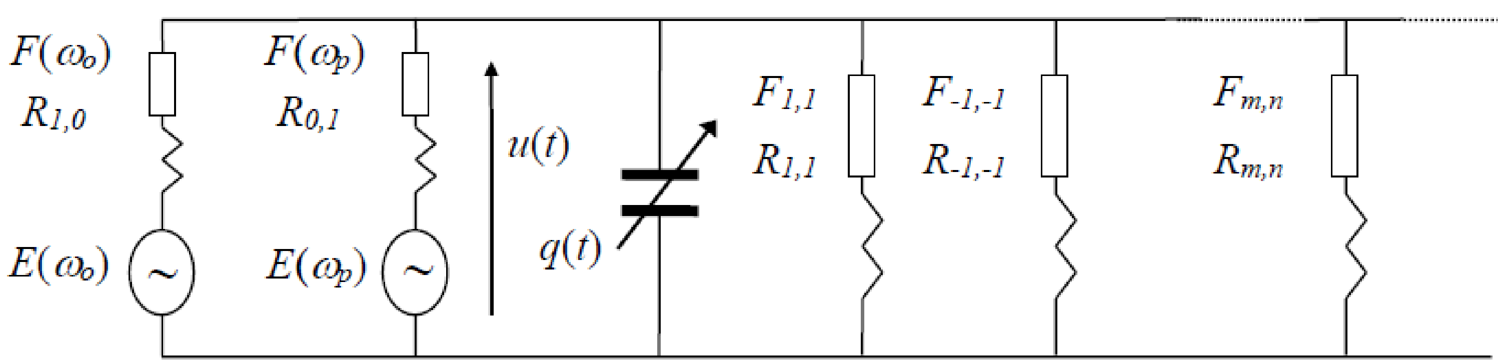
In practice, the diode is not ideal and this rule is played by the series resistances noted $R_{m,n}$ (where the $F_{m,n}$ are ideal pass-band filters centered on frequency $f_{m,n}$).



In general, for optimum efficiency, real powers due to undesirable frequencies must not be present.

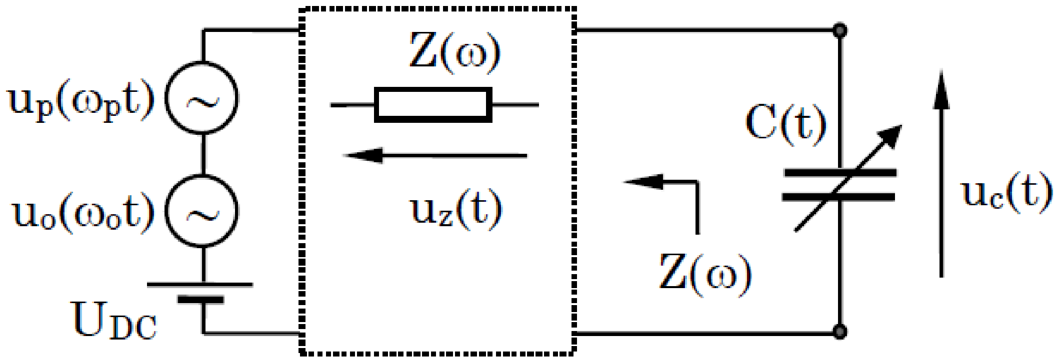
This condition is assured by loading the diode with reactive impedances at all frequencies except the desired one. But the series resistance R_s of the diode cannot allow this condition to be satisfied only if the loads are **open circuited** in order to eliminate the power dissipation in R_s. *This implies **a zero current** in the resistance for all undesirable frequencies, which is practically impossible to realize.*

An alternative approach consists to put short-circuits and then to provide power dissipation in the output network. In practice, this is done by adding circuits, called "**idlers**", for each undesirable frequency. But this will significantly complicate the final circuit, so that analyzing such circuits is relatively complex. Therefore, a more simplified, but approximate approach can be utilized, namely the coupled circuit analysis.



COUPLED CIRCUIT REPRESENTATION

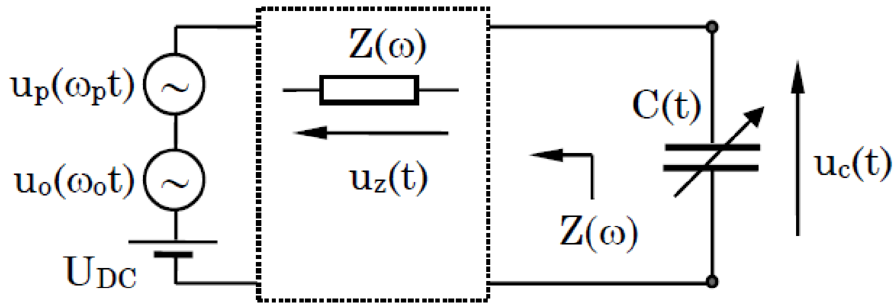
To analyze parametric amplifiers, the nonlinear element namely, the capacitance is separated from the rest of the circuit represented by its Thevenin's equivalent impedance noted $Z(\omega)$ [**Chapter 3**]



Thus, the source voltage u(t) can be expressed on the form of an immittance (*).

(*) The term *immittance* was invented by H. W. Bode. It refers to a complex number which may be either the impedance or the admittance of a system.

COUPLED CIRCUIT REPRESENTATION



$$u(t) = u_c(q(t)) - \int_0^t Z(t-\tau)\, q(\tau)\, d\tau = U_{dc} + u_p(\omega_p t) + u_o(\omega_o t)$$

$$= U_{dc} + 2U_{acp}\cos\left(\omega_p t\right) + 2U_{aco}\cos\left(\omega_p t + \varphi\right)$$

$$= U_{dc} + U_{acp}\, e^{j\omega_p t} + U_{acp}\, e^{-j\omega_p t} + \left\{U_{aco} e^{j\varphi}\right\} e^{j\omega_p t} + \left\{U_{aco}\, e^{j\varphi}\right\} e^{-j\omega_p t}$$

$$\Rightarrow \quad u(t) = U_{dc} + 2\,\mathbf{Re}\left\{U_p\, e^{j\omega_p t}\right\} + 2\,\mathbf{Re}\left\{U_o\, e^{j\omega_p t}\right\}$$

Integration, with initial conditions equal to zero, gives

$$u_c(q(t)) + Z(0)\, q(t) - \int_0^t Z(t-\tau)\, q(\tau)\; d\tau = u_p\left(\omega_p t\right) + u_o(\omega_o t) + Z(t-0)\, q(0)$$

$$u_c(q(t)) - \int_0^t Z(t-\tau)\, q(\tau)\; d\tau = u_p\left(\omega_p t\right) + u_o(\omega_o t)$$

To be solved using the large-signal small-signal analysis.

I.- Large-signal analysis for the nonlinear operator: Excitation by the pump

The operator F can be used to express the solution in the form of the Frechet's series:

$$F\left\lfloor q(\omega_p t)\right\rfloor = u(\omega_p t)$$

II.- Small signal analysis for the operator: Excitation by the pump and the signal

$$F\left\lfloor q_p(\omega_p t) + q_o(\omega_o t)\right\rfloor = u(\omega_p t) + u(\omega_o t)$$

$$F\left\lfloor q_p(\omega_p t) + q_o(\omega_o t)\right\rfloor - F\left\lfloor q_p(\omega_p t)\right\rfloor = u(\omega_o t)$$

The operator F can be expended using the Frechet's series:

$$dF\Big|_{q_p} q_o(t) = u(\omega_o t)$$

$$\Rightarrow \quad \frac{du_c}{dq}\Bigg|_{q_p} q_o(t) - \int_0^t Z(t-\tau) q_o(\tau)\, d\tau = u_o(\omega_o t)$$

III - Steady-state case

In the steady-state case, the above equation will be on the following form:

$$S(\omega_p t)\int_0^t i_o(\tau)d\tau + Z(\omega)i_o(\omega_o t) = u_o(\omega_o t)$$

which allows bringing out an impedance matrix

$$[\boldsymbol{Z}][\cdots \; I_{-1} \; I_0 \; I_1 \; \cdots]^t = [\cdots \; 0 \; U_0 \; 0 \; \cdots]^t$$

$$[\mathbf{Z}] = \begin{array}{c|c} \text{Line \# } m & \\ \begin{array}{c} \vdots \\ -1 \\ 0 \\ 1 \\ \vdots \end{array} & \begin{bmatrix} & \vdots & \vdots & \vdots & \\ \cdots & Z_{-1,-1} & Z_{-1,0} & Z_{-1,1} & \cdots \\ \cdots & Z_{0,-1} & Z_{0,0} & Z_{0,1} & \cdots \\ \cdots & Z_{1,-1} & Z_{1,0} & Z_{1,1} & \cdots \\ & \vdots & \vdots & \vdots & \end{bmatrix} \\ & \begin{array}{ccccc} \cdots & -1 & 0 & 1 & \cdots \end{array} \; \text{Column \# } n \end{array}$$

$$Z_{m,n} = \begin{cases} \dfrac{S_o}{j\omega_m} + Z(\omega_m) & m = n \\[2ex] \dfrac{S_{m-n}}{j\omega_n} & m \neq n \end{cases}$$

where the exponent *t* indicates the transpose matrix and where ω_n can be equivalent to $\omega_{m,n}$ upon the condition where the amplitude of the pump signal is much more greater than the one of the input signal:

$$\omega_{m,n} = m\omega_o + n\omega_p \approx \omega_o + n\omega_p = \omega_n$$

Now, a question may come to mind:

WHY PARAMETRIC AMPLIFIERS ?

Advantages ?

- **They can exhibit a very small noise figure.**
- **They are cost effective.**

Disadvantages ?

- **They are based on “old” devices (Varactors).**
- **They use diodes while transistor amplifiers are much more competitive.**
- **Recent transistors can exhibit quite low noise figure.**
- **Their theory is so complicated, making their design uncertain.**

So … obsolete ??

IEEE database:

419 papers on **parametric amplifiers : 1958 - 1970 … in 12 years ≈ *35 papers / year***

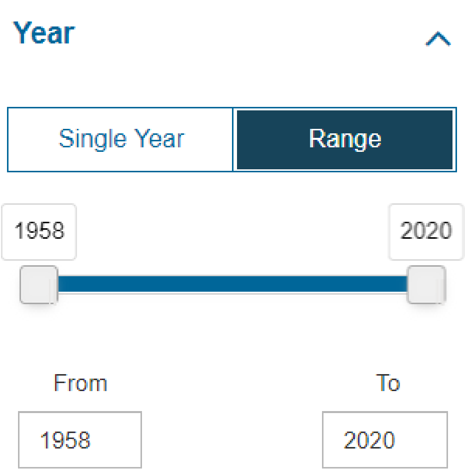
574 papers on **parametric amplifiers : 1970 - 2000 … in 30 years ≈ *19 papers / year***

In comparison:

4,139 papers on **transistors amplifiers** during the same period **(1970 - 2000)**

Sure … obsolete !!

Perhaps … not !



3,743 papers on **parametric amplifiers : 1958 - 2020**

IEEE database:

419 papers on **parametric amplifiers : 1958 - 1970 … in 12 years ≈ 35 paper / year**

574 papers on **parametric amplifiers : 1970 - 2000 … in 30 years ≈ 19 paper / year**

1,415 papers on **parametric amplifiers : 2000 - 2010**

1,623 papers on **parametric amplifiers : 2010 - 2020**

Applications?

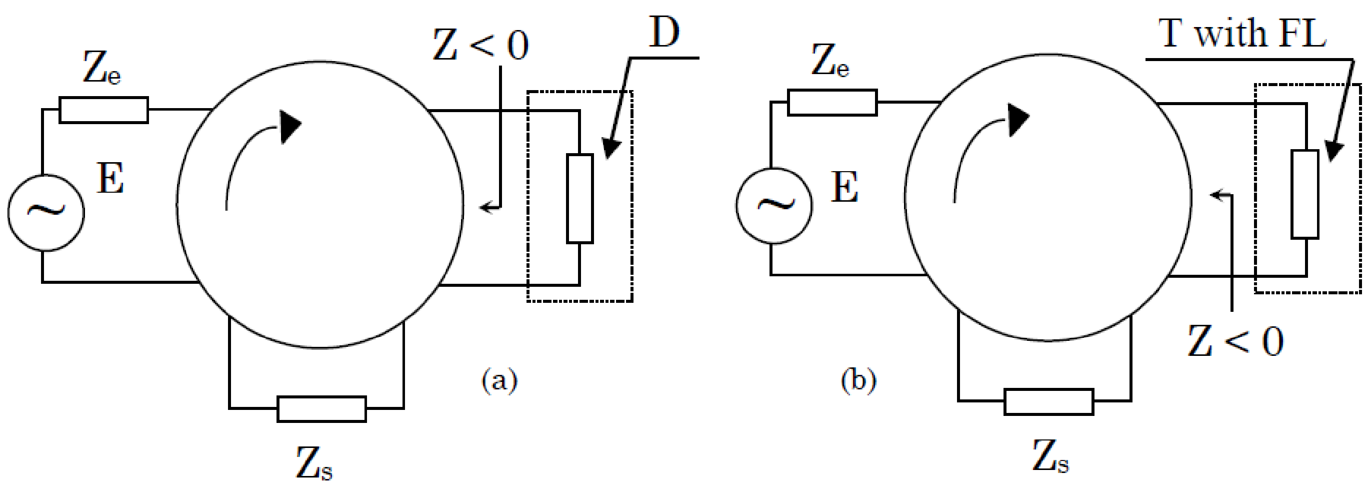
Optical Parametric Amplifiers for Applications in Modern Communication Networks

Coming back to the justifications provided:

- **They are based on old devices (Varactors) while transistor amplifiers are so competitive.**
  - **Recent optical parametric generators and amplifiers (OPG/OPA) offer the unique possibility to produce pico- and femtosecond pulses based on *transistors*.**
- **Their theory is so complicated**
  - **Nowadays, CAD tools are so efficient. Also, even if the theory is still based on the same foundations, the design of OPG/OPA devices is relatively mature.**

TRANSISTOR-BASED PARAMETRIC AMPLIFICATION

Transistors are also used for parametric amplification. In this case, the negative resistance is obtained by a feedback loop, which makes the transistor unstable. Compared to usual transistor amplifiers, these amplifiers are more stable, exhibit better *gain*Bandwidth* product, and present less sensitivity to transistor parameters variations.



Diode (D) amplifier. Transistor amplifier (T) with the feedback loop (FL).

Parametric amplifier with coupling by a circulator

DIFFERENT TYPES OF AMPLIFIERS

TRANSISTOR AMPLIFIERS

The most important design considerations in a microwave transistor amplifier are stability, power gain, bandwidth, noise and dc requirements. Usually, transistor amplifiers can be classified as

- reactively matched,
- lossy matched,
- feedback,
- distributed,
- balanced
- Doherty …

REACTIVELY MATCHED AMPLIFIERS

Reactively matched circuits are the **most common types** of amplifiers. In fact, circuits like "wideband amplifiers" and "broadband amplifiers" refer to amplifiers with a bandwidth of typically one or two octaves, i.e., that essentially have a pass-band characteristic.

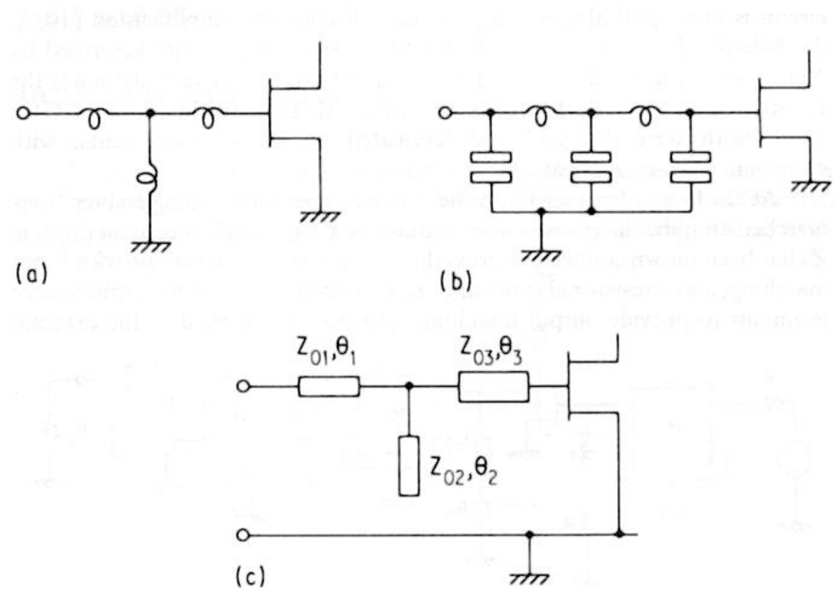
Such amplifiers are often of the reactively matched type and find application in phased-array radars, electronic counter measure systems, civil radars and satellite communication systems.

Amplifiers having **bandwidths** of a **few tens of percent** mostly employ **reactive matching**, and find application in communications such as direct broadcast from satellite systems.

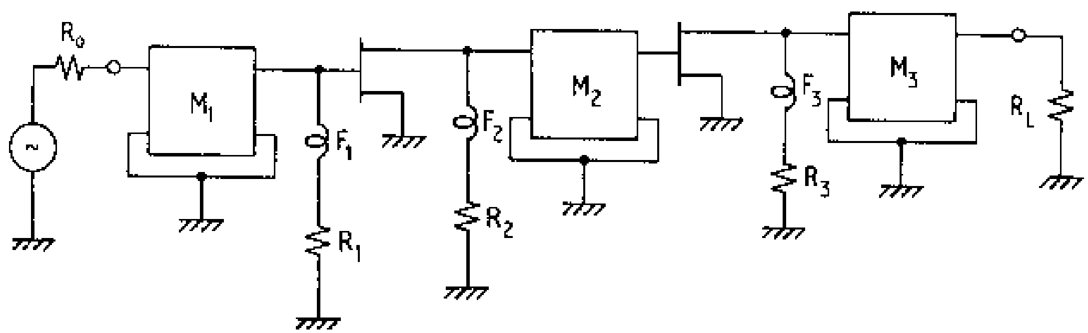
On the other hand, **ultra-broadband amplifiers**, as a class, have a bandwidth of **several octaves**, and often have a low-frequency response down to dc, or at most a few hundred megahertz. Those circuits, which respond at dc, can rightly be regarded as low-pass. They are required for electronic warfare systems, guided weapons, instrumentation and pulse amplifiers.

LOSSY MATCHED AMPLIFIERS

So instead of using reactive matching networks …



one means of improving the match at low frequencies is to provide resistive elements, which become more tightly coupled to the main signal path through the amplifier at low frequencies.

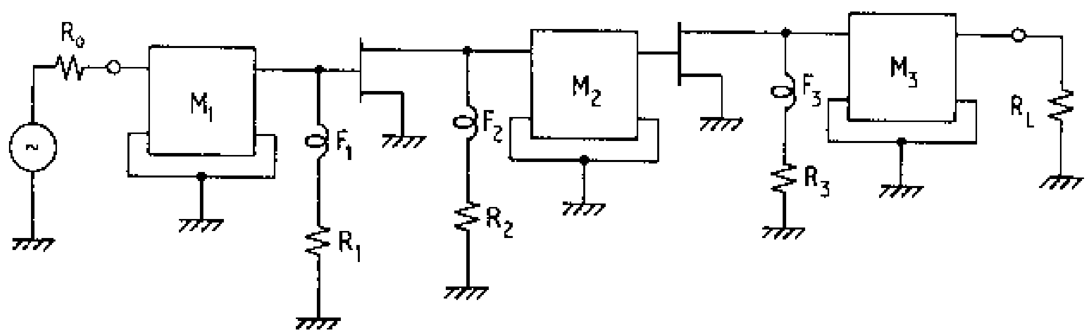


LOSSY MATCHED AMPLIFIERS

M_i are lossless impedance matching networks and F_i are low-pass lossless impedance transformers.

At the higher range of the specified frequency band, the transformers transform the value of R_i into high impedances that do not load the signal path through the amplifier. In this limit, the circuit behaves as a reactively matched amplifier.

The gain of the amplifier is then dependent upon the resistor values. Therefore, a lossy matched amplifier consists of two circuits both having the same gain: **a resistively-coupled** amplifier providing gain and matching at the lower frequencies in the band to be covered, and a **reactively-matched** amplifier providing gain and matching at the higher frequencies. The circuit is thus applicable to flat-gain, ultra-broadband amplification frequencies.



FEEDBACK AMPLIFIERS

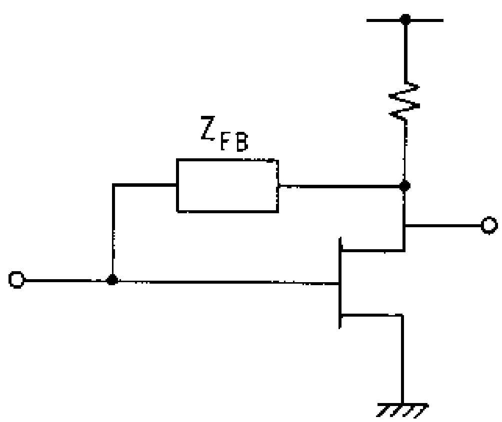
Feedback amplifiers could be designed through a first-trial method, finding initial circuit values that can then be iterated by computer methods to achieve the desired performance. The connection has the following attributes:

(a) It offers gain equalization over a fairly wide bandwidth,

(b) It can improve the $|S_{21}|$ at the highest frequencies, compared with the simple FET,

(c) The matches can be improved concurrently with (a) and (b), thus simplifying the task of matching network design,

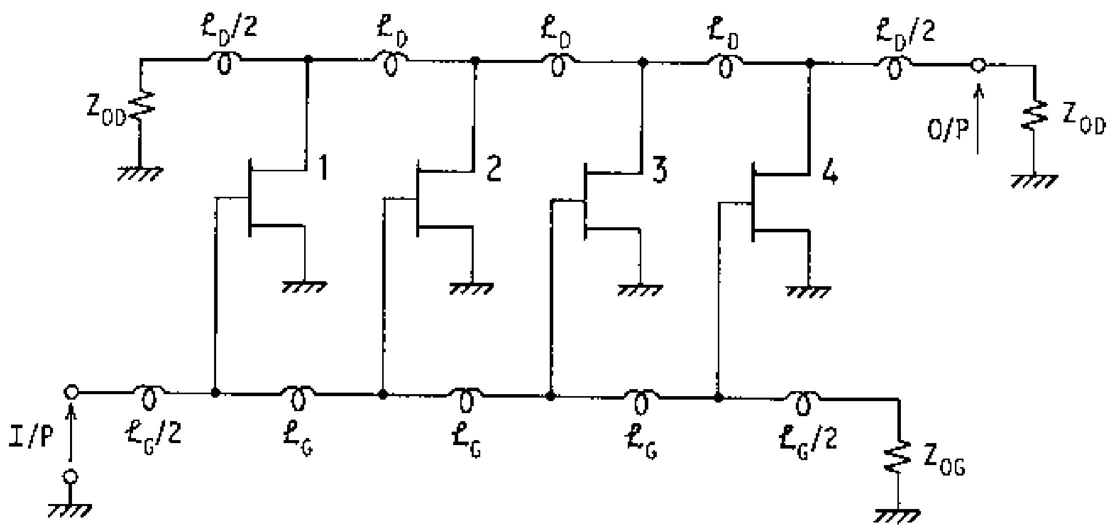
(d) It reduces the sensitivity of the overall circuit to device variations



DISTRIBUTED AMPLIFIERS

With a conventional amplifier, any attempt to increase the gain by increasing the FET transconductance usually results in an increase in input capacitance as well.

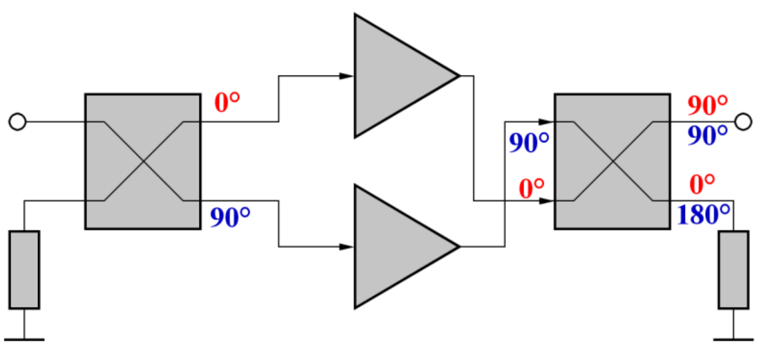
The distributed amplifier, also called the travelling wave amplifier, offers a means of combining the transconductance of several FETs without combining their input capacitances.

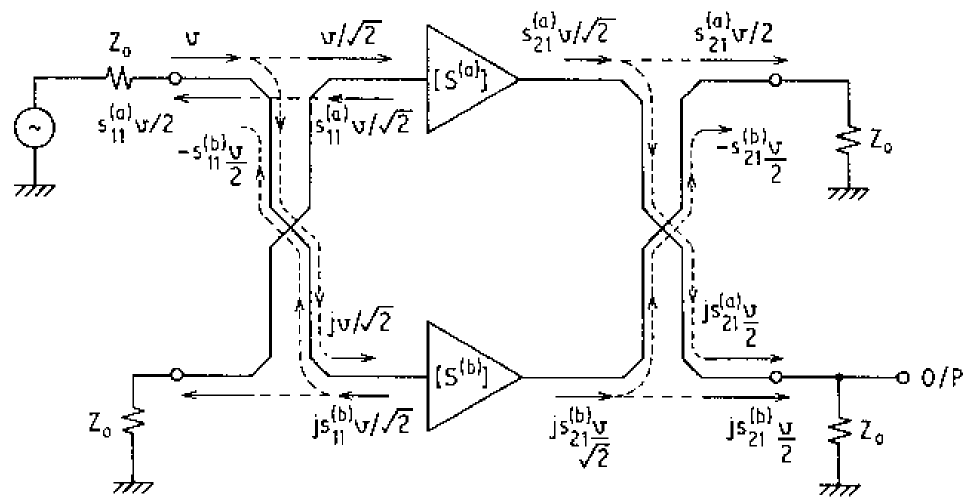


BALANCED AMPLIFIERS

Usually, the input reflection coefficient for a wideband reactively matched amplifier can easily be 0.8 or greater, i.e. a return loss as high as –2dB. To prevent the reflected energy giving rise to standing waves on any transmission line between the source and the amplifier, and to prevent the energy being dissipated in the source, a balanced amplifier configuration can be employed.

Two quadrature couplers are used: one at the input, to split the signal from the generator (source) into two equal parts to be fed to two amplifiers operating in parallel, the other to combine the outputs of the amplifiers into a single load. If the amplifier is truly balanced, $S_{11}^{(a)} = S_{11}^{(b)}$, and $S_{22}^{(a)} = S_{22}^{(b)}$.

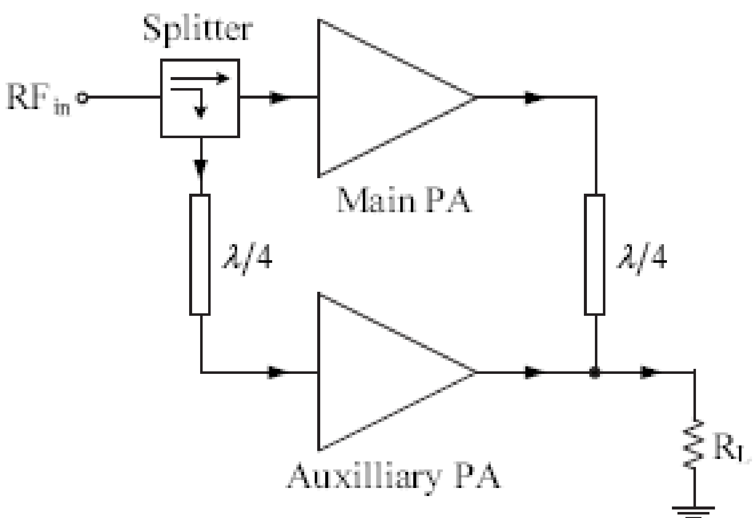


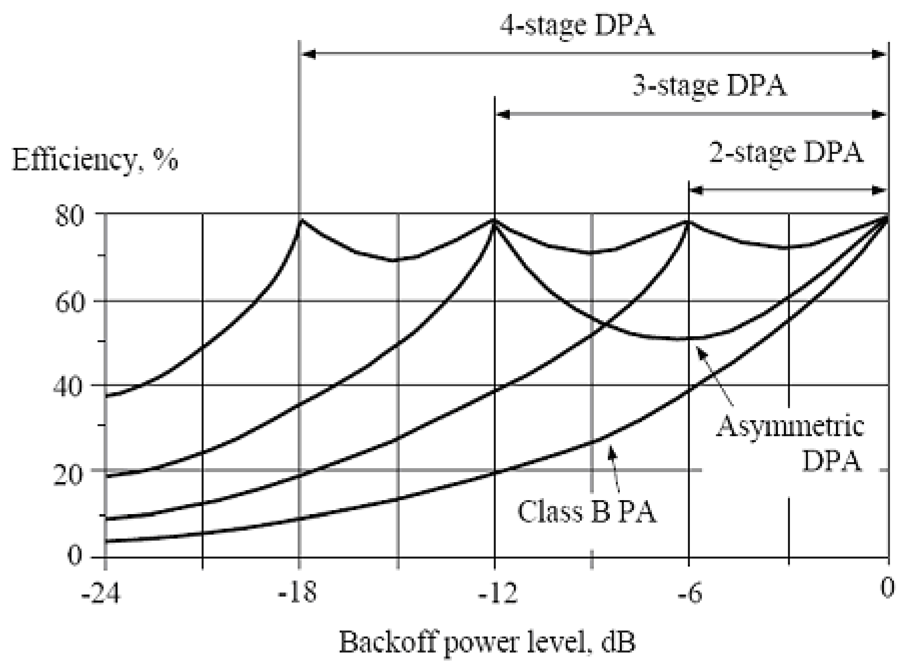


DOHERTY AMPLIFIERS

W.H. Doherty first introduced the Doherty technique in 1936, originally designed using vacuum tubes.

This is one of the most implemented techniques today for improving efficiency at back-off output power levels. It involves the implementation of efficiency enhancement on a power amplifier circuit that requires linear amplification. The most conventional configuration consists of two amplifiers, namely the main and the peaking ("auxiliary" amplifier). The amplifiers are connected in parallel with their outputs joined by a quarter-wave transmission line, which performs impedance transformation.





LINEAR AMPLIFIERS :

- **Low noise amplifiers**
- **High gain amplifiers**

TRANSISTOR AMPLIFIERS

As a guide, reactively-matched amplifiers allow the most gain to be derived from a given FET (and they also offer the lowest noise).

Design of an amplifier usually starts with a set of specifications and the selection of the proper transistor. Then, a systematic mathematical solution, aided by graphical methods, is developed to determine the transistor loading (source and load reflection coefficients) for a particular stability and gain criteria.

But what is the meaning of the term "gain"?

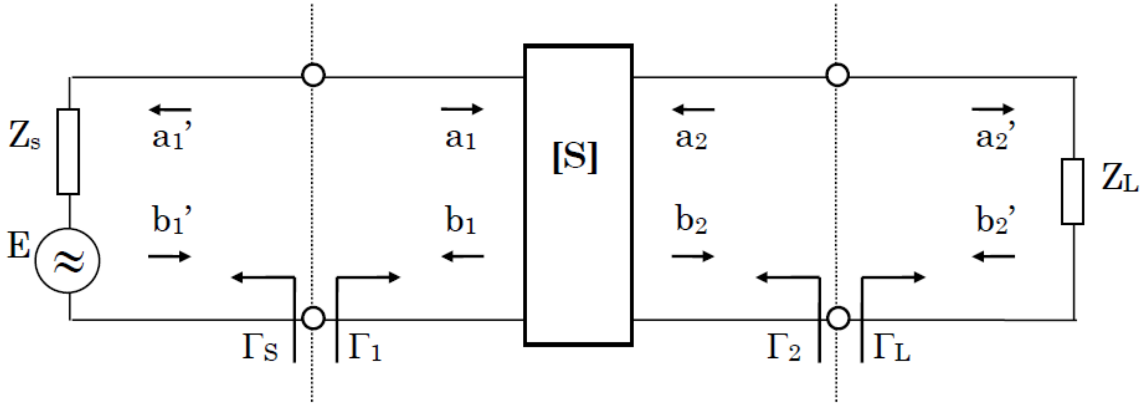
We will base our definitions ***on the reactively matched transistor configuration***.

Let us use the S-parameters

$$b_1 = S_{11}\,a_1 + S_{12}\,a_2 \qquad\qquad b_2 = S_{21}\,a_1 + S_{22}\,a_2$$

where a_k and b_k are the incident and reflected waves of port k respectively.

Now, let us consider the simple circuit below.

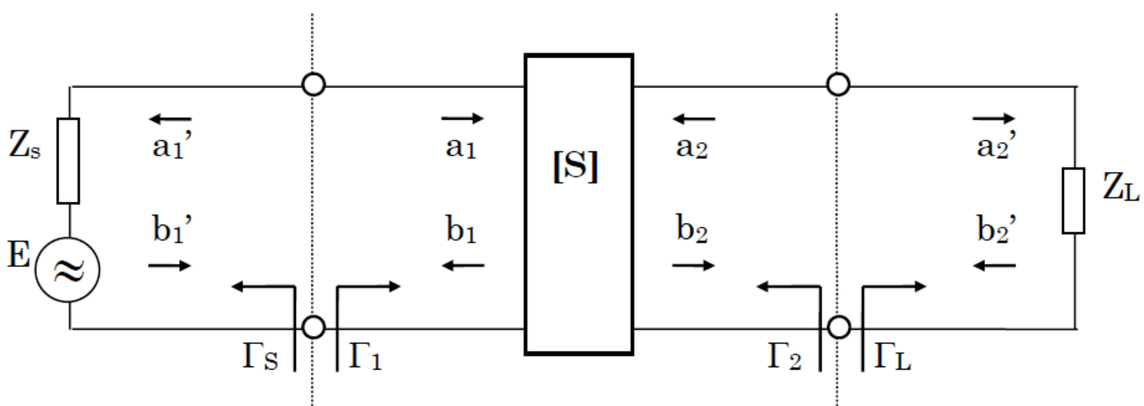


The transistor is exited by a voltage source of magnitude E and internal impedance Z_s and loaded by Z_L. The reflection coefficients of these impedances are noted Γ_S and Γ_L respectively.

From this mathematical representation, we can introduce some definitions of powers used in amplifier design:

- Power Delivered to a load: The power delivered to the load is defined as the difference between the incident and reflected power namely,

$$P_L = \frac{1}{2}|b_2|^2 - \frac{1}{2}|a_2|^2 = \frac{1}{2}|b_2|^2\left(1-|\Gamma_L|^2\right)$$



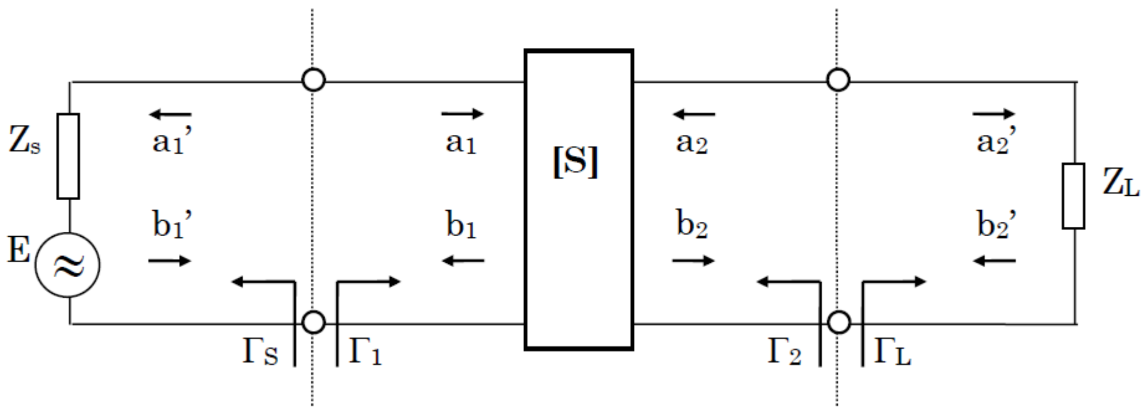
- Power available from a source: The power available from a source is defined as the power delivered by a source to the conjugately matched load, i.e.,

$$\Gamma_L = \Gamma_S^*$$

$$P_{AV} = \frac{1}{2}\left|b_1'\right|^2 - \frac{1}{2}\left|a_1'\right|^2 = \frac{\frac{1}{2}\left|b_1\right|^2}{1-\left|\Gamma_S\right|^2}$$

- Input power to the transistor: the input power to the transistor is defined as

$$P_{IN} = P_1 = \frac{1}{2}\left|a_1\right|^2 - \frac{1}{2}\left|b_1\right|^2 = \frac{1}{2}\left|a_1\right|^2\left(1-\left|\Gamma_1\right|^2\right)$$



Gain definitions

The transducer power gain G_T, is the ratio of the power delivered to a load to the power available from the source

$$G_T = \frac{P_L}{P_{AV}} = \frac{|b_2|^2}{|b_1|^2}\left(1-|\Gamma_L|^2\right)\left(1-|\Gamma_S|^2\right)$$

Similarly, the power gain is defined as the ratio of the power delivered to the load to the input power of the network

$$G_P = \frac{P_L}{P_1} = \frac{|b_2|^2}{|a_1|^2}\frac{\left(1-|\Gamma_L|^2\right)}{\left(1-|\Gamma_1|^2\right)}$$

Finally the available power gain G_A is the ratio of the power available from the transistor (P_2) to the power available from the source. The power available from the transistor is the power delivered by the transistor to a conjugate matched load:

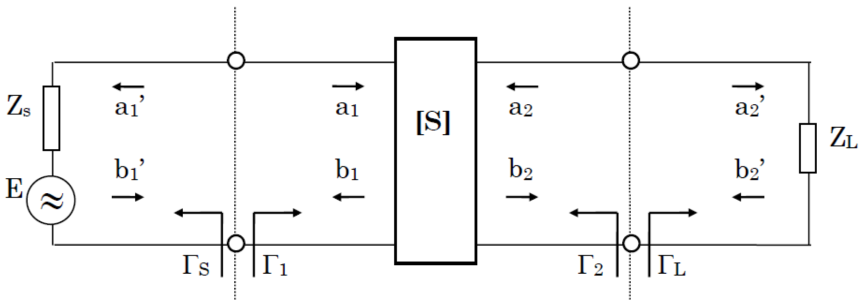
$$P_2 = \left[\frac{1}{2}|b_2|^2 - \frac{1}{2}|a_2|^2\right]_{\Gamma_L = \Gamma_2^*} = \frac{1}{2}|b_2|^2\left(1-|\Gamma_2|^2\right) \qquad G_A = \frac{1-|\Gamma_S|^2}{|1-S_{11}\,\Gamma_S|^2}|S_{21}|^2\frac{1}{1-|\Gamma_2|^2}$$

http://web.cecs.pdx.edu/~brano/ee431/ee431.html

$$G_T = \frac{P_L}{P_{avs}} = \frac{\text{power delivered to load}}{\text{power available from source}} = \text{ transducer}$$

$$G_p = \frac{P_L}{P_{in}} = \frac{\text{power delivered to load}}{\text{power input to network}} = \text{ operating}$$

$$G_A = \frac{P_{avn}}{P_{avs}} = \frac{\text{power available from network}}{\text{power available from source}} = \text{ available}$$



$$G_T = \frac{1-|\Gamma_s|^2}{|1-\Gamma_{IN}\Gamma_s|^2}|S_{21}|^2\frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2}$$

$$G_T = \frac{1-|\Gamma_s|^2}{|1-S_{11}\Gamma_s|^2}|S_{21}|^2\frac{1-|\Gamma_L|^2}{|1-\Gamma_{OUT}\Gamma_L|^2}$$

$$G_P = \frac{1}{1-|\Gamma_{IN}|^2}|S_{21}|^2\frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2} = f(\Gamma_L, [S])$$

$$G_A = \frac{1-|\Gamma_s|^2}{|1-S_{11}\Gamma_s|^2}|S_{21}|^2\frac{1}{1-|\Gamma_L|^2} = f(\Gamma_s, [S])$$

$$\Gamma_{IN} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1-S_{22}\Gamma_L} \quad \Gamma_{OUT} = S_{22} + \frac{S_{12}S_{21}\Gamma_s}{1-S_{11}\Gamma_s}$$

- Source and load mismatch factors:

$$M_s = \frac{P_{IN}}{P_{AVS}} = \frac{(1-|\Gamma_s|^2)(1-|\Gamma_{IN}|^2)}{|1-\Gamma_s\Gamma_{IN}|^2}\left(= \frac{4R_sR_{IN}}{|Z_s+Z_{IN}|^2}\right)$$

$$M_L = \frac{P_L}{P_{AVN}} = \frac{(1-|\Gamma_L|^2)(1-|\Gamma_{OUT}|^2)}{|1-\Gamma_L\Gamma_{OUT}|^2}\left(= \frac{4R_{OUT}R_L}{|Z_{OUT}+Z_L|^2}\right)$$

Use M_s and M_L to relate various powers and gains:

$$P_{IN} = P_{AVS}M_s = P_{AVS}|_{\Gamma_{IN}=\Gamma_s^*}$$

$$P_L = P_{AVN}M_L = P_{AVN}|_{\Gamma_L=\Gamma_{OUT}^*}$$

$$G_T = \frac{P_L}{P_{AVS}} = \frac{P_L}{P_{IN}}\frac{P_{IN}}{P_{AVS}} = G_p\frac{P_{IN}}{P_{AVS}} = G_pM_s$$

$$G_A = \frac{P_{AVN}}{P_{AVS}} = \frac{P_L}{P_{AVS}}\frac{P_{AVN}}{P_L} = \frac{G_T}{M_L}$$

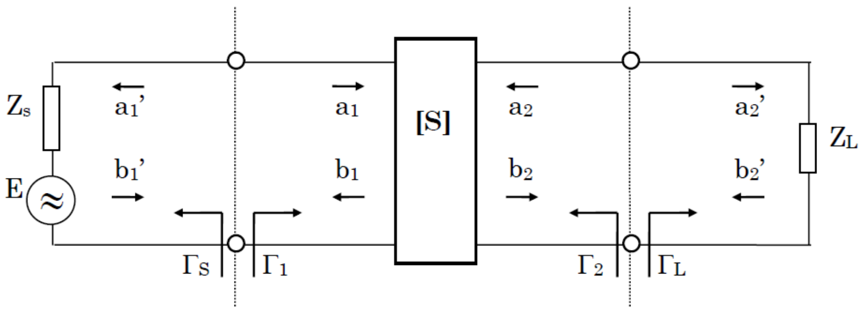
$$G_T = \frac{1-|\Gamma_s|^2}{|1-\Gamma_{IN}\Gamma_s|^2}|S_{21}|^2\frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2}$$

$$G_T = \frac{1-|\Gamma_s|^2}{|1-S_{11}\Gamma_s|^2}|S_{21}|^2\frac{1-|\Gamma_L|^2}{|1-\Gamma_{OUT}\Gamma_L|^2}$$

$$G_P = \frac{1}{1-|\Gamma_{IN}|^2}|S_{21}|^2\frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2} = f(\Gamma_L,[S])$$

$$G_A = \frac{1-|\Gamma_s|^2}{|1-S_{11}\Gamma_s|^2}|S_{21}|^2\frac{1}{1-|\Gamma_L|^2} = f(\Gamma_s,[S])$$

$$\Gamma_{IN} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1-S_{22}\Gamma_L} \quad \Gamma_{OUT} = S_{22} + \frac{S_{12}S_{21}\Gamma_s}{1-S_{11}\Gamma_s}$$



AMPLIFIER DESIGN: LINEAR AMPLIFIERS

Any amplifier design is mainly based on the right choice of source and load impedances Z_S and Z_L. This is essential since the amplifier performance and parameters (gain, input/output impedance, etc.) are function of these impedance values.

The purpose of any linear amplifier design is to determine the appropriate reflection coefficients Γ_S and Γ_L (and then, the corresponding impedances).

For linear and quasi-linear amplifiers, this design is achieved from the small-signal S-parameters of the active component, i.e., the transistor.

AMPLIFIER DESIGN: LINEAR AMPLIFIERS

Stability

The **stability** of an amplifier, or its resistance to oscillate, is a fundamental consideration in a design and can be determined from the S-parameters, the matching networks, and the terminations:

- Making an amplifier stable is to load it with source and load impedances exhibiting real positive parts **($|\Gamma_S| < 1$ and $|\Gamma_L| < 1$).** In other words, the loads should be passive.

Matching

The **matching** of an amplifier is its ability to transfer a maximum of power from the source to the transistor and from the transistor to the load:

- Matching an amplifier means to built a **maximum power transfer** from source to circuit and from circuit to load.

AMPLIFIER DESIGN: LINEAR AMPLIFIERS

Let us use the S-parameters $b_1 = S_{11}\,a_1 + S_{12}\,a_2 \quad b_2 = S_{21}\,a_1 + S_{22}\,a_2$

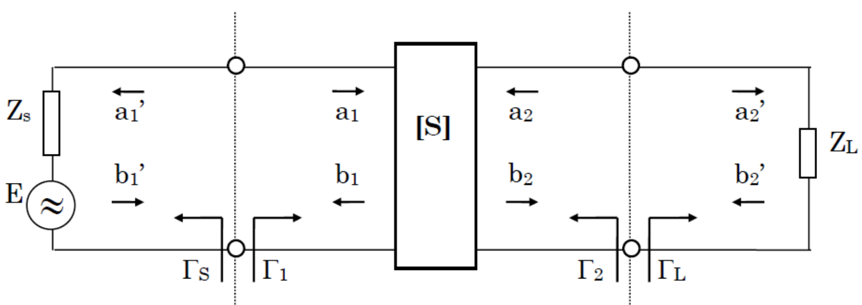
where a_k and b_k are the incident and reflected waves of port k respectively

Let Γ_1 be the reflection coefficient at the transistor input when its output port is matched ($a_2 = 0$) and Γ_2 be the reflection coefficient at the transistor output when the input port is matched ($a_1 = 0$).

$$\Gamma_1 = \frac{b_1}{a_1} = \Gamma_S^* \qquad \Gamma_2 = \frac{b_2}{a_2} = \Gamma_L^*$$

CONDITIONS OF MATCHING

where the (*) sign indicates the complex conjugate.



AMPLIFIER DESIGN: LINEAR AMPLIFIERS

$$b_1 = S_{11}\,a_1 + S_{12}\,a_2 \qquad\qquad b_2 = S_{21}\,a_1 + S_{22}\,a_2$$

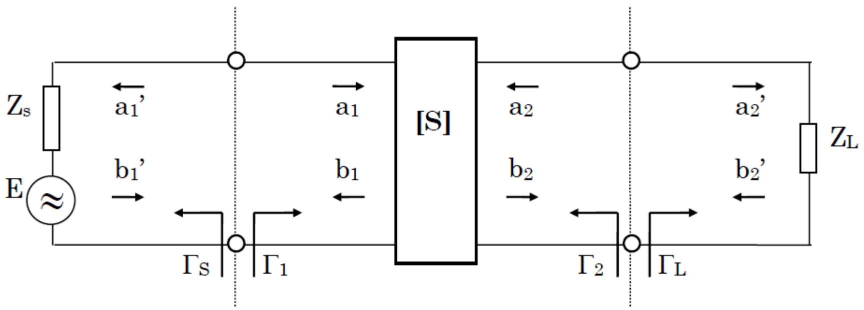
$$\Gamma_1 = \frac{b_1}{a_1} = \frac{S_{11}\,a_1 + S_{12}\,a_2}{a_1} = S_{11} + S_{12}\,\frac{a_2}{a_1}$$

$$a_2 = \Gamma_L b_2 = \Gamma_L\left(S_{21}\,a_1 + S_{22}\,a_2\right) \quad\Longrightarrow\quad \frac{a_2}{a_1} = \frac{\Gamma_L\,S_{21}}{1-\Gamma_L\,S_{22}}$$

Therefore, $$\Gamma_1 = S_{11} + \frac{\Gamma_L\,S_{12}\,S_{21}}{1-\Gamma_L\,S_{22}}$$

Similarly, $$\Gamma_2 = S_{22} + \frac{\Gamma_S\,S_{12}\,S_{21}}{1-\Gamma_S\,S_{11}}$$

CONDITIONS OF STABILITY



$$\Rightarrow \quad \left|S_{11} + \frac{\Gamma_L\,S_{12}\,S_{21}}{1-\Gamma_L\,S_{22}}\right| < 1 \qquad \left|S_{22} + \frac{\Gamma_S\,S_{12}\,S_{22}}{1-\Gamma_S\,S_{11}}\right| < 1$$

AMPLIFIER DESIGN: LINEAR AMPLIFIERS

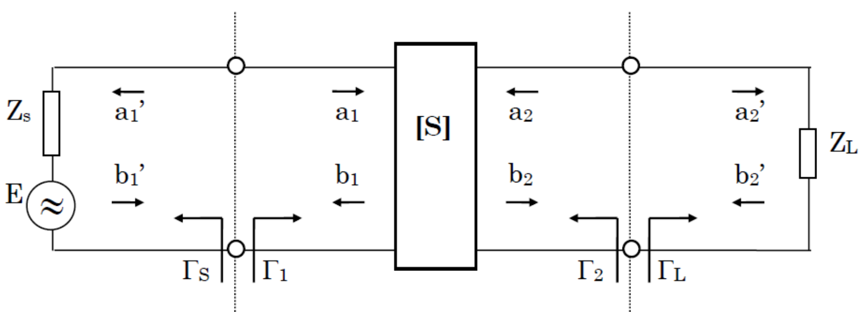
$$\Gamma_1 = \frac{b_1}{a_1} = \Gamma_S^* \qquad \Gamma_2 = \frac{b_2}{a_2} = \Gamma_L^* \qquad \left| S_{11} + \frac{\Gamma_L S_{12} S_{21}}{1-\Gamma_L S_{22}} \right| < 1 \qquad \left| S_{22} + \frac{\Gamma_S S_{12} S_{22}}{1-\Gamma_S S_{11}} \right| < 1$$

CONDITIONS OF MATCHING + CONDITIONS OF STABILITY

$$\Gamma_S = \Gamma_1^* = \left(S_{11} + \frac{\Gamma_L\, S_{12}\, S_{21}}{1-\Gamma_L\, S_{22}} \right)^* = S_{11}^* + \frac{\Gamma_L^*\, S_{12}^*\, S_{21}^*}{1-\Gamma_L^*\, S_{22}^*} \qquad \text{similarly} \qquad \Gamma_2 = \Gamma_L^* = S_{22} + \frac{\Gamma_S\, S_{12}\, S_{21}}{1-\Gamma_S\, S_{11}}$$

After substitution,

$$\Gamma_S = S_{11}^* + \frac{\left(S_{22} + \frac{\Gamma_S\, S_{12}\, S_{21}}{1-\Gamma_S\, S_{11}} \right) S_{12}^*\, S_{21}^*}{1 - \left(S_{22} + \frac{\Gamma_S\, S_{12}\, S_{21}}{1-\Gamma_S\, S_{11}} \right) S_{22}^*} = S_{11}^* + \frac{\left(S_{22} - S_{22}\,\Gamma_S\, S_{11} + \Gamma_S\, S_{12}\, S_{21} \right) S_{12}^*\, S_{21}^*}{1-\Gamma_S\, S_{11} - \left(S_{22} - S_{22}\,\Gamma_S\, S_{11} + \Gamma_S\, S_{12}\, S_{21} \right) S_{22}^*}$$



$$\Gamma_S = S_{11}^* + \frac{\left(S_{22} - \Delta\Gamma_S \right) S_{12}^*\, S_{21}^*}{1-\Gamma_S\, S_{11} - \left| S_{22} \right|^2 + S_{22}^*\, \Delta\Gamma_S}$$

$$\Delta = S_{11}\, S_{22} - S_{12}\, S_{21}$$

AMPLIFIER DESIGN: LINEAR AMPLIFIERS

$$\Gamma_1 = \frac{b_1}{a_1} = \Gamma_S^* \qquad \Gamma_2 = \frac{b_2}{a_2} = \Gamma_L^* \qquad \left| S_{11} + \frac{\Gamma_L S_{12} S_{21}}{1 - \Gamma_L S_{22}} \right| < 1 \qquad \left| S_{22} + \frac{\Gamma_S S_{12} S_{22}}{1 - \Gamma_S S_{11}} \right| < 1$$

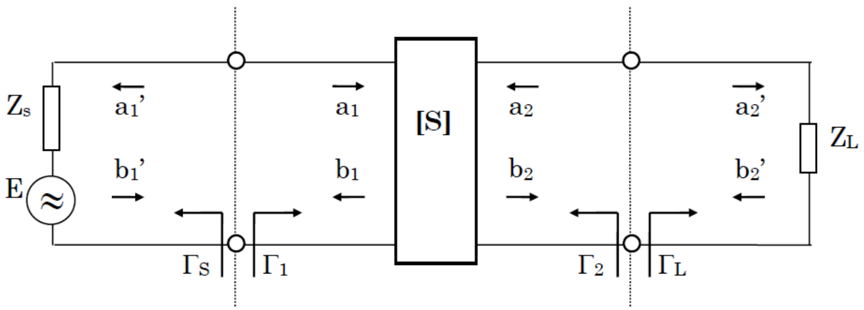
$$\Gamma_S = S_{11}^* + \frac{(S_{22} - \Delta \Gamma_S) S_{12}^* S_{21}^*}{1 - \Gamma_S S_{11} - |S_{22}|^2 + S_{22}^* \Delta \Gamma_S}$$

Let us expand this expression,

$$\Rightarrow \Gamma_S \left(1 - \Gamma_S S_{11} - |S_{22}|^2 + S_{22}^* \Delta \Gamma_S\right) = S_{11}^* \left(1 - \Gamma_S S_{11} - |S_{22}|^2 + S_{22}^* \Delta \Gamma_S\right) + (S_{22} - \Delta \Gamma_S) S_{12}^* S_{21}^*$$

$$\Rightarrow \Gamma_S \left(1 - |S_{22}|^2\right) + \Gamma_S^2 \left(S_{22}^* \Delta - S_{11}\right) = \Gamma_S \left(S_{11}^* S_{22}^* \Delta - |S_{11}|^2 - \Delta S_{12}^* S_{21}^*\right) + S_{11}^* - S_{11}^* |S_{22}|^2 + S_{22} S_{12}^* S_{21}^*$$

$$\Rightarrow \Gamma_S^2 \left(S_{22}^* \Delta - S_{11}\right) + \Gamma_S \left(1 - |S_{22}|^2 - S_{11}^* S_{22}^* \Delta + |S_{11}|^2 + \Delta S_{12}^* S_{21}^*\right) = S_{11}^* - S_{11}^* |S_{22}|^2 + S_{22} S_{12}^* S_{21}^*$$



with

$$S_{11}^* S_{22}^* \Delta - \Delta S_{12}^* S_{21}^* = |\Delta|^2$$

AMPLIFIER DESIGN: LINEAR AMPLIFIERS

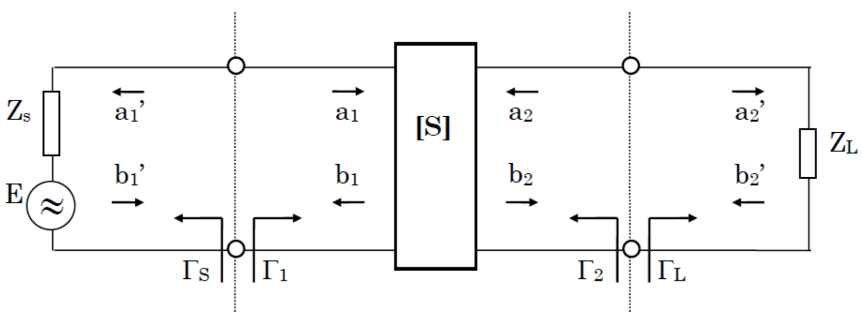
$$\Gamma_1 = \frac{b_1}{a_1} = \Gamma_S^* \qquad \Gamma_2 = \frac{b_2}{a_2} = \Gamma_L^* \qquad \left| S_{11} + \frac{\Gamma_L S_{12} S_{21}}{1 - \Gamma_L S_{22}} \right| < 1 \qquad \left| S_{22} + \frac{\Gamma_S S_{12} S_{22}}{1 - \Gamma_S S_{11}} \right| < 1$$

$$\Gamma_S^2 \left(S_{22}^* \Delta - S_{11} \right) + \Gamma_S \left(1 - |S_{22}|^2 - S_{11}^* S_{22}^* \Delta + |S_{11}|^2 + \Delta S_{12}^* S_{21}^* \right) = S_{11}^* - S_{11}^* |S_{22}|^2 + S_{22} S_{12}^* S_{21}^*$$

$$\Rightarrow \quad -\Gamma_S^2 \left(S_{22}^* \Delta - S_{11} \right) - \Gamma_S \left(1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2 \right) + \left(S_{11}^* - S_{11}^* S_{22} S_{22}^* + S_{22} S_{12}^* S_{21}^* \right) = 0$$

$$\Rightarrow \quad \Gamma_S^2 \left(S_{11} - S_{22}^* \Delta \right) - \Gamma_S \left(1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2 \right) + \left(S_{11}^* - S_{22} \Delta^* \right) = 0$$

Let $\quad B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2 \qquad C_1 = S_{11} - \Delta S_{22}^*$



$$\Rightarrow \quad \Gamma_S^2 C_1 - \Gamma_S B_1 + C_1^* = 0$$

$$a x^2 + b x + c = 0$$

AMPLIFIER DESIGN: LINEAR AMPLIFIERS

$$\Gamma_1 = \frac{b_1}{a_1} = \Gamma_S^* \qquad \Gamma_2 = \frac{b_2}{a_2} = \Gamma_L^* \qquad \left| S_{11} + \frac{\Gamma_L S_{12} S_{21}}{1 - \Gamma_L S_{22}} \right| < 1 \qquad \left| S_{22} + \frac{\Gamma_S S_{12} S_{22}}{1 - \Gamma_S S_{11}} \right| < 1$$

$$\Gamma_S^2 \, C_1 - \Gamma_S \;\, B_1 + C_1^* = 0$$

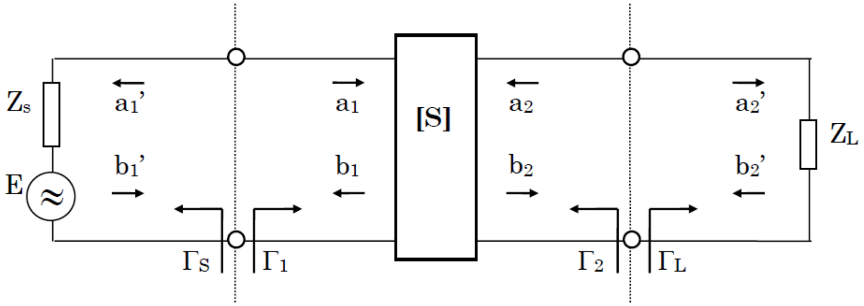
Two solutions:

$$\Gamma_S = \frac{B_1 \pm \sqrt{B_1^2 - 4\left|C_1\right|^2}}{2\,\mathrm{C}_1}$$

Similarly, let $\Gamma_L^2 \, C_2 - \Gamma_L \;\, B_2 + C_2^* = 0$

$$B_2 = 1 + \left|S_{22}\right|^2 - \left|S_{11}\right|^2 - \left|\Delta\right|^2$$

$$C_2 = S_{22} - \Delta\, S_{11}^*$$



Two solutions : $\Gamma_L = \dfrac{B_2 \pm \sqrt{B_2^2 - 4\left|C_2\right|^2}}{2\,\mathrm{C}_2}$

AMPLIFIER DESIGN: LINEAR AMPLIFIERS

$$\Gamma_1 = \frac{b_1}{a_1} = \Gamma_S^* \qquad \Gamma_2 = \frac{b_2}{a_2} = \Gamma_L^*$$

$$\left| S_{11} + \frac{\Gamma_L S_{12} S_{21}}{1 - \Gamma_L S_{22}} \right| < 1 \qquad \left| S_{22} + \frac{\Gamma_S S_{12} S_{22}}{1 - \Gamma_S S_{11}} \right| < 1$$

$$\Gamma_S = \frac{B_1 \pm \sqrt{B_1^2 - 4|C_1|^2}}{2\,C_1} \qquad \Gamma_L = \frac{B_2 \pm \sqrt{B_2^2 - 4|C_2|^2}}{2\,C_2}$$

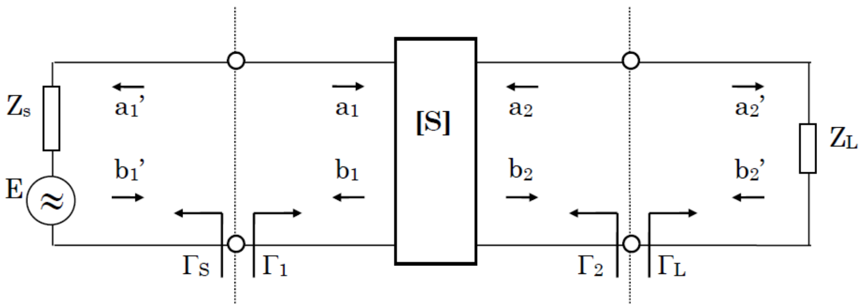
$$B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2 \qquad B_2 = 1 + |S_{22}|^2 - |S_{11}|^2 - |\Delta|^2$$

$$C_1 = S_{11} - \Delta\, S_{22}^* \qquad C_2 = S_{22} - \Delta\, S_{11}^*$$

Note that a permutation of the subscripts 1 and 2 does not affect the above expressions.

Let us define a stability factor called the Rollett's factor

$$K = \frac{1 + |\Delta|^2 - |S_{11}|^2 - |S_{22}|^2}{2|S_{12}||S_{21}|}$$



STABILITY FACTOR

http://www.calculatoredge.com/electronics/rf%20amplifier.htm

AMPLIFIER DESIGN: LINEAR AMPLIFIERS

$$\Gamma_1 = \frac{b_1}{a_1} = \Gamma_S^* \qquad \Gamma_2 = \frac{b_2}{a_2} = \Gamma_L^*$$

$$\left| S_{11} + \frac{\Gamma_L S_{12} S_{21}}{1-\Gamma_L S_{22}} \right| < 1 \qquad \left| S_{22} + \frac{\Gamma_S S_{12} S_{22}}{1-\Gamma_S S_{11}} \right| < 1$$

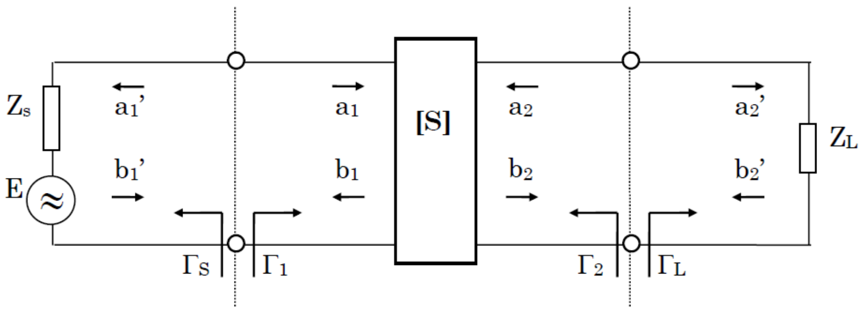
$$\Gamma_S = \frac{B_1 \pm \sqrt{B_1^2 - 4|C_1|^2}}{2\,C_1} \qquad \Gamma_L = \frac{B_2 \pm \sqrt{B_2^2 - 4|C_2|^2}}{2\,C_2}$$

$$B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2 \qquad B_2 = 1 + |S_{22}|^2 - |S_{11}|^2 - |\Delta|^2$$

$$C_1 = S_{11} - \Delta\, S_{22}^* \qquad C_2 = S_{22} - \Delta\, S_{11}^*$$

$$K = \frac{1 + |\Delta|^2 - |S_{11}|^2 - |S_{22}|^2}{2|S_{12}||S_{21}|}$$

$$\Rightarrow \quad B_1^2 - 4|C_1|^2 = B_2^2 - 4|C_2|^2 = 4\left(K^2 - 1\right)|S_{12}\, S_{21}|^2$$



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$$\Gamma_1 = \frac{b_1}{a_1} = \Gamma_S^* \qquad \Gamma_2 = \frac{b_2}{a_2} = \Gamma_L^* \qquad \left| S_{11} + \frac{\Gamma_L S_{12} S_{21}}{1-\Gamma_L S_{22}} \right| < 1 \qquad \left| S_{22} + \frac{\Gamma_S S_{12} S_{22}}{1-\Gamma_S S_{11}} \right| < 1$$

$$\Gamma_S = \frac{B_1 \pm \sqrt{B_1^2 - 4|C_1|^2}}{2\,C_1} \qquad \Gamma_L = \frac{B_2 \pm \sqrt{B_2^2 - 4|C_2|^2}}{2\,C_2}$$

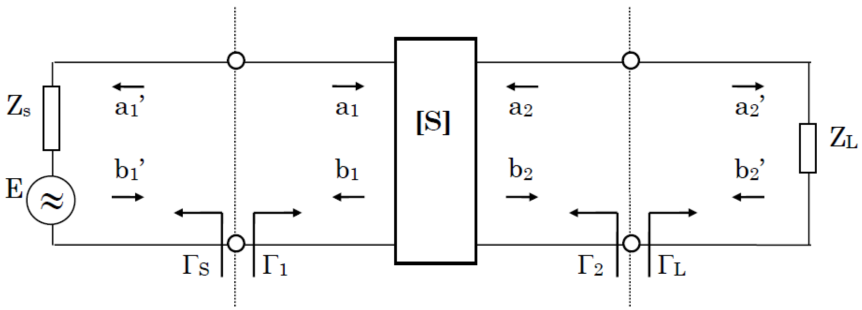
$$B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2 \qquad B_2 = 1 + |S_{22}|^2 - |S_{11}|^2 - |\Delta|^2$$

$$C_1 = S_{11} - \Delta\, S_{22}^* \qquad C_2 = S_{22} - \Delta\, S_{11}^*$$

$$B_1^2 - 4|C_1|^2 = B_2^2 - 4|C_2|^2 = 4\left(K^2 - 1\right)|S_{12}\ S_{21}|^2$$

What is the sign of the radicand?

The sign of the radicand depends on the sign of { $K^2 - 1$ } !!!



$$K^2 - 1 = \left(\frac{1 + |\Delta|^2 - |S_{11}|^2 - |S_{22}|^2}{2|S_{12}||S_{21}|} \right)^2 - 1$$

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$$\Gamma_S = \frac{B_1 \pm \sqrt{B_1^2 - 4|C_1|^2}}{2\,C_1} \qquad \Gamma_L = \frac{B_2 \pm \sqrt{B_2^2 - 4|C_2|^2}}{2\,C_2}$$

$$B_1^2 - 4|C_1|^2 = B_2^2 - 4|C_2|^2 = 4\left(K^2 - 1\right)|S_{12}\,S_{21}|^2$$

a) ***K = 1*** it is not a practical case since $|\Gamma_S| = |\Gamma_L| = 1$

b) ***K > 1*** expanding the two solutions of Γ_S and Γ_L gives

$$\Gamma_S' = \frac{B_1 - \sqrt{B_1^2 - 4|C_1|^2}}{2\,C_1}; \qquad \Gamma_S'' = \frac{B_1 + \sqrt{B_1^2 - 4|C_1|^2}}{2\,C_1}$$

$$\Gamma_L' = \frac{B_2 - \sqrt{B_2^2 - 4|C_2|^2}}{2\,C_2}; \qquad \Gamma_L'' = \frac{B_2 + \sqrt{B_2^2 - 4|C_2|^2}}{2\,C_2}$$

Note that

$$|\Gamma_S'||\Gamma_S''| = 1 \qquad |\Gamma_L'||\Gamma_L''| = 1$$

To explicit which solution has to be retained, we have to consider Δ:

- $|\Delta| < 1$ The pair of solutions is $\{\, \Gamma_S',\, \Gamma_L' \,\}$.

  Unconditional stability and maximum gain

$$G_{T\,\max} = \left|\frac{S_{21}}{S_{12}}\right|\left(K - \sqrt{K^2 - 1}\right)$$

- $|\Delta| > 1$ The pair of solutions is $\{\, \Gamma_S'',\, \Gamma_L'' \,\}$.

  Conditional stability and minimum gain

$$G_{T\,\min} = \left|\frac{S_{21}}{S_{12}}\right|\left(K + \sqrt{K^2 + 1}\right)$$

Note :

The **maximum available gain** ($G_{T\,MAG}$ or $G_{T\,MAX}$) of a device is only defined where K > 1.

Algebraically, this is because the term under the square-root becomes negative for values of K less than 1. Another way to look at it is that maximum available gain is infinite. Infinite gain means oscillator.

The **maximum stable gain (MSG)** of a device is defined when maximum available gain is undefined (K<1).

It is merely the ratio of $|S_{21}| / |S_{12}|$.

Do not try to get more than this amount of gain from a conditionally stable device !

$$G_{T\max} = \left|\frac{S_{21}}{S_{12}}\right| \left(K - \sqrt{K^2 - 1}\right)$$

https://www.microwaves101.com/encyclopedias/stability-factor#kfactor

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c) **$K \leq -1$** For this value of K, the two possible solutions are { Γ_S', Γ_L'' } or { Γ_S'' , Γ_L' }.

However, these solutions **are not useful** for amplifiers for two raisons:

- First, it is not possible to match simultaneously the input and the output.
- Second, the amplifier is unstable because we associate a solution with a magnitude > 1 with another of magnitude < 1 : **naturally instable** !

d) **$-1 \leq K \leq 1$** Simultaneous matching is impossible.

The amplifier is **conditionally stable** (or potentially unstable), because only **some values** of Γ_S and Γ_L can be used to **avoid oscillations**.

In this case, we need to **determine** the **allowed regions** in which we can select the **load impedances** that **assure stability**.

This is the more frequent case !!

AMPLIFIER DESIGN: LINEAR AMPLIFIERS

⇒ Stability regions ?

Reported to the reflection coefficient space (Smith chart), these regions are equivalent to circles, and thus are called "stability circles".

The locus of Γ_S given the critical value { $\Gamma_L = 1$ } is a circle in the complex plane (center Ω_1, radius R_1).

$$\Omega_1 = \frac{\left(S_{11} - \Delta S_{22}^*\right)^*}{\left|S_{11}\right|^2 - \left|\Delta\right|^2} \qquad R_1 = \frac{\left|S_{21} S_{12}\right|}{\left|\left|S_{11}\right|^2 - \left|\Delta\right|^2\right|}$$

The locus of Γ_L given the critical value { $\Gamma_S = 1$ } is a circle in the complex plane (center Ω_2, radius R_2).

$$\Omega_2 = \frac{\left(S_{22} - \Delta S_{11}^*\right)^*}{\left|S_{22}\right|^2 - \left|\Delta\right|^2} \qquad R_2 = \frac{\left|S_{12} S_{21}\right|}{\left|\left|S_{22}\right|^2 - \left|\Delta\right|^2\right|}$$

Circles ?

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In fact, let $$|\Gamma_S| = 1 = \left| S_{11} + \frac{\Gamma_L\, S_{12}\, S_{21}}{1 - \Gamma_L\, S_{22}} \right|$$

Given the squares $$|S_{11} - \Gamma_L\, \Delta|^2 = (S_{11} - \Gamma_L\, \Delta)(S_{11} - \Gamma_L\, \Delta)^* = (1 - \Gamma_L\, S_{22})(1 - \Gamma_L\, S_{22})^*$$

We get $$|\Gamma_L|^2 - \frac{2}{|S_{22}|^2 - |\Delta|^2}.\mathrm{Re}\left[\left(S_{22} - \Delta S_{11}^*\right)\Gamma_L\right] = \frac{|S_{11}|^2 - 1}{|S_{22}|^2 - |\Delta|^2}$$

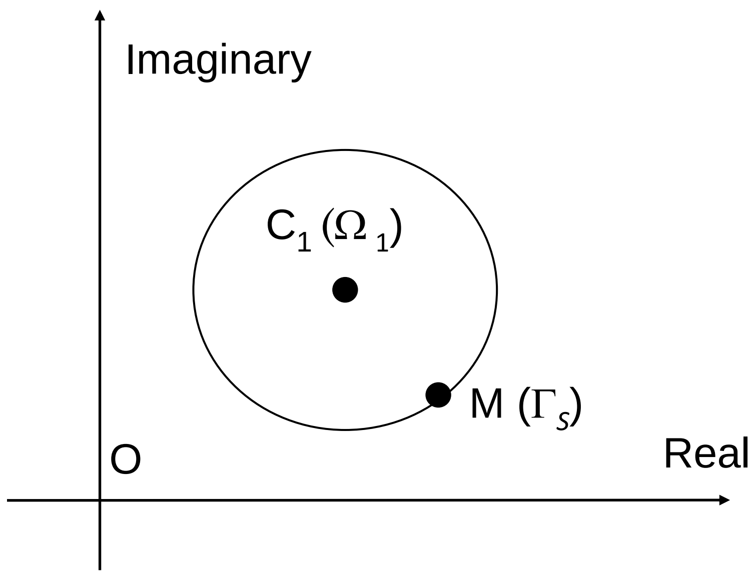
Similarly, $$|\Gamma_S|^2 - \frac{2}{|S_{11}|^2 - |\Delta|^2}.\mathrm{Re}\left[\left(S_{11} - \Delta S_{22}^*\right)\Gamma_S\right] = \frac{|S_{22}|^2 - 1}{|S_{11}|^2 - |\Delta|^2}$$

Equation of a circle in the complex plane:

$$\overline{OM} = \overline{OC_1} + \overline{C_1M}$$

In magnitude : $$\left|\overline{C_1M}\right|^2 = \left|\overline{OM} - \overline{OC_1}\right|^2$$

$$\Rightarrow \qquad R_1^2 = \left(\Gamma_S - \Omega_1\right)\left(\Gamma_S - \Omega_1\right)^*$$



$$\left|\Gamma_S\right|^2 - 2\,\mathrm{Re}\left(\Gamma_S\,\Omega_1^*\right) = R_1^2 - \left|\Omega_1\right|^2$$

Equation of a circle in the complex plane: $$\left|\Gamma_S\right|^2 - 2\,\mathrm{Re}\left(\Gamma_S\,\Omega_1^*\right) = R_1^2 - \left|\Omega_1\right|^2$$

We have :
$$\left|\Gamma_S\right|^2 - \frac{2}{\left|S_{11}\right|^2 - \left|\Delta\right|^2}.\mathrm{Re}\left[\left(S_{11} - \Delta S_{22}^*\right)\Gamma_S\right] = \frac{\left|S_{22}\right|^2 - 1}{\left|S_{11}\right|^2 - \left|\Delta\right|^2}$$

$$\left|\Gamma_L\right|^2 - \frac{2}{\left|S_{22}\right|^2 - \left|\Delta\right|^2}.\mathrm{Re}\left[\left(S_{22} - \Delta S_{11}^*\right)\Gamma_L\right] = \frac{\left|S_{11}\right|^2 - 1}{\left|S_{22}\right|^2 - \left|\Delta\right|^2}$$

$$\Omega_1 = \frac{\left(S_{11} - \Delta\,S_{22}^*\right)^*}{\left|S_{11}\right|^2 - \left|\Delta\right|^2} \qquad \mathrm{R}_1 = \frac{\left|S_{21}\,S_{12}\right|}{\left|\left|S_{11}\right|^2 - \left|\Delta\right|^2\right|}$$

After identification:

$$\Omega_2 = \frac{\left(S_{22} - \Delta\,S_{11}^*\right)^*}{\left|S_{22}\right|^2 - \left|\Delta\right|^2} \qquad \mathrm{R}_2 = \frac{\left|S_{12}\,S_{21}\right|}{\left|\left|S_{22}\right|^2 - \left|\Delta\right|^2\right|}$$

Reported into the reflection coefficient space (Smith chart), these regions are equivalent to circles, and thus are called "stability circles".

To summarise, when the amplifier is potentially unstable, there may be values of Γ_S and Γ_L for which the real parts of Z_S and Z_L are positive.

The set of values of Γ_S and Γ_L (i.e., regions in the Smith chart) can be determined using the following:

I - First, the regions were values of Γ_S and Γ_L produce respectively { $|\Gamma_1| = 1$ } and { $|\Gamma_2| = 1$ } are determined (boundary cases). Set:

$$|\Gamma_S| = 1 \Rightarrow |\Gamma_1| = \left| S_{11} + \frac{\Gamma_L \; S_{12} \; S_{21}}{1 - \Gamma_L \; S_{22}} \right| = 1 \qquad |\Gamma_L| = 1 \Rightarrow |\Gamma_2| = \left| S_{22} + \frac{\Gamma_S \; S_{21} \; S_{12}}{1 - \Gamma_S \; S_{11}} \right| = 1$$

II - Solve for the values of Γ_S and Γ_L showing that the solutions for Γ_S and Γ_L lie on circles whose equations are given by

$$\left| \Gamma_S - \frac{\left(S_{11} - \Delta \; S_{22}^*\right)^*}{|S_{11}|^2 - |\Delta|^2} \right| = \left| \frac{S_{21} \; S_{12}}{|S_{11}|^2 - |\Delta|^2} \right| \qquad \left| \Gamma_L - \frac{\left(S_{22} - \Delta \; S_{11}^*\right)^*}{|S_{22}|^2 - |\Delta|^2} \right| = \left| \frac{S_{12} \; S_{21}}{|S_{22}|^2 - |\Delta|^2} \right|$$

The derivations (see the course notes) show two families of circles:

Γ_L values for $|\Gamma_1| = 1$ output stability circle: Centre Ω_2 and radius R_2

Γ_S values for $|\Gamma_2| = 1$ output stability circle: Centre Ω_1 and radius R_1

In the Γ_L plane, on **one side of the circle boundary** we will have $|\Gamma_1| > 1$ and **on the other side** $|\Gamma_1| < 1$.

In the Γ_S plane, on **one side of the circle boundary** we will have $|\Gamma_2| > 1$ and **on the other side** $|\Gamma_2| < 1$.

Next we need to determine which area in the Smith chart represents the stable region.

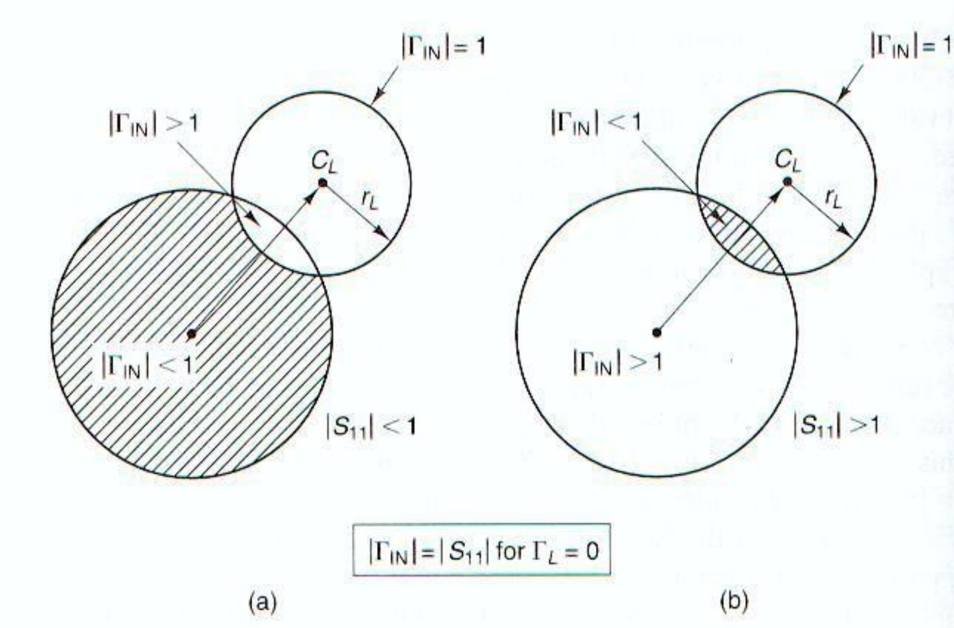
In other words, the regions where

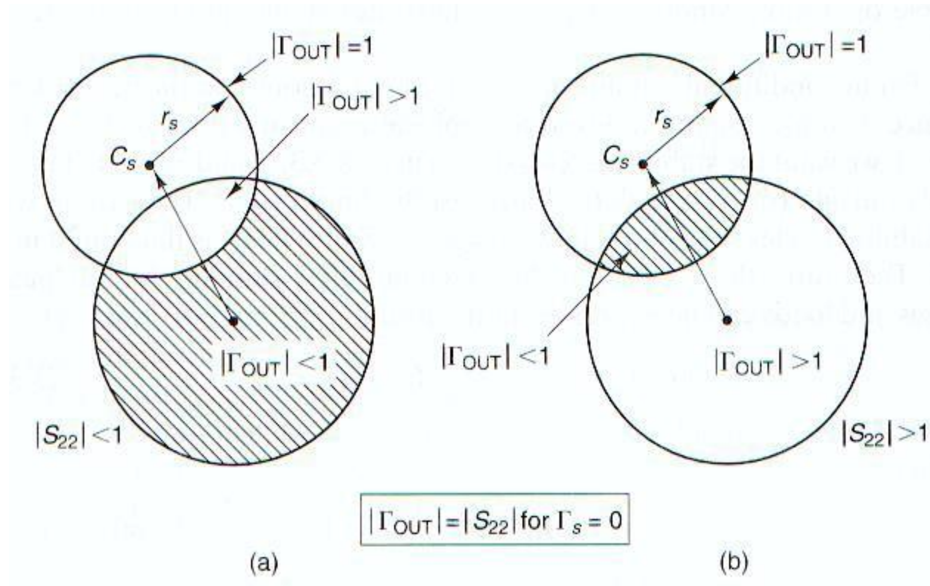
- values of Γ_L (where $|\Gamma_L| < 1$) produce $|\Gamma_1| < 1$ and where
- values of Γ_S (where $|\Gamma_S| < 1$) produce $|\Gamma_2| < 1$.

To this end, we observe that if $Z_L = Z_0$, then $\Gamma_L = 0$ and $\Gamma_1 = S_{11}$. If the magnitude of S_{11} is less than 1, then $|\Gamma_1| < 1$ when $\Gamma_L = 0$.

That is, the centre of the Smith chart represents, **by default**, a stable operating region.

On the other hand, if the magnitude of S_{11} is greater than 1, then the centre of the Smith chart is in an unstable operating region for the input.





Stable and unstable regions in the Γ_L plane

Stable and unstable regions in the Γ_S plane

However, these are “regions” !

To find the exact appropriate impedance values for a transistor amplifier, the designer needs to specify the values of the gain and the noise factor he is targeting for the designed amplifier.

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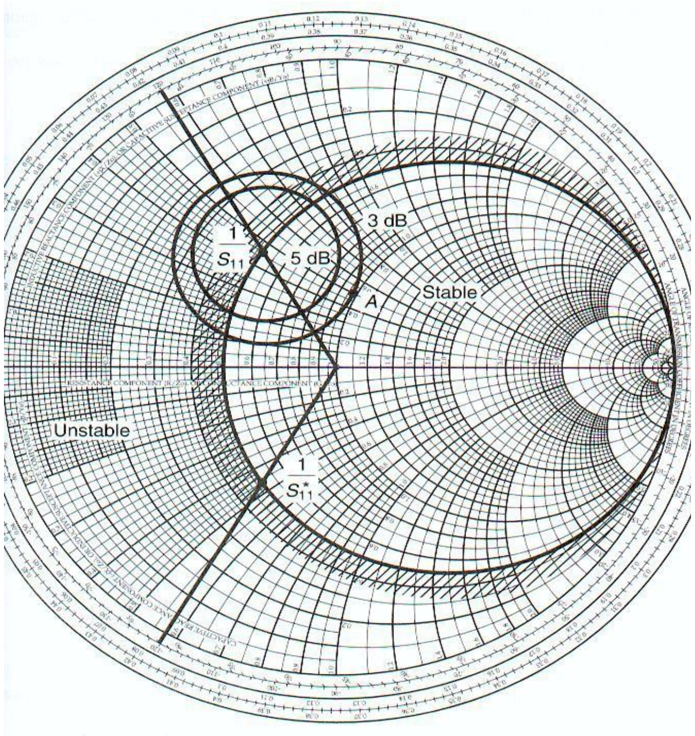
Power Gain

When the input is matched, the power gain G_p becomes

$$G_p = \frac{|S_{21}|^2 \left(1-|\Gamma_L|^2\right)}{1-|S_{11}|^2 + |\Gamma_L|^2.\left(|S_{22}|^2 - |\Delta|^2\right) - 2\,\mathrm{Re}\left(C_2\,\Gamma_L\right)}$$

Let us introduce the normalized gain g_p

$$g_p = \frac{G_p}{|S_{21}|^2} = \frac{\left(1-|\Gamma_L|^2\right)}{1-|S_{11}|^2 + |\Gamma_L|^2.\left[\left(|S_{22}|^2 - |\Delta|^2\right)\right] - 2\,\mathrm{Re}\left(C_2\Gamma_L\right)}$$



Constant-gain circles for a given transistor, including stable/unstable regions of the load reflection coefficient.

We can show that the Γ_L that give a constant gain g_p are located on a "constant-gain circle" of radius R_p centered at Ω_p

$$\Omega_p = \frac{g_p\left(S_{22}^* - \Delta * S_{11}\right)}{1+g_p\left(|S_{22}|^2 - |\Delta|^2\right)}$$

$$R_p = \frac{\sqrt{1 - 2\,K\,.|S_{12}.S_{21}|.g_p + |S_{12}.S_{21}|^2.g_p^2}}{\left|1+g_p\left(|S_{22}|^2 - |\Delta|^2\right)\right|}$$

AMPLIFIER DESIGN: LINEAR AMPLIFIERS

Power Gain

Note 1: **The gain is related to the output impedance**, i.e., the load reflection coefficient Γ_L. Choosing a specific value for the gain implies choosing a specific value for the load.

Note 2: **Any value for the load has to be inside the stable region** for potentially unstable cases. For stable transistors, any load value can be selected (with the restriction: Re(Z) > 0).

Once Γ_L is selected, maximum output power is obtained with a conjugate match at the input namely, with $\Gamma_S = \Gamma_1^*$

$$\Gamma_S = \left(\frac{S_{11} - \Delta \Gamma_L}{1 - \Gamma_L S_{22}} \right)^*$$

Verify that the obtained Γ_S value will satisfy the stability criterion

AMPLIFIER DESIGN: LINEAR AMPLIFIERS

Noise Figure (NF)

A designer has to know that if the **gain is related to the load** (output reflection coefficient Γ_L),

the noise figure of an amplifier is **related to the source impedance** (i.e., Γ_S).

So, selecting a given noise factor will imply selecting a specific source impedance.

The noise figure of a two-port network is given by

$$F = F_{\min} + 4\frac{R_n}{Z_o}\frac{\left|\Gamma_S - \Gamma_{opt}\right|^2}{\left|1+\Gamma_{opt}\right|^2\left(1-\left|\Gamma_S\right|^2\right)} \Rightarrow F = F_{\min} + \frac{r_n}{g_S}\left|y_S - y_{opt}\right|^2$$

$r_n = R_n/Z_o$: equivalent normalized noise resistance of the two-port (e.g., transistor),

$y_S = g_S + \mathrm{j}\, b_S$: normalized source admittance,

$$y_S = \frac{1-\Gamma_S}{1+\Gamma_S}$$

$Y_{opt} = g_{opt} + \mathrm{j}\, b_{opt}$: normalized source admittance that results in the minimum (or optimum) noise figure called F_{min}

$$y_{opt} = \frac{1-\Gamma_{opt}}{1+\Gamma_{opt}}$$

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Noise Figure (NF)

$$F = F_{\min} + \frac{r_n}{g_S}\left|y_S - y_{opt}\right|^2 \qquad y_S = \frac{1-\Gamma_S}{1+\Gamma_S} \qquad y_{opt} = \frac{1-\Gamma_{opt}}{1+\Gamma_{opt}}$$

F_{min}, r_n and Γ_{opt} are called the noise parameters and are given by the manufacturer of the transistor or have to be determined experimentally.

The value of F_{min} occurs when $\Gamma_{opt} = \Gamma_S$.

The noise resistance r_n can be measured by reading the noise figure $F = F_S$ when $\Gamma_S = 0$, leading to

$$F_S = F_{\min} + 4r_n \frac{\left|\Gamma_{opt}\right|^2}{\left|1+\Gamma_{opt}\right|^2}$$

$$r_n = \left(F_S - F_{\min}\right) + \frac{\left|1+\Gamma_{opt}\right|^2}{4\left|\Gamma_{opt}\right|^2}$$

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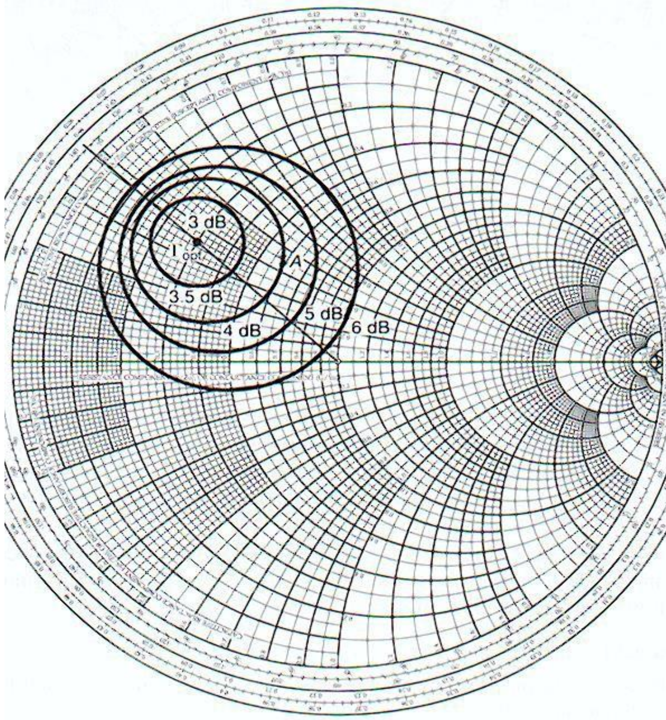
Noise Figure (NF)

$$F_S = F_{\min} + 4r_n \frac{\left|\Gamma_{opt}\right|^2}{\left|1+\Gamma_{opt}\right|^2}$$

We can rearrange the above equation as follows

$$N = \frac{\left|\Gamma_S - \Gamma_{opt}\right|^2}{1-\left|\Gamma_S\right|^2} = \frac{F - F_{\min}}{4}\frac{Z_o}{R_n}\left|1+\Gamma_{\min}\right|^2$$

to show that it can be recognized as the equation of a circle in the Γ_S plane.



Constant-NF circles for a given transistor, including stable/unstable regions of the source reflection coefficient.

So, for a given N, the center C_n and radius R_n of such circles called "constant noise figure circles", are

$$R_n = \frac{\sqrt{N^2 + N\left(1-\left|\Gamma_{\min}\right|^2\right)}}{1+N} \qquad C_n = \frac{\Gamma_{\min}}{1+N}$$

As for the gain, the value of Γ_S allows to select the corresponding value of Γ_L.

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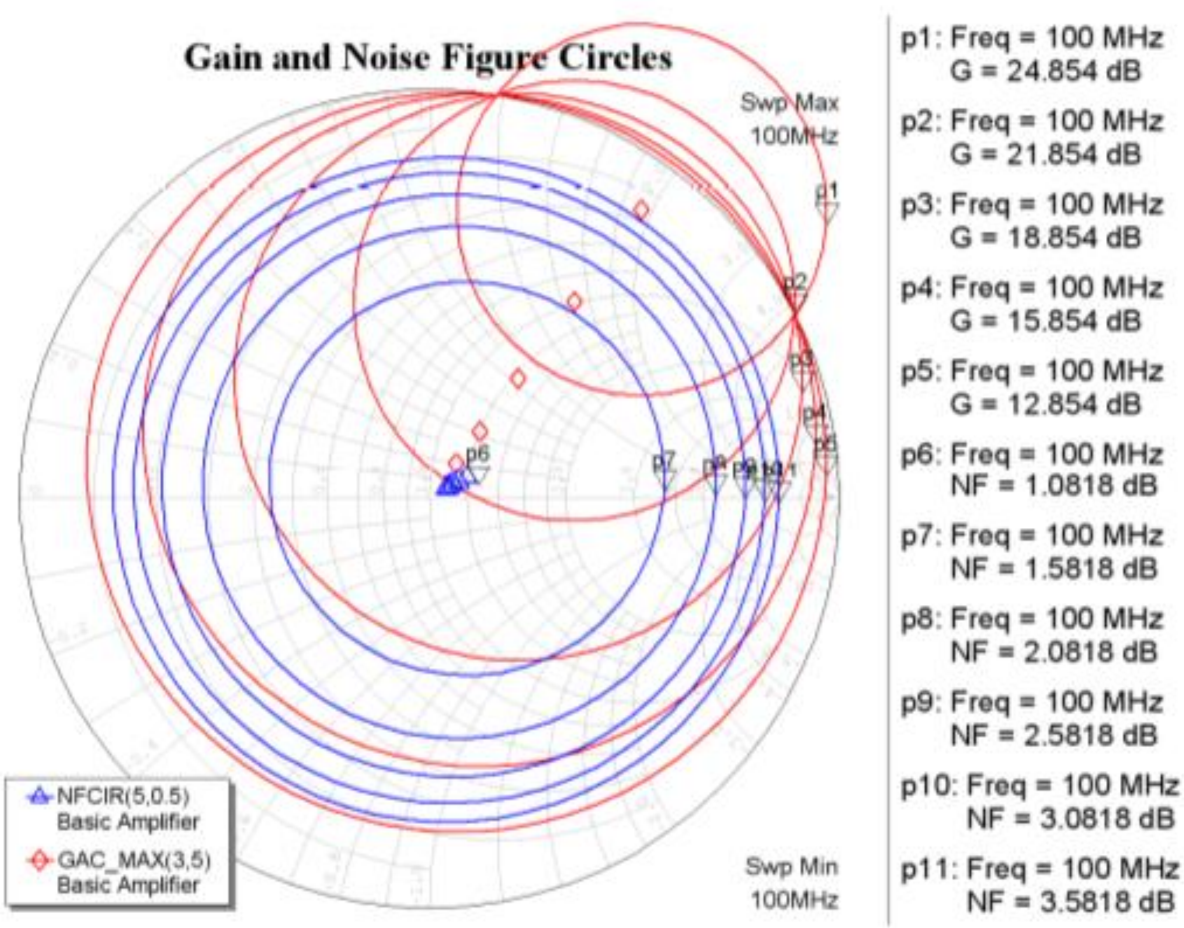
Gain and Noise Figure

Note:

Constant-gain circles:	when the radius increases, gain decreases.
Constant-noise figure circles:	when the radius increases, noise figure increases.



http://mwl.diet.uniroma1.it/people/pisa/SISTEMI_RF/MATERIALE%20INTEGRATIVO/Kikkert_RF_Electronics_Course/11-RF_Electronics_Kikkert_Ch8_AmplifierStabilityNoise.pdf

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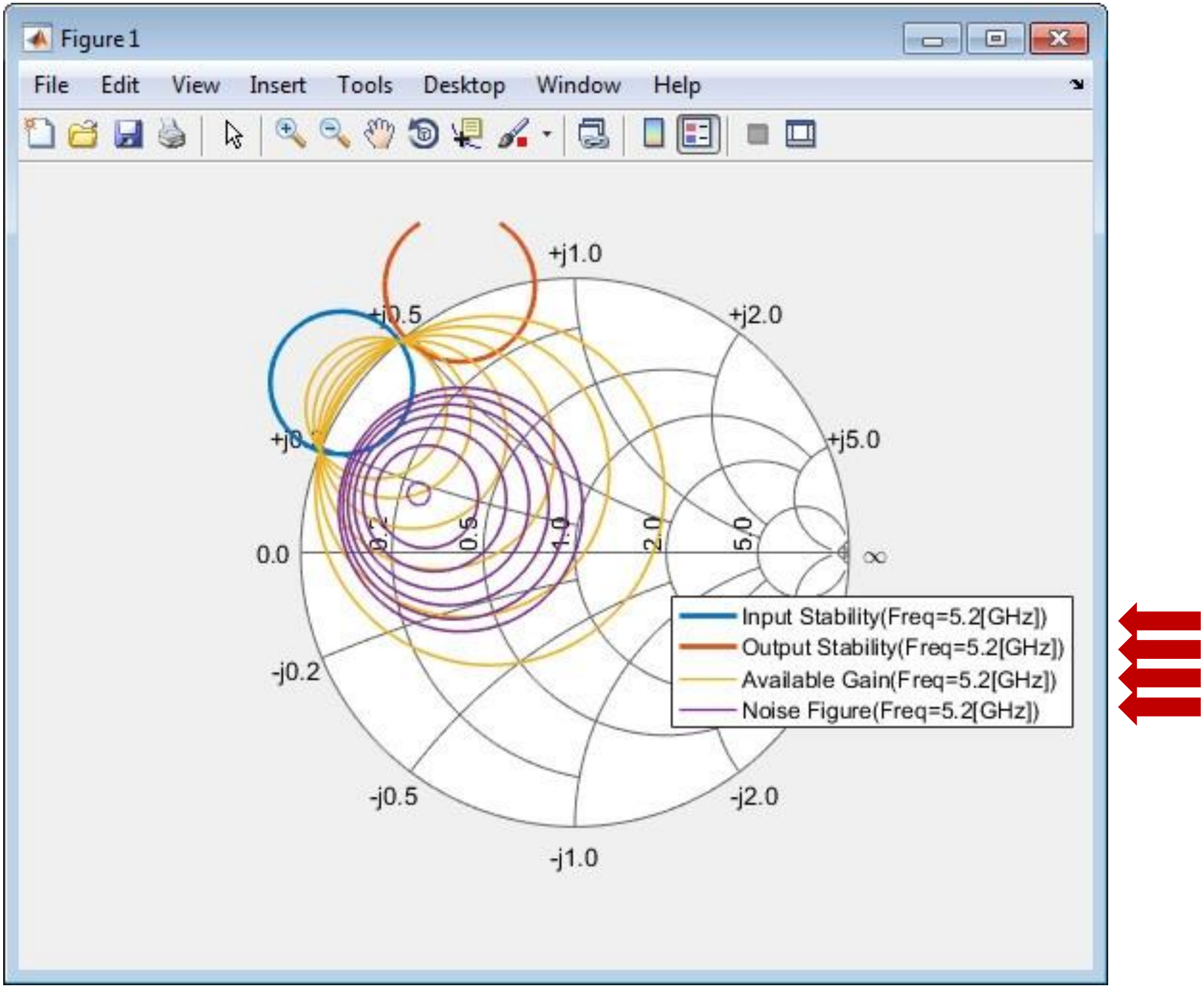
Summary



https://www.mathworks.com/help/rf/ug/designing-matching-networks-part-1-networks-with-an-lna-and-lumped-elements.html

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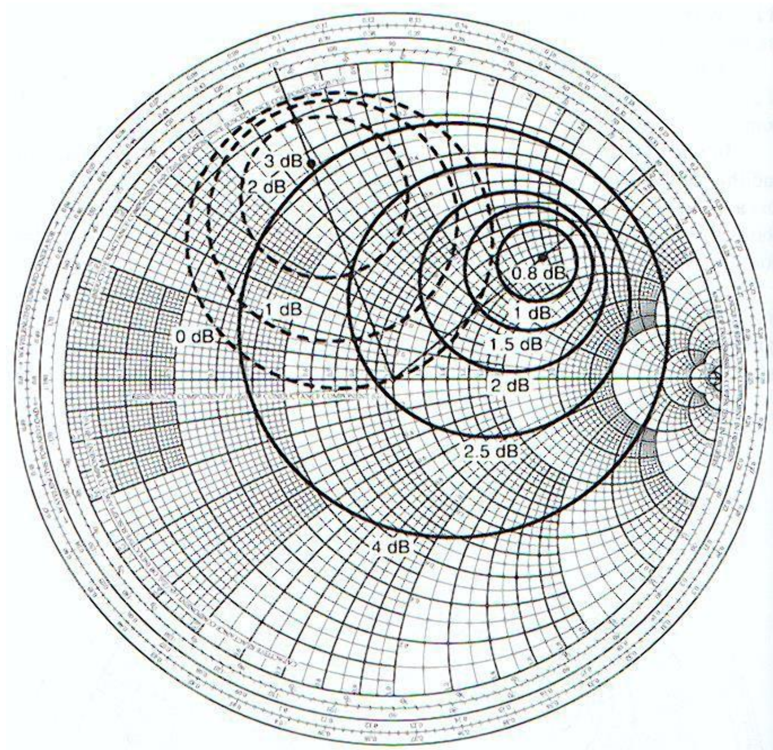
Trade-off Gain/Noise Figure



Constant-gain circles (dashed curves) and constant -noise figure circles (solid curves) for a given transistor.

Three options:

- For a **given** gain and NF, the constant-gain circle and the constant-NF circle has **two** intersection points. Hence, we have two possible values for Γ_S. The designer has to assure that the selected point belongs to the input stability region and that the corresponding impedance value is realistic (in terms of technology limitations, substrate performance, matching network, etc.). Deduce Γ_L (same restrictions).

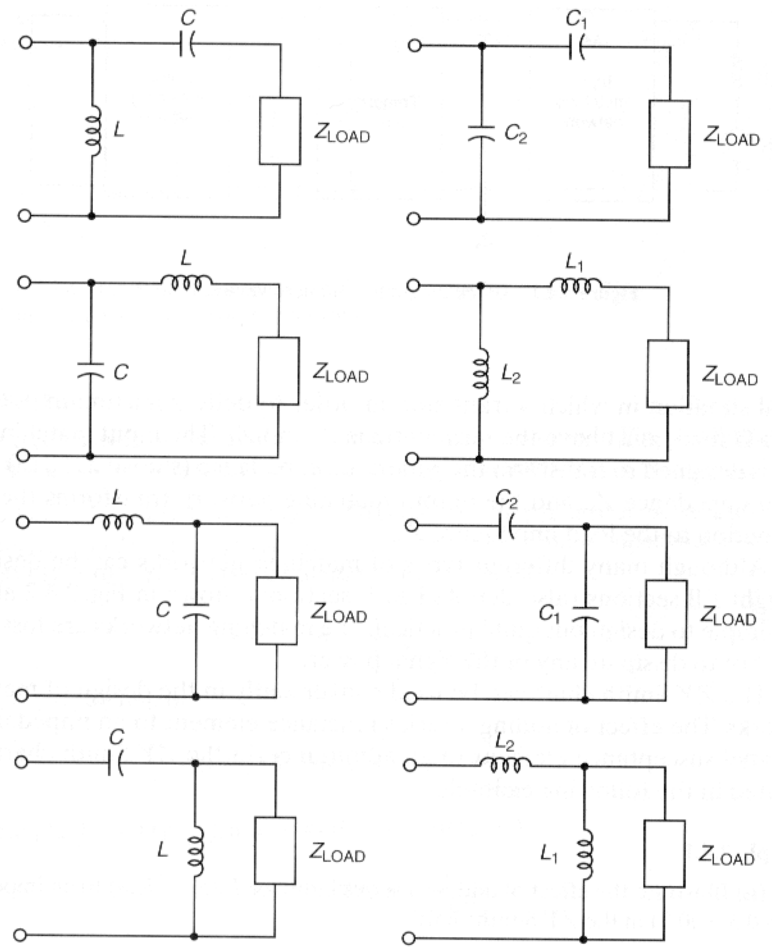
- **One** intersection point, thus one possible value for Γ_S.

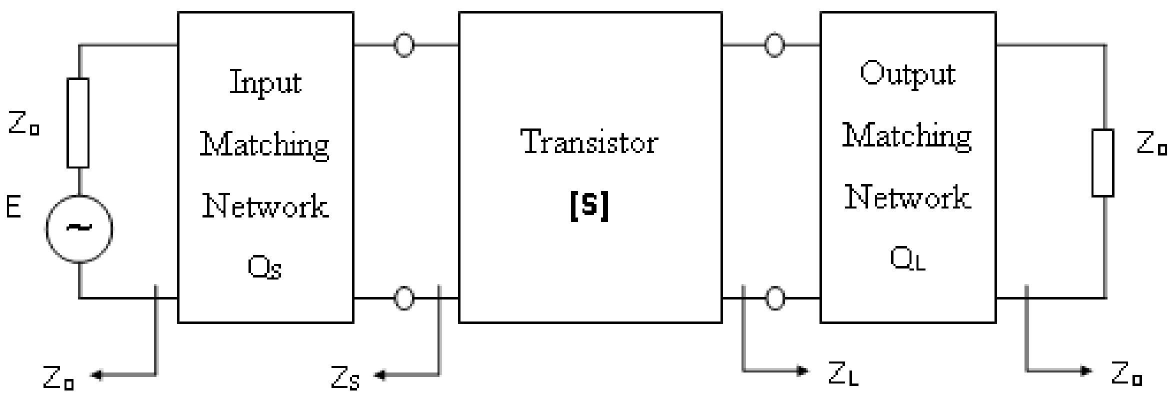
- **No** intersection. Impossible to design an amplifier with such gain and noise figure specifications.

TYPICAL CONFIGURATIONS OF MATCHING NETWORKS

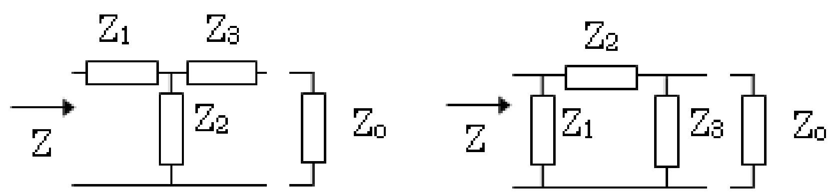
The need for matching networks arises because amplifiers, in order to deliver maximum power to a load or to perform in a certain desired way, must be properly terminated at both the input and the output ports: this is the role of the matching networks.

For **narrow band amplifiers**, the analysis of matching circuits at microwave frequencies can be cumbersome in analytical form. This figure shows some basic narrow band matching networks (usually LC networks).



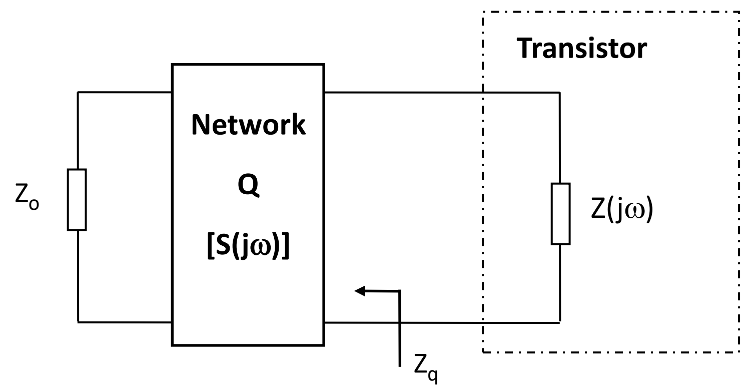


For amplifiers with a **relatively narrow band**, designers can use T or Π matching networks as shown here.



For **wide band amplifiers**, a designer needs an input/output matching over a large bandwidth. Accordingly, the impedance is not a constant, but varies with frequency and can be written as a transfer function { $Z(j\omega) = Z(s)$ }.

We have then to extract the number of poles and zeros to determine the matching cell.



Let **[S(jω)]** be the scattering matrix of two-port Q.

The power gain of the network is equal to

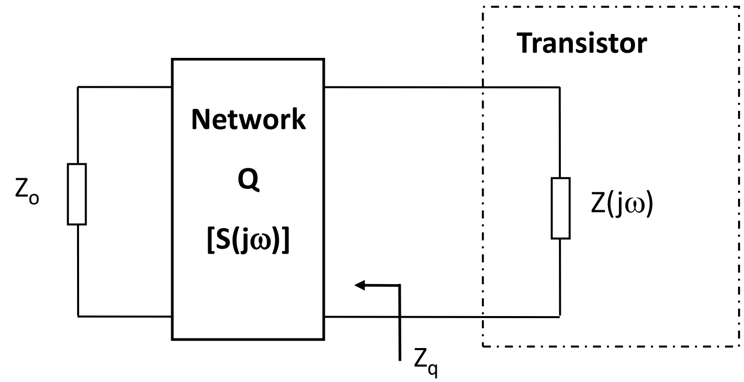
$$G(\omega^2) = |S_{21}(j\omega)|^2 = 1 - |S_{22}(j\omega)|^2 = 1 - \left|\frac{Z_q(s) - Z(-s)}{Z_q(s) + Z(-s)}\right|^2$$

where $Z_q(s)$ is the impedance seen by the transistor. The network is now equivalent to a wideband network.

Therefore, the procedure to obtain the two-port network configuration can be summarized in the following steps:

* Verify that the function $G(\omega^2)$ is rational wrt ω^2 and satisfy to passive systems criteria :

$$0 \le G(\omega^2) \le 1$$



* Determine the reflection coefficient ρ(s) with minimum phase using spectral factorization

$$\rho(-s)\,\rho(s) = 1 - G(-s^2) = \frac{N(s^2)}{M(s^2)}$$

* Expand $N(s^2)$ and $M(s^2)$ in a product of two polynomials

$$\frac{N(s^2)}{M(s^2)} = \frac{n(-s)\,n(s)}{m(-s)\,m(s)}$$

where n(s) and m(s) are Hurwitz's polynomials built on the zeros of $N(s^2)$ and $M(s^2)$.

The reflection coefficient ρ(s) can be written relatively to an all-pass function B(s):

$$\rho(s) = \pm\frac{n(s)}{m(s)} = \prod_{i=1}^{m} \frac{s - s_i}{s + s_i}\, S_{22}(s) = B(s)\, S_{22}(s)$$

where the s_i are the poles of $Z(-s)$ for { Re $(s) > 0$ }. By substituting S_{22}, we obtain:

$$Z_q(s) = \frac{\left[Z(s) + Z(-s)\right]B(s)}{B(s) - \rho(s)} - Z(s)$$

* Use the impedance as the input impedance of the matching network.

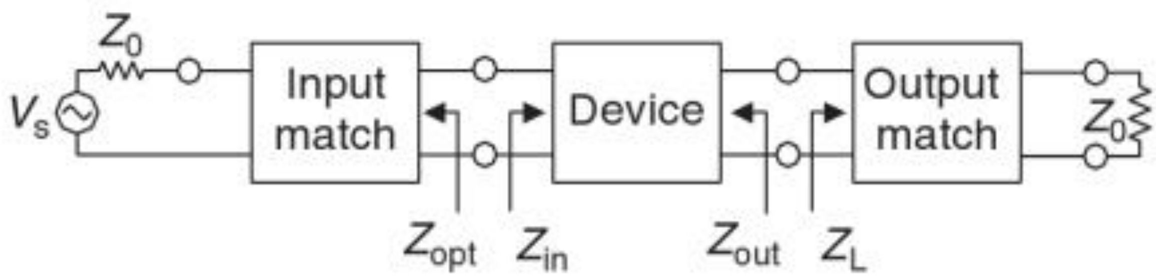
NONLINEAR AMPLIFIERS :

- High power amplifiers

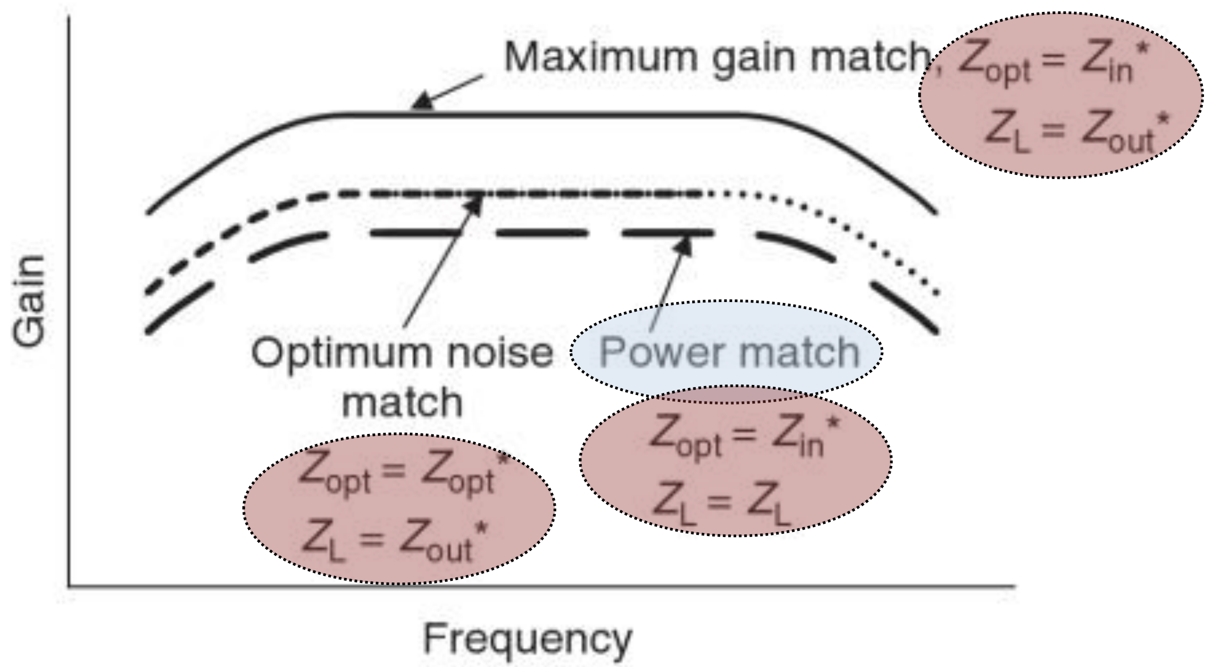
HIGH-POWER AMPLIFIERS

What is different vs. linear amplifiers?

Favouring Power … but what about **Gain**?



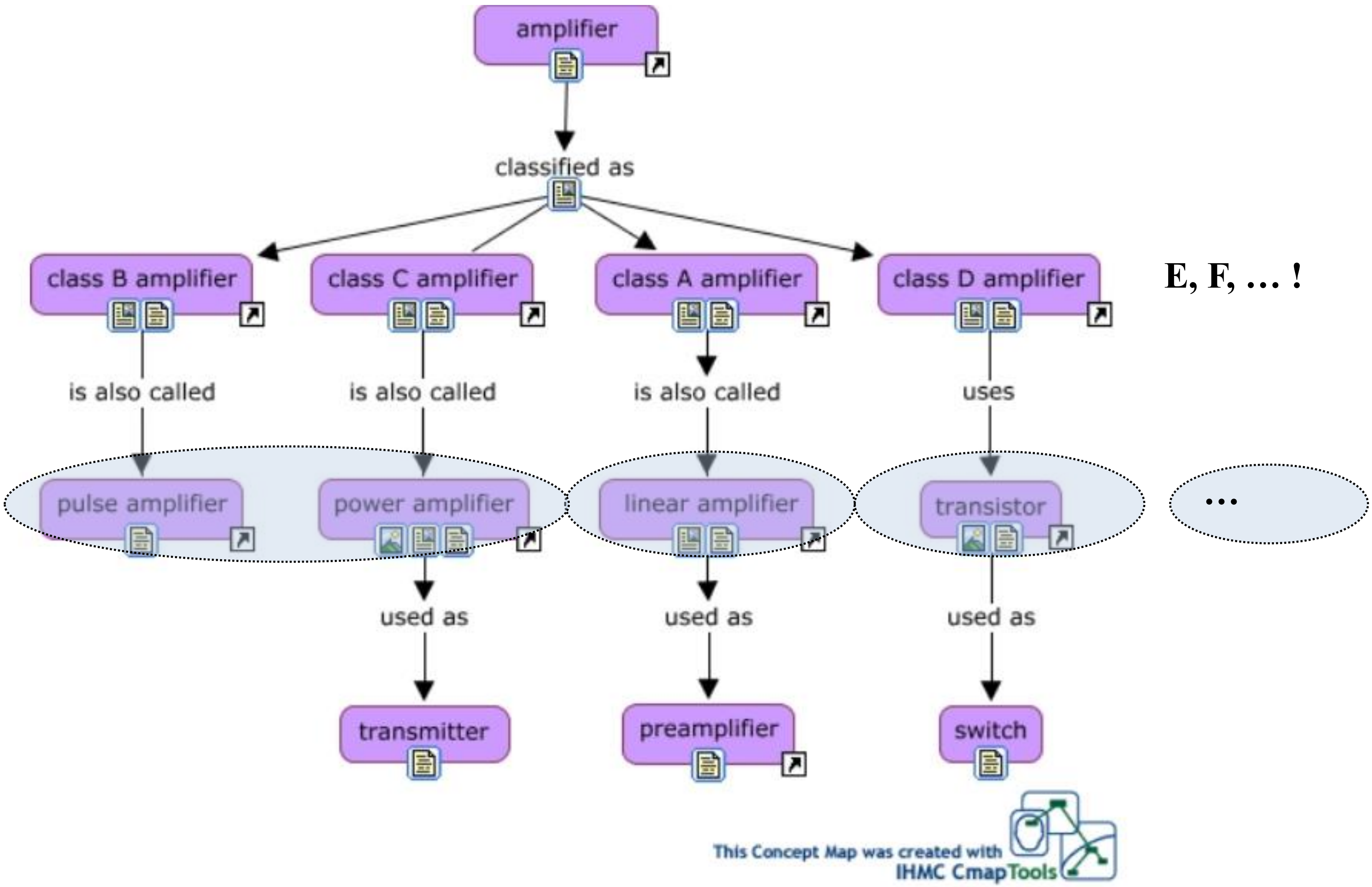
(a)



HIGH-POWER AMPLIFIERS

What is different vs. linear amplifiers?

Classes !



E, F, … !

HIGH-POWER AMPLIFIERS :

- **Classes**

HIGH-POWER AMPLIFIERS

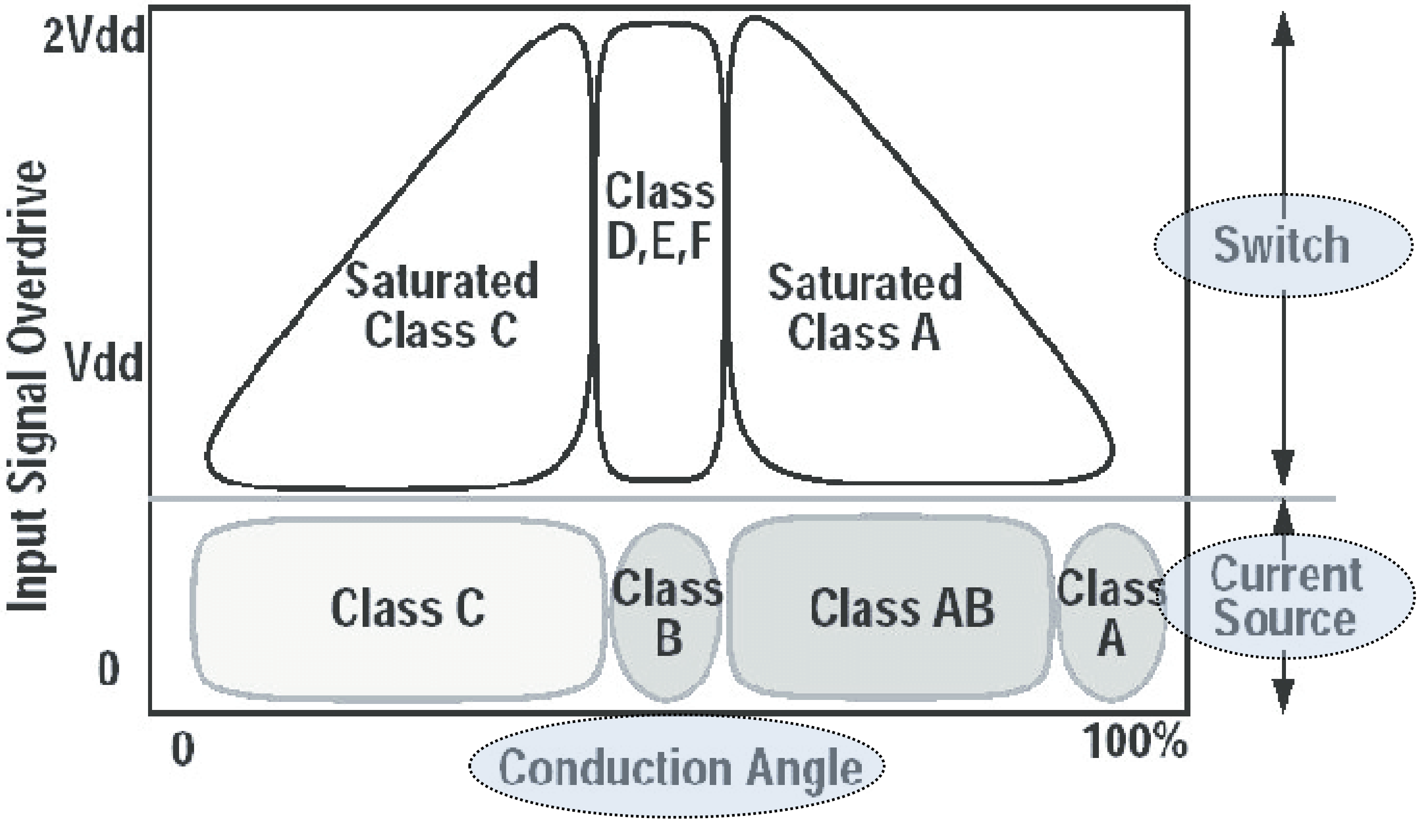
Classes

Why class A (linear amplifier) is not efficient in **power amplification** ?

- Class A output stage is a simple **linear** current amplifier.
- Single transistor can only conduct in one direction.
- D.C. bias current is needed to cope with negative going signals.
  - 75 % (or more) of the supplied power is dissipated by D.C.
    - ➢ It is **very inefficient**, typical maximum efficiency between 10 and 20 %.
- Only suitable for **low power** applications.
- **High power** requires much **better efficiency**.

HIGH-POWER AMPLIFIERS

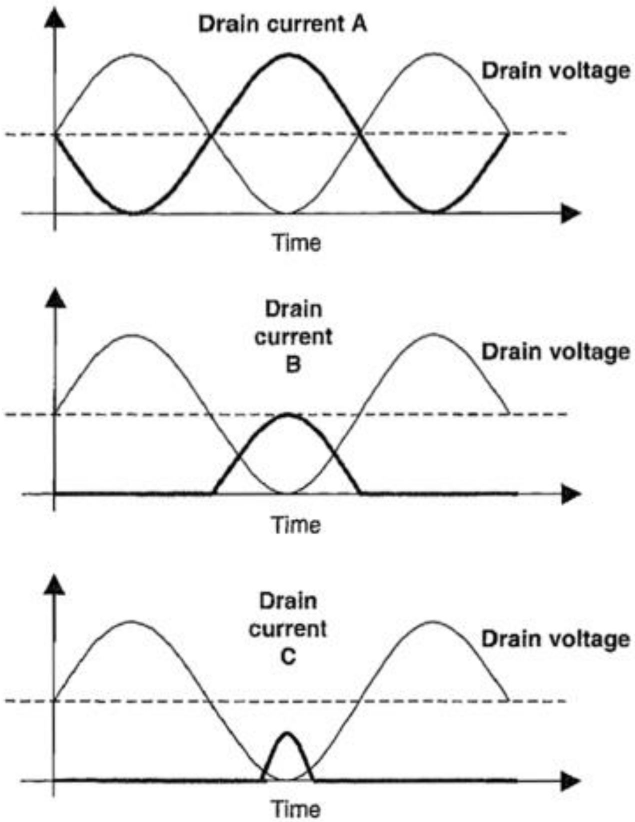
Classes

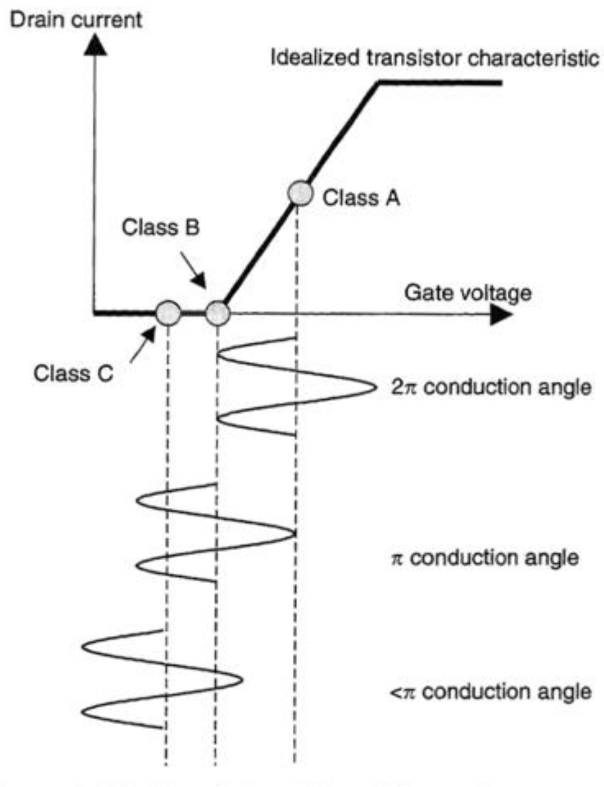


https://www.researchgate.net/figure/Figure219-Classical-Definition-of-power-amplifier-classes_fig15_310706437

HIGH-POWER AMPLIFIERS

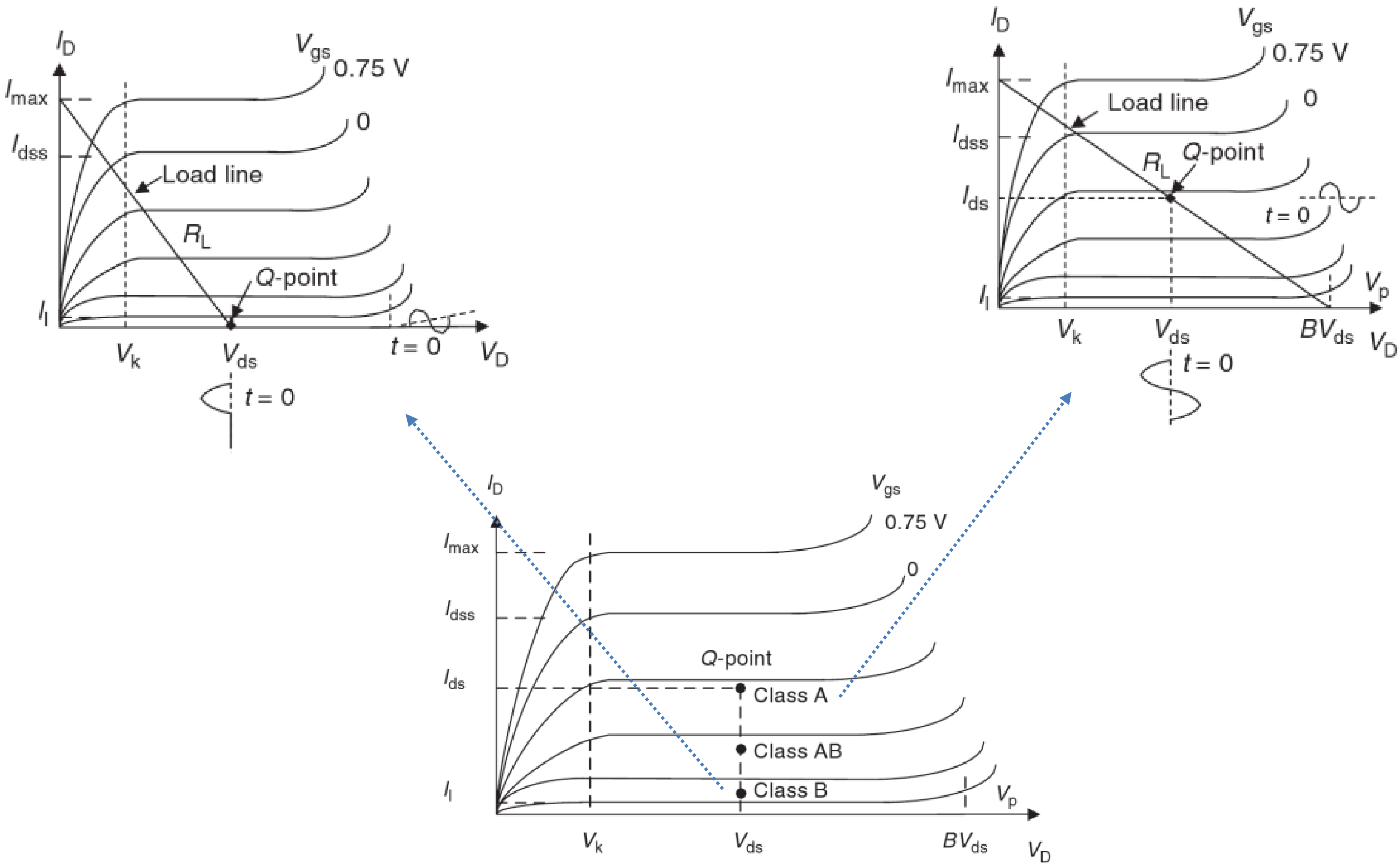
Classes: Conduction angle?





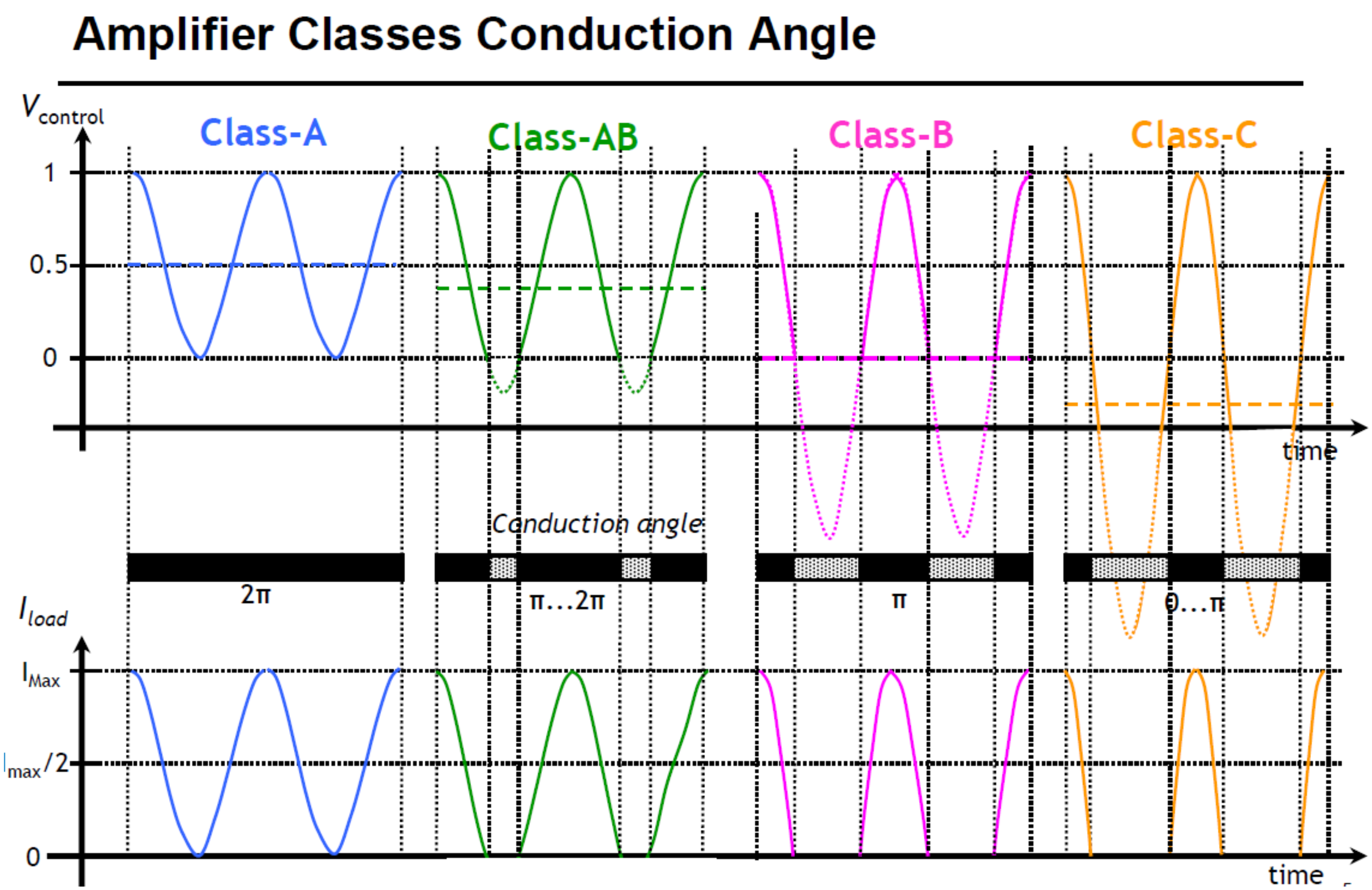
HIGH-POWER AMPLIFIERS

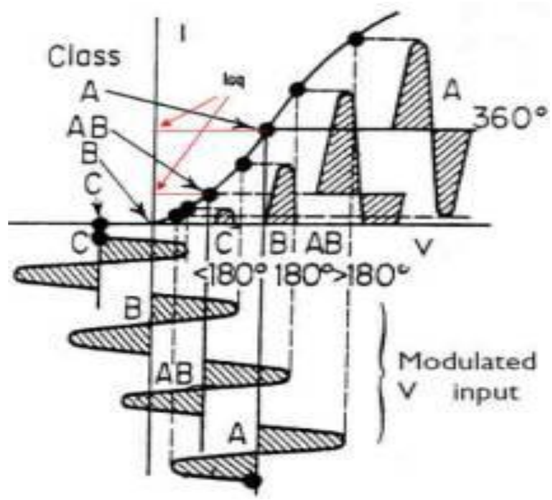
Classes: Conduction angle?



HIGH-POWER AMPLIFIERS

Classes: Signal waveforms !





HIGH-POWER AMPLIFIERS :

- **Current source amplifiers**

  (Classes A, AB, B, C)

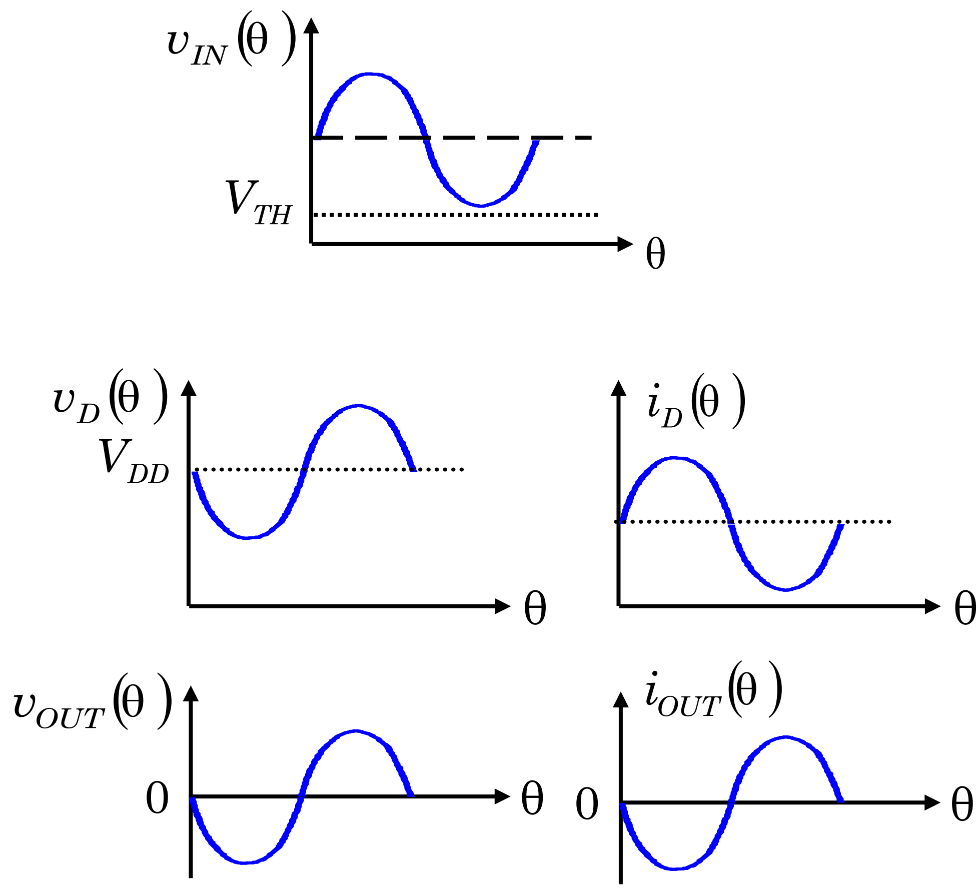
HIGH-POWER AMPLIFIERS

Classes: A, B, AB, C

Class A

$$P_{RFout} = \frac{V_{om}^{2}}{2R} \leq \frac{V_{DD}^{2}}{2R}$$

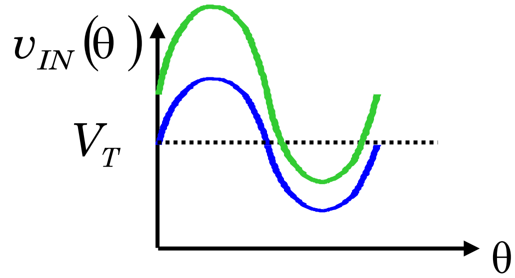
$$\eta_{Drain} = \frac{P_{RFout}}{P_{DC}} = \frac{V_{om}^{2}/2R}{V_{DD}^{2}/R} = \frac{V_{om}^{2}}{2V_{DD}^{2}} \leq \frac{1}{2}$$



HIGH-POWER AMPLIFIERS

Classes: A, B, AB, C

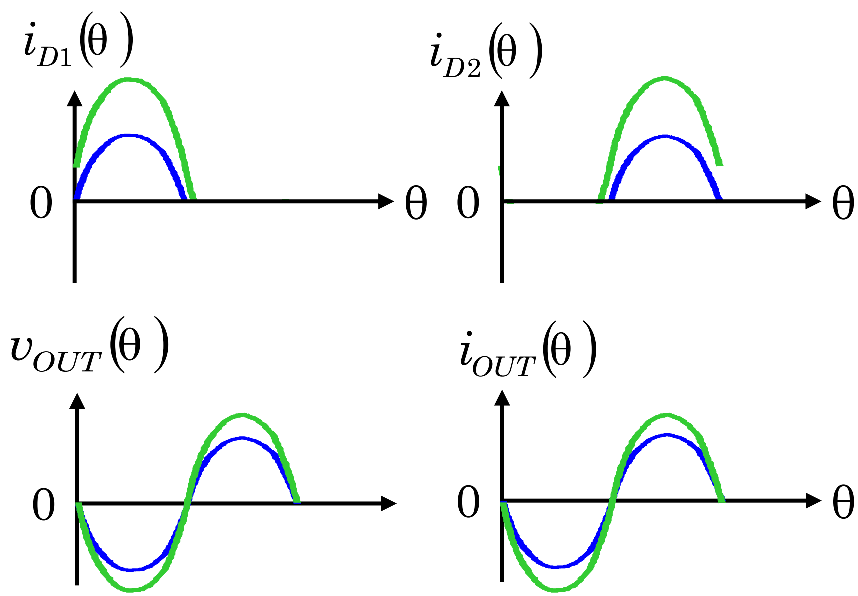
Class B, AB



$$P_{RFout} = \frac{V_{om}^2}{2R} \le \frac{V_{DD}^2}{2R}$$

$$I_{DC} = \frac{1}{2\pi}\int_0^{2\pi} I_D|\sin\theta|d\theta = \frac{2I_D}{\pi} = \frac{2}{\pi}\frac{V_{om}}{R}$$

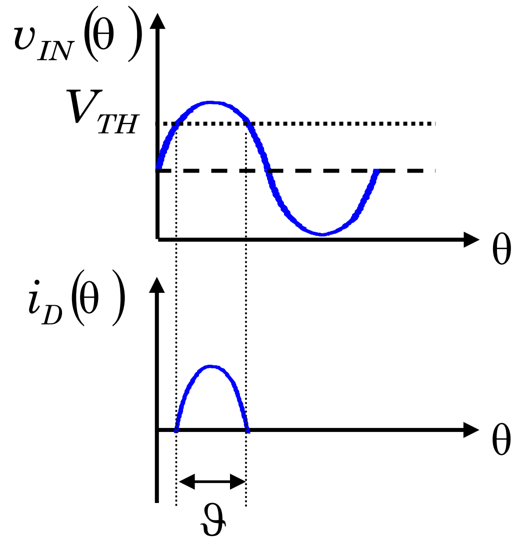
$$\eta_{Drain} = \frac{P_{RFout}}{P_{DC}} = \frac{V_{om}^2/2R}{\frac{2}{\pi}\frac{V_{om}}{R}V_{DD}} = \frac{\pi}{4}\frac{V_{om}}{V_{DD}} \le \frac{\pi}{4} \approx 0.785$$



HIGH-POWER AMPLIFIERS

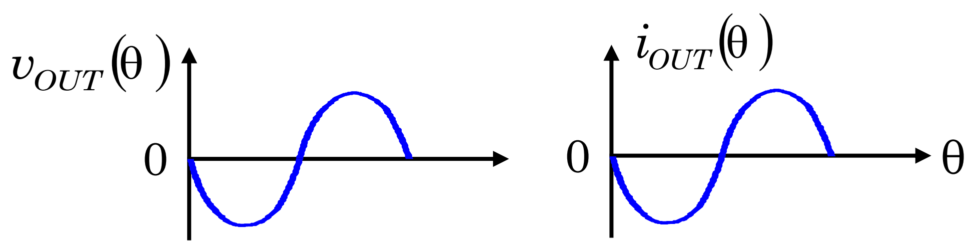
Classes: A, B, AB, C

Class C



$$P_{RFout} \propto \frac{\vartheta - \sin\vartheta}{1 - \cos(\vartheta/2)}$$

$$\eta_{Drain} = \frac{P_{RFout}}{P_{DC}} = \frac{1}{4}\,\frac{\vartheta - \sin\vartheta}{\sin(\vartheta/2) - \vartheta/2\cos(\vartheta/2)}$$



HIGH-POWER AMPLIFIERS

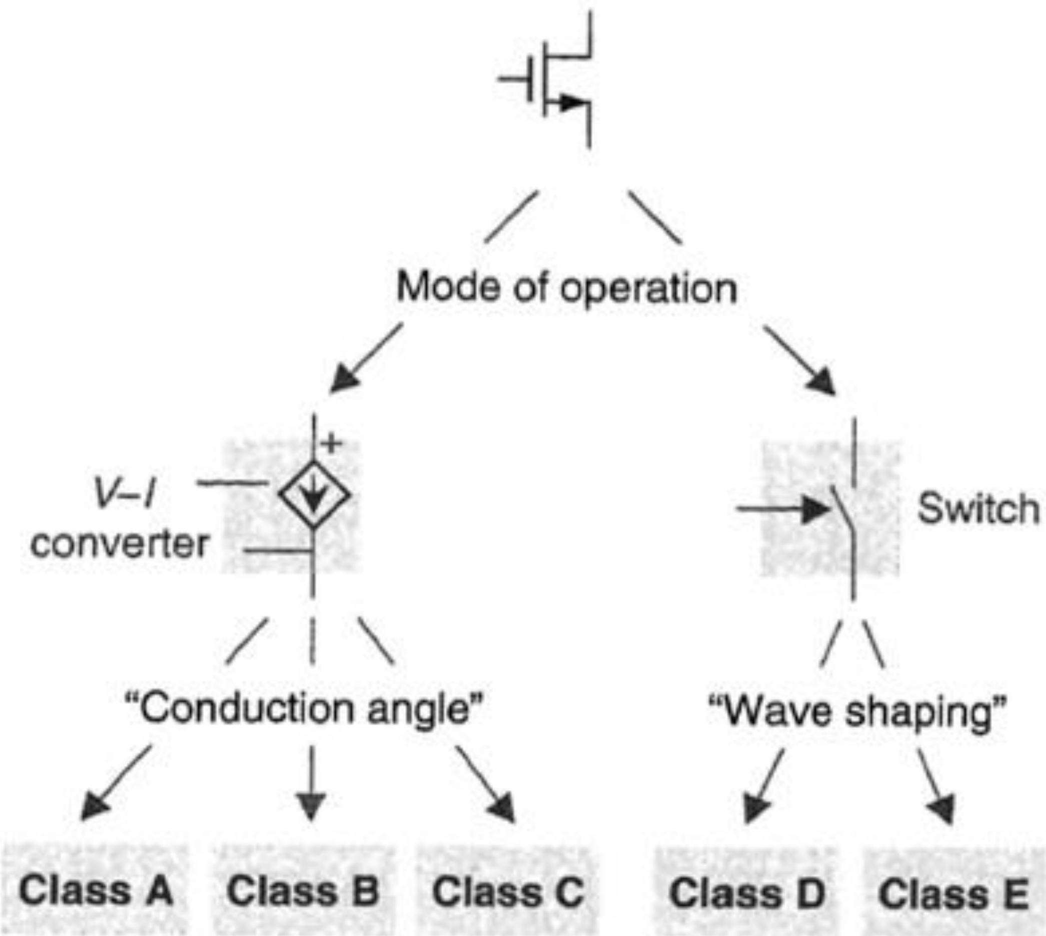
Classes: A, B, AB, C

Class	A	B	C	AB
Conduction Angle	360^0	180^0	Less than 90^0	180 to 360^0
Position of the Q-point	Centre Point of the Load Line	Exactly on the X-axis	Below the X-axis	In between the X-axis and the Centre Load Line
Overall Efficiency	Poor 25 to 30%	Better 70 to 80%	Higher than 80%	Better than A but less than B 50 to 70%
Signal Distortion	None if Correctly Biased	At the X-axis Crossover Point	Large Amounts	Small Amounts

https://www.homemade-circuits.com/understanding-amplifier-circuits/

HIGH-POWER AMPLIFIERS

Classes: Switching amplifiers?



- Class A: conduction angle 360º
- Class B: conduction angle 180º
- Class AB: conduction angle > 180º
- Class C: conduction angle < 180º
- Class D: an extension of class C
- Classes E, F: switch modes

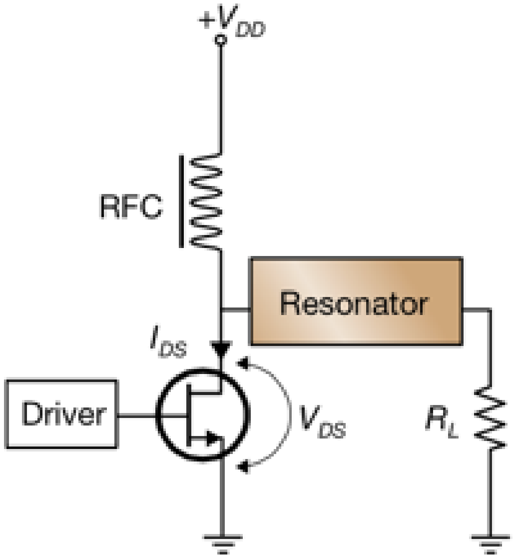
HIGH-POWER AMPLIFIERS :

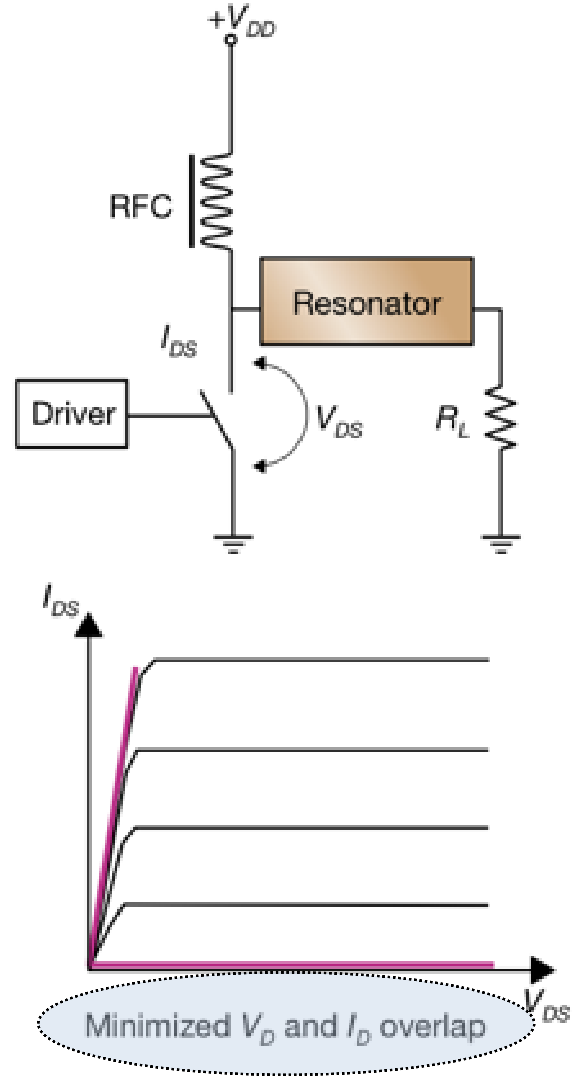
- **Switching amplifiers**

  (Classes D, E, F, F^{-1} ...)

HIGH-POWER AMPLIFIERS

Classes: Switching amplifiers





Classes: Switch Mode !

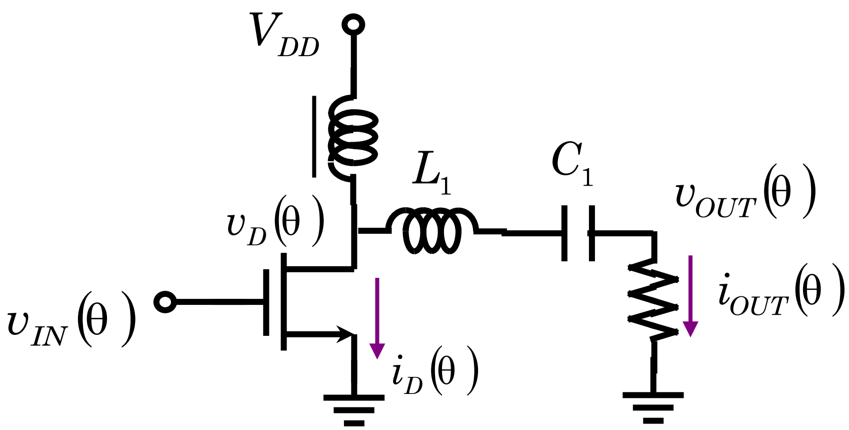
Highly Efficient Amplifier Classes - D, E, F, S -

- Amplifiing Element operates in Switching Mode
- States:
  - Device is fully switched – Current through device is maximal - but Voltage at device is zero – There is no loss power!
  - Device is fully open – Voltage at device is maximal – but Current through device is zero – There is again no loss power!
- Highly Efficient Operating Mode:
- Classes are different by their termination (complex load) of the harmonics
- More complicated amplifier architectures are necessary to amplify modulations with amplitude variations
- Very high efficiency also for small signal operation of the amplifier (good back-off efficiency)

HIGH-POWER AMPLIFIERS

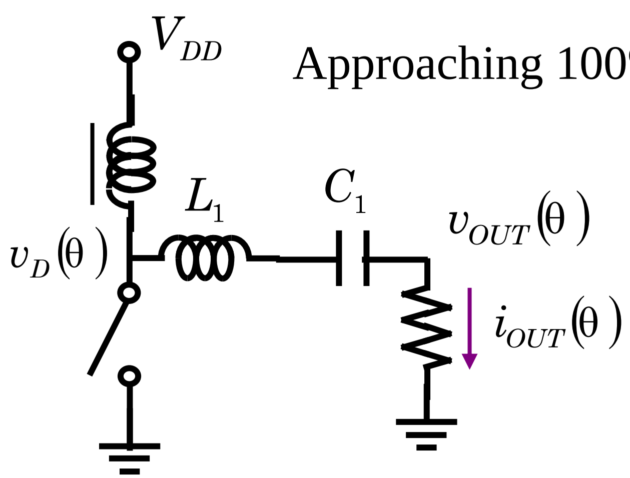
Classes: E, F, F^{-1}

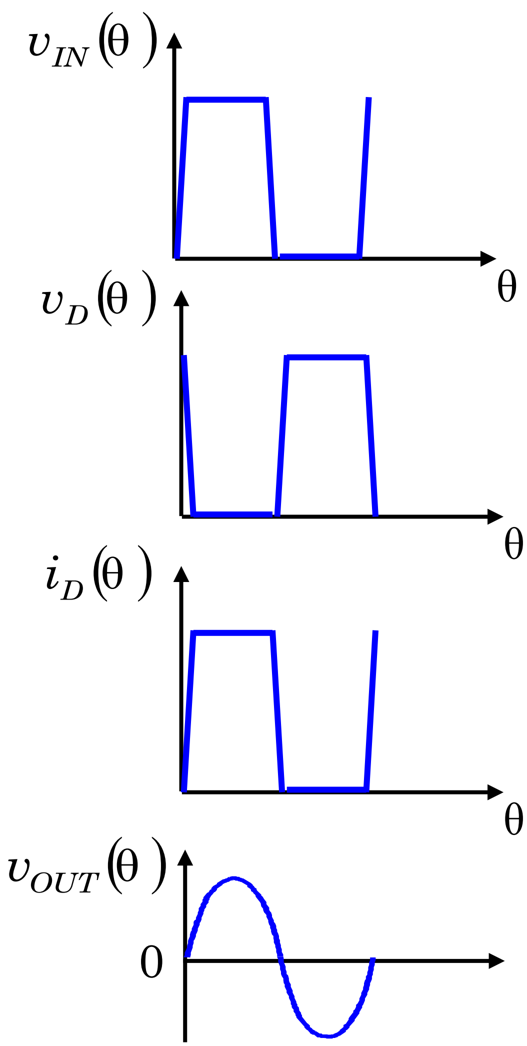
Class E



Switch mode

Approaching 100% efficiency

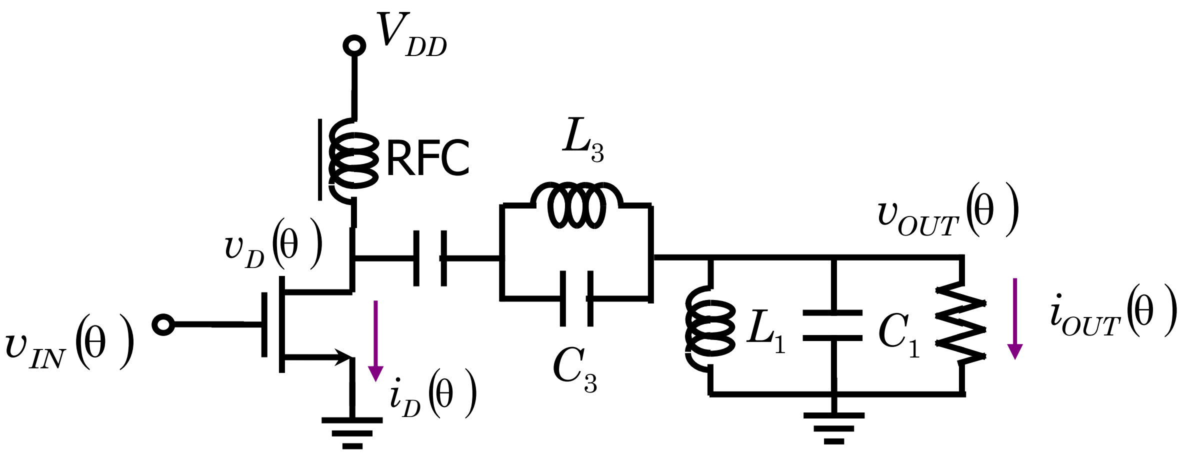


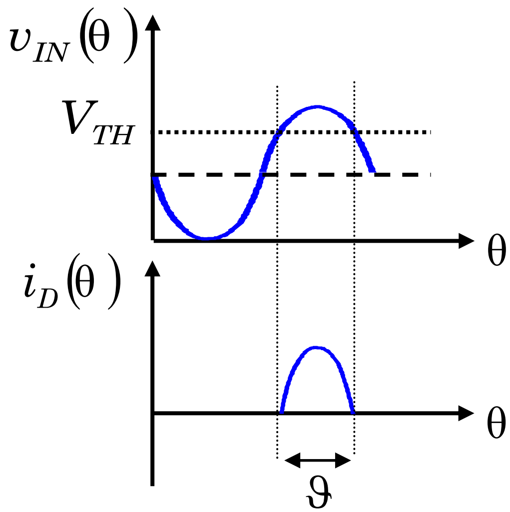


HIGH-POWER AMPLIFIERS

Classes: E, F, F^{-1}

Class F



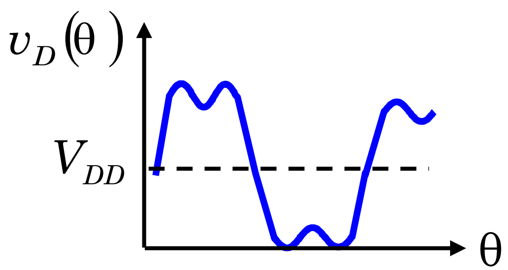


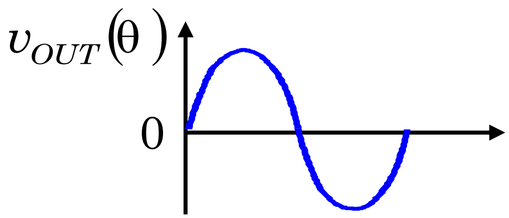
L_3C_3 tuned to the 2nd or 3rd harmonics

Peak efficiency

88% for 3rd harmonics peaking

85% for 2nd harmonics peaking.



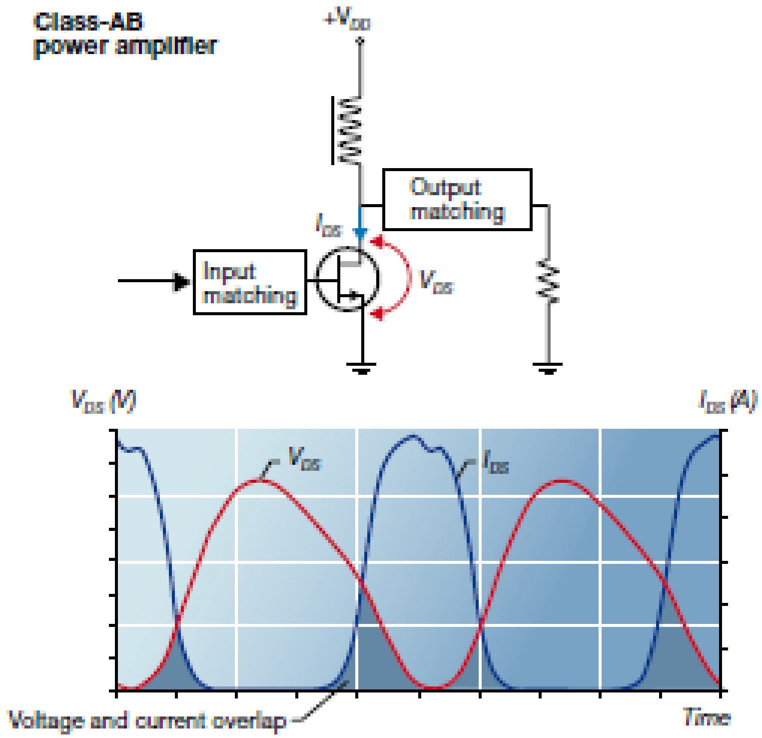


HIGH-POWER AMPLIFIERS :

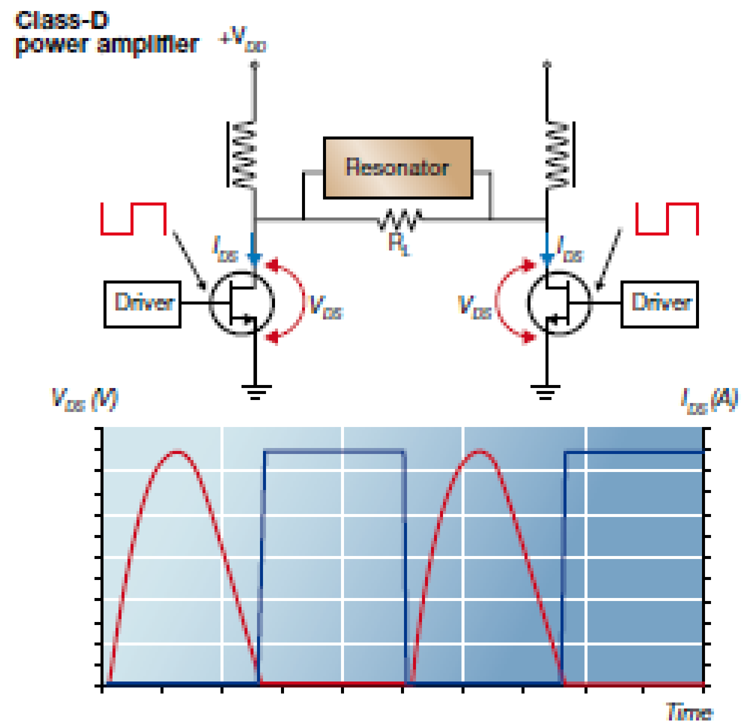
- **Comparison between classes**

HIGH-POWER AMPLIFIERS

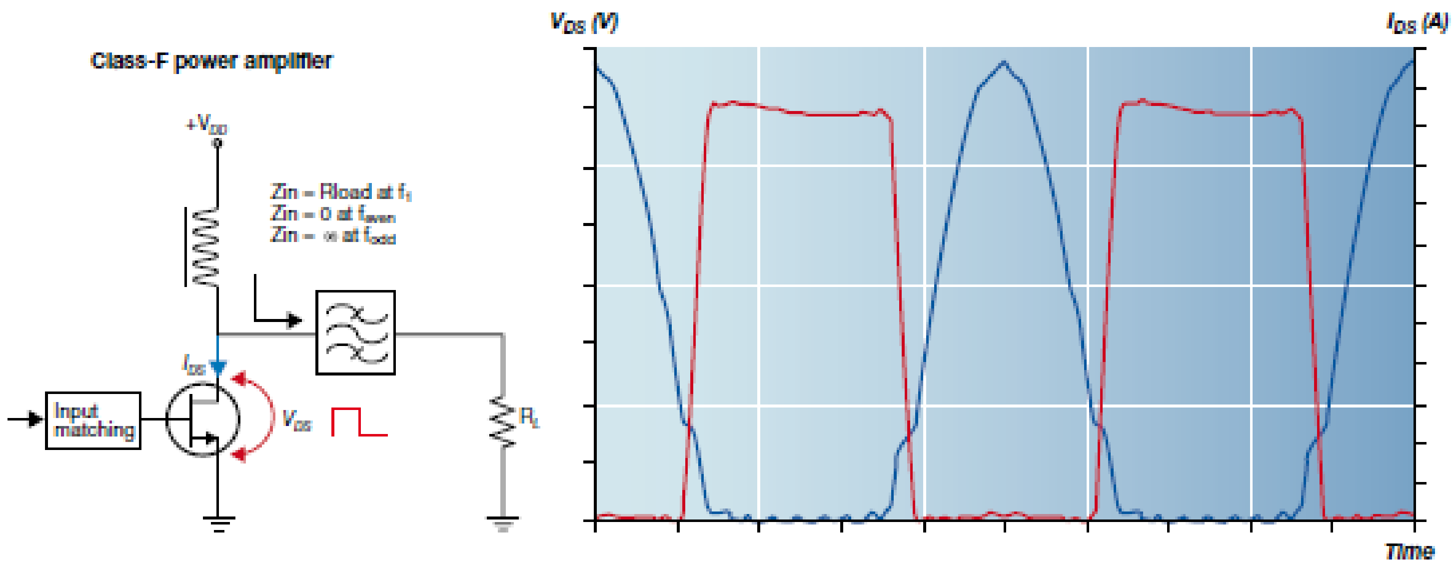
Classes: Comparison between AB, D, and F !



67% efficiency

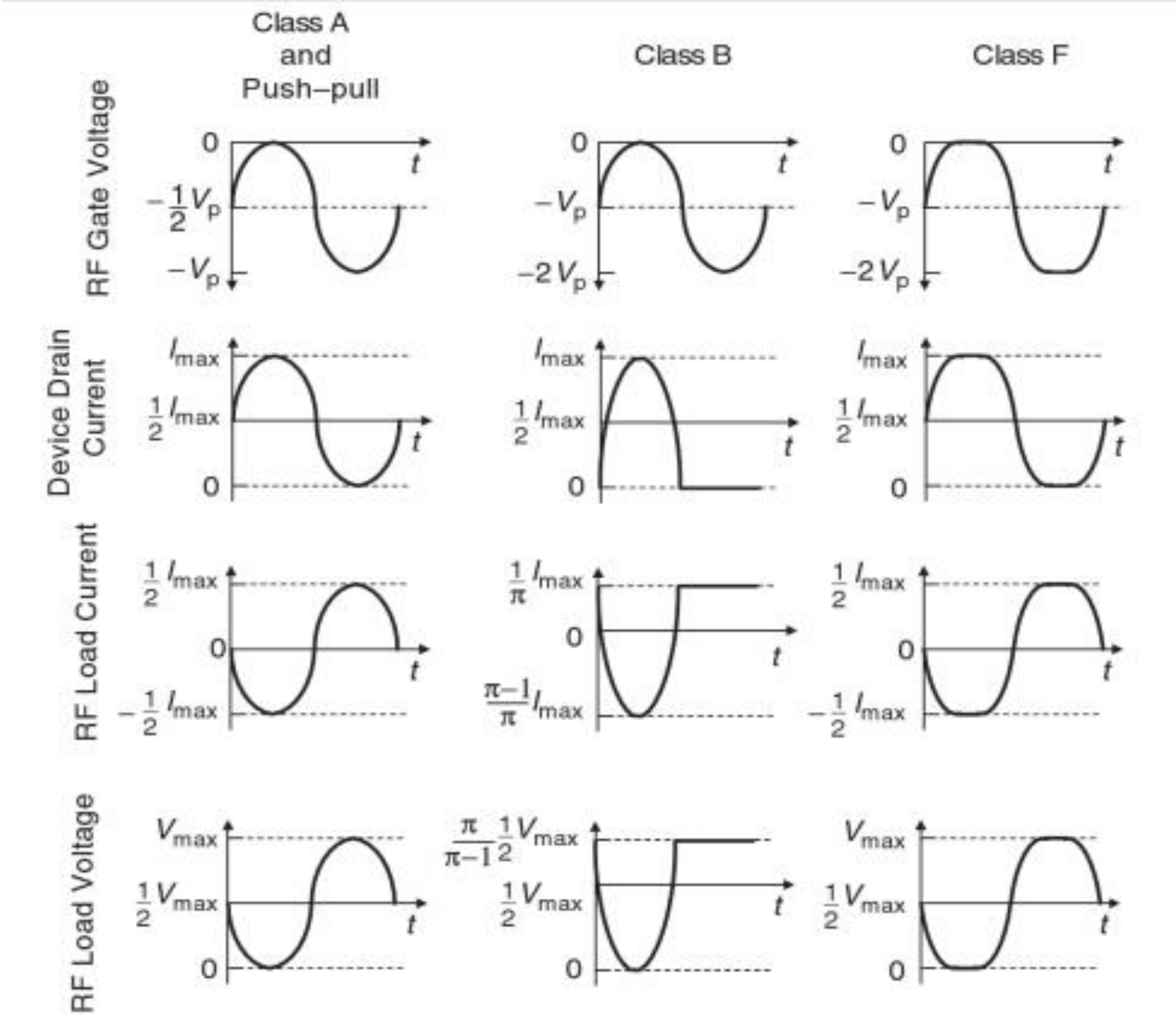


96% efficiency



HIGH-POWER AMPLIFIERS

Classes: Comparison between waveforms !



Classes: Comparison table

Design specifications	Suitable classes of amplifiers
High linearity	Class A, Class AB
High efficiency	Class AB, Class B, Class C, Class D, Class E, Class F, Class G, Class H
Low complexity	Class A, Class B, Class AB, Class C
Low noise	Class A, Class AB
Digital circuit design	Class D, Class E, Class F

HIGH-POWER AMPLIFIERS

Classes: Comparison table

Class	Type (1)	Linearity	Power Efficiency	Ease of implementation	Power capability	Gain (2)
A	LCA	Excellent	Poor	Very Good	Low	High
B	LCA	Medium	Good	Very Good	High	Medium/High
AB	LCA	Good	Fair	Good	Medium/High	High
C	LCA	Poor	Good	Fair	High	Low
D	Switcher	(3)	Very Good	Very Good	High	High
E	Switcher	Poor	Excellent	Good	High	Low/Medium
F	Switcher	Poor	Excellent	Fair	High	Medium
Inverse F	Switcher	Poor	Excellent	Fair	High	Medium

https://www.researchgate.net/figure/Brief-performance-comparison-between-various-amplifier-classes_tbl1_263115694

Classes of PA & Comparison

Power Amplifier Classes

Power Amplifier

Class A
- Total conduction angle is 360°
- Ideal efficiency of 25% is possible
- DUF of 0.125 is possible
- Used for linear applications

Class B Push-pull
- Total conduction angle is 360°
- Ideal efficiency of 78.4% is possible
- DUF of 0.125 is possible
- Used for linear applications

Class C
- Total conduction angle is less than 180°
- Ideal efficiency of 89.6% at 120° conduction
- DUF of 0.0981 is at 120° conduction

Class D
- Push-Pull arrangement of Class F
- Ideal efficiency of 100% is possible
- DUF of 0.318 is possible

Class E
- Optimum output power at 50% Duty cycle switching
- Ideal efficiency of 100% is possible
- DUF of 0.0981 is possible

Class F
- Odd order peaking resonators are used at odd harmonics
- Ideal efficiency of 100% is possible
- DUF of 0.159 is possible

HIGH-POWER AMPLIFIERS :

- **Trade-offs: linearity/distortions !!**

HIGH-POWER AMPLIFIERS

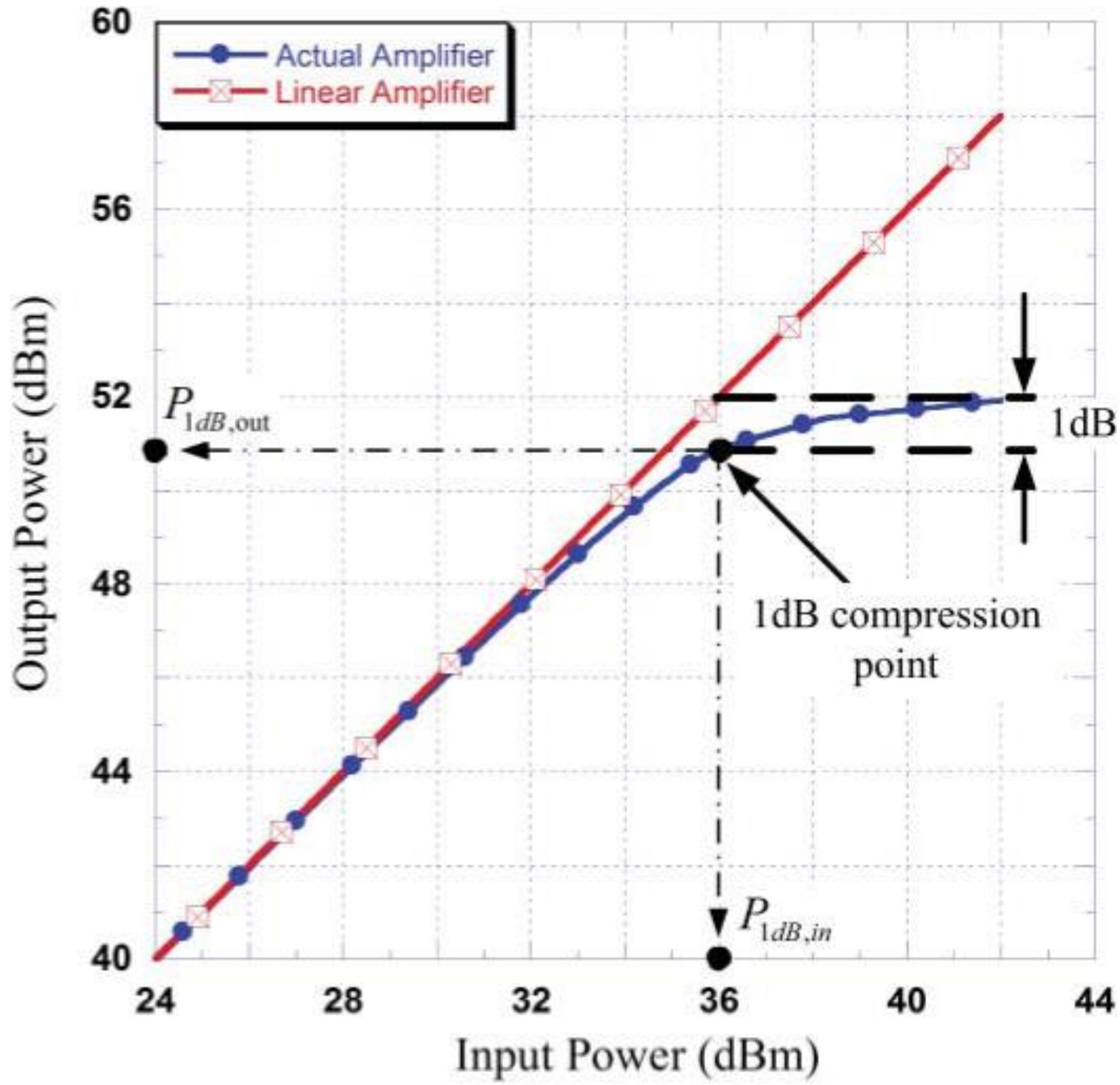
Power transfer function: not linear !

1 dB compression point from P_{out} vs. P_{in} characteristic

The 1-dB gain compression point is defined as

$$G_{1dB}(dB) = G_o(dB) - 1$$

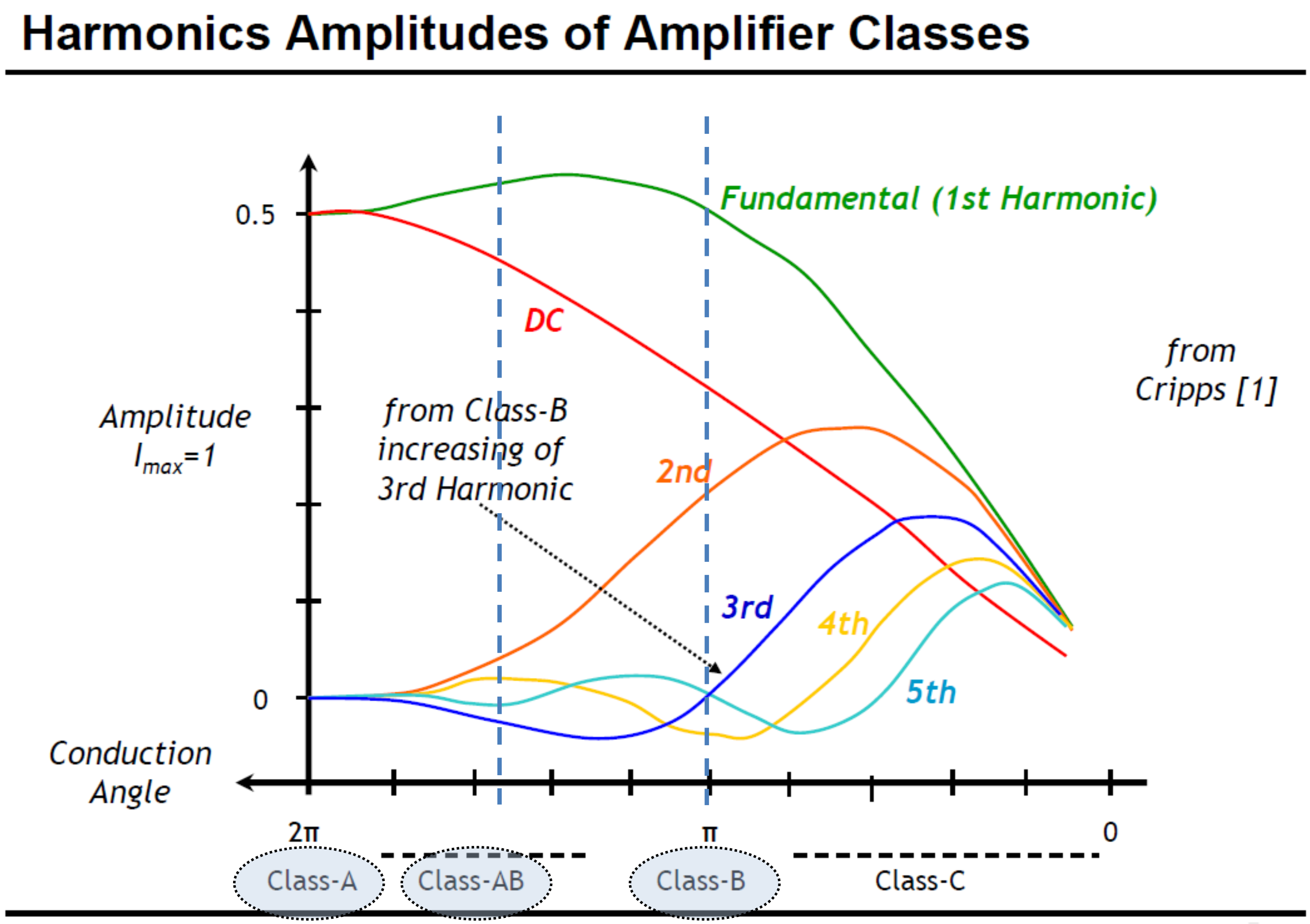
where G_o(dB) is the small signal linear power gain in decibels.



https://ebrary.net/78582/computer_science/compression_point

HIGH-POWER AMPLIFIERS

One input : Generation of harmonics !



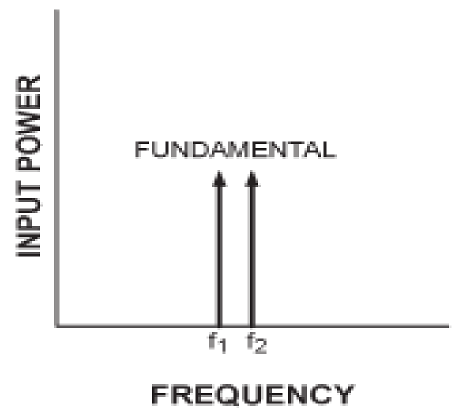
How to express such behaviour ?

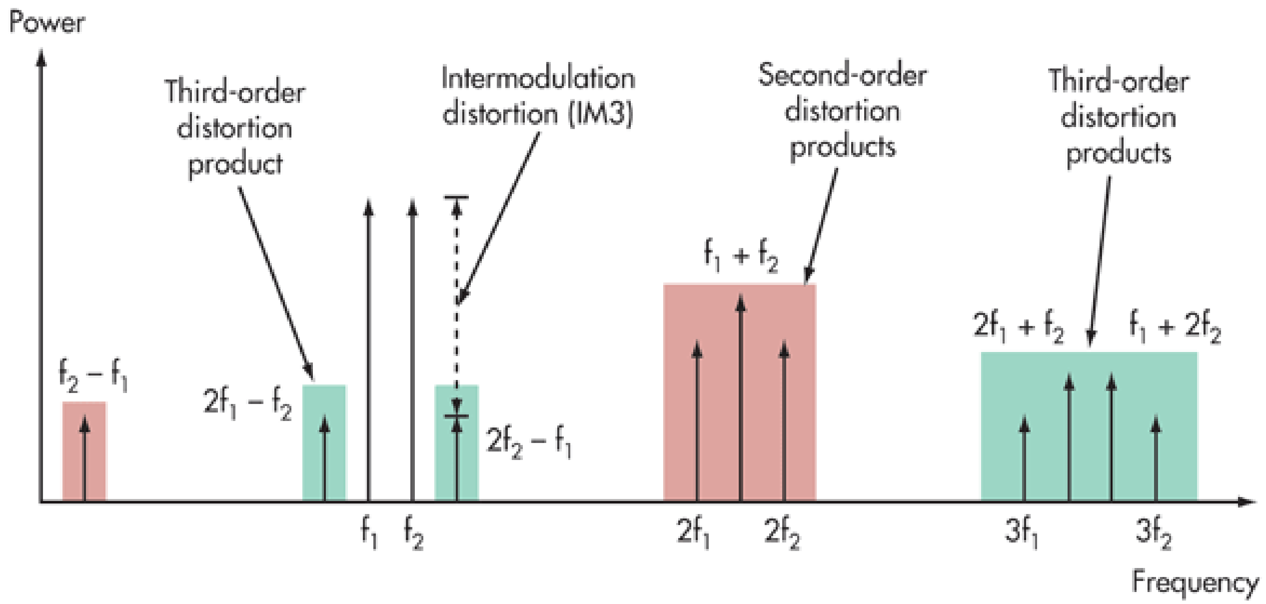
HIGH-POWER AMPLIFIERS

Two inputs : Generation of intermodulation products !

Intermodulation distortions (IMD)

$$V_{out} = a_0 + a_1.V_{in} + a_2.V_{in}^{\ 2} + a_3.V_{in}^{\ 3} + ...$$

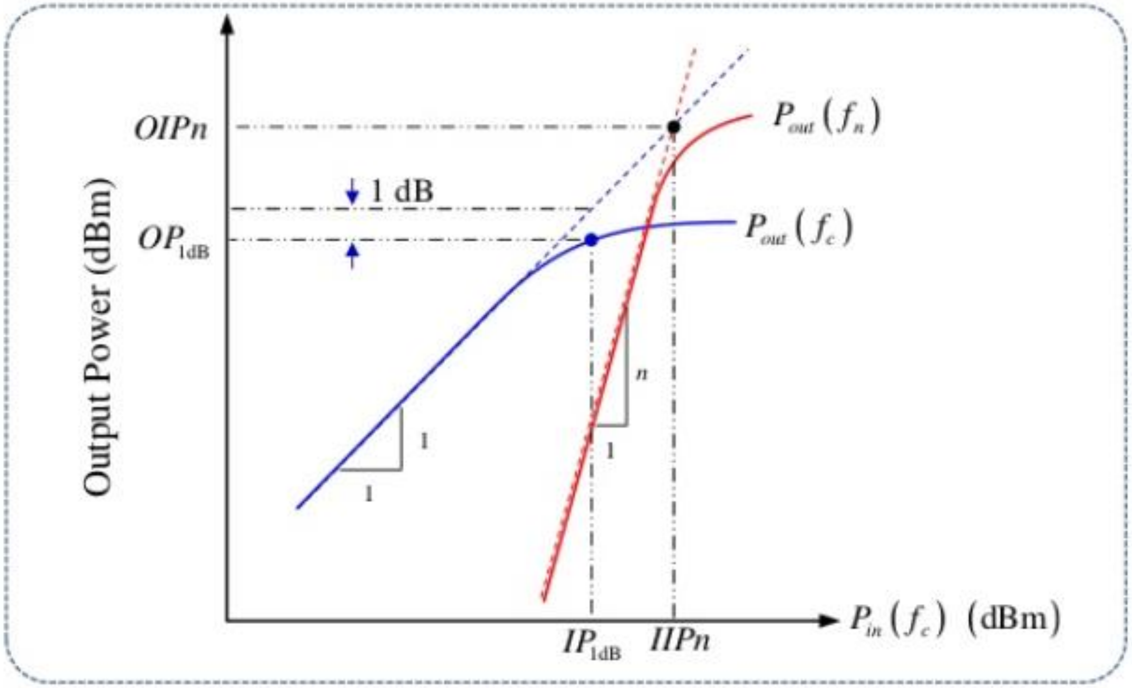




HIGH-POWER AMPLIFIERS

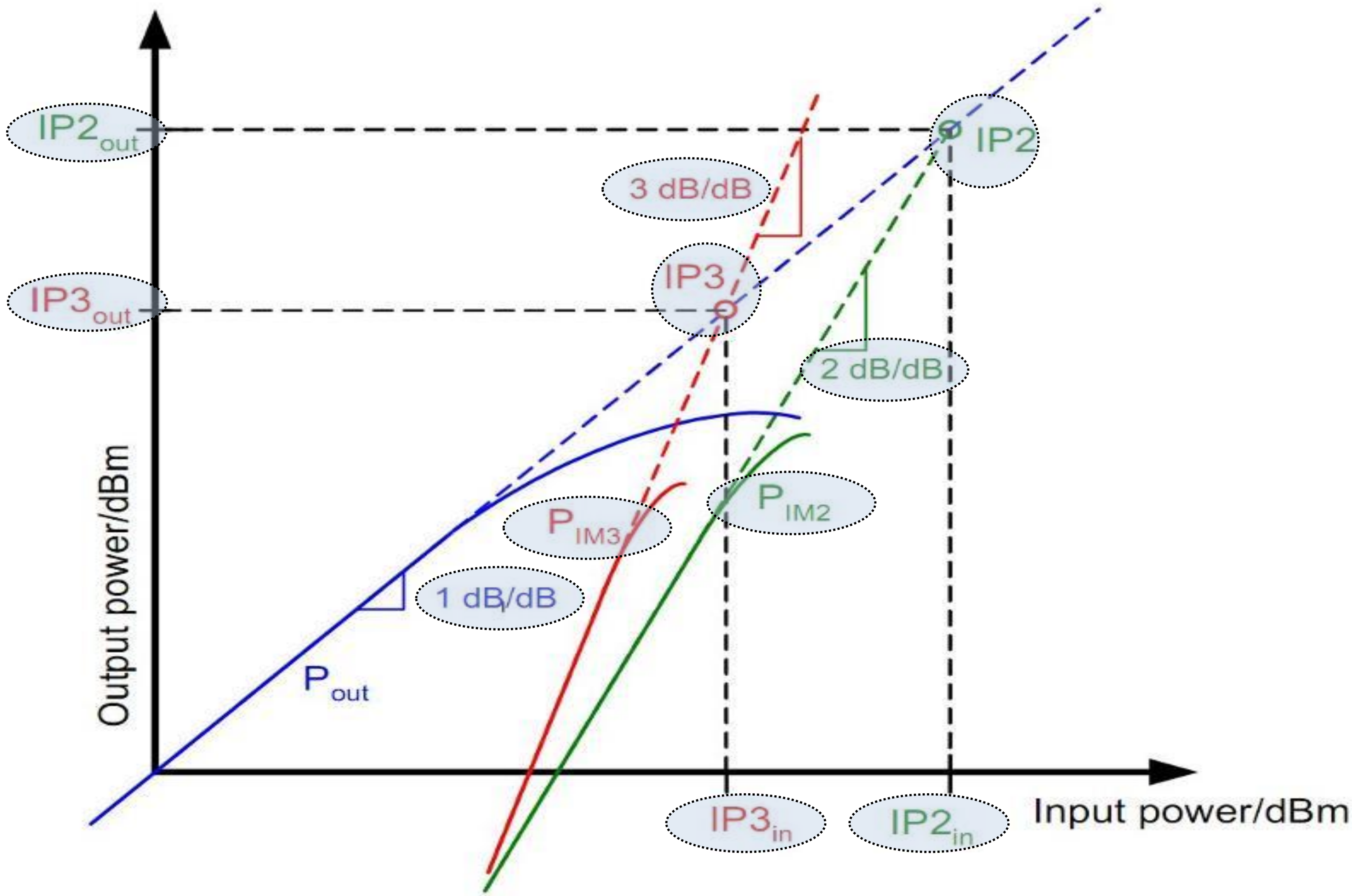
How to quantify that ? Intercept points !

- The nonlinear properties can be described by the concept of ***intercept points*** **(IPs)**. The input intercept point (IIPn) is a fictitious input power where the desired output signal component equals in amplitude the undesired component.



HIGH-POWER AMPLIFIERS

Intercept points !

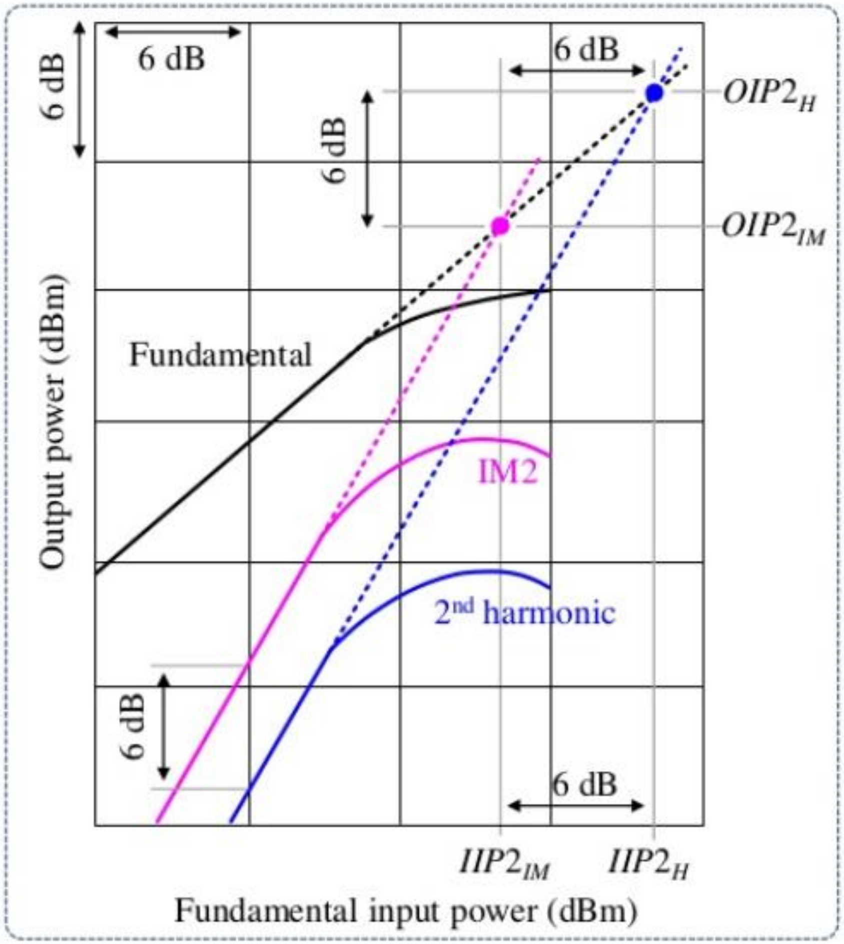


HIGH-POWER AMPLIFIERS

Intercept points !

Second-Order Intercept Point

- The 2nd-order products increase twice as fast as the desired fundamental, the straight lines cross. At the crossing point, either for the intermod or the harmonic, the fundamental and the 2nd-order product have equal output powers.
- Since the slopes of the straight lines are known, these crossing points, called intercept points (IPs), define the 2nd-order products **at low levels**.
- Typically, the larger of the input or output intercept points is specified; so amplifiers use OIPs and mixers use IIPs. Some may even add the power of the two fundamentals, increasing the value of the IP by 3 dB.



Intercept points !

Third-Order Nonlinear Effect (I)

- Consider only the first-order and the third-order effect of a nonlinear device, i.e., $v_{out} = \alpha_1 v_{in} + \alpha_3 v_{in}^3$.
- **Single-tone excitation**:

  The input signal contains only a sinusoidal signal $v_i = A\cos\omega_1 t$, where its available power can be obtained as $P_{in} = A^2/(2Z_{in})$.

- **In-band and out-of-band distortions**

  The output voltage becomes $v_{out} = \alpha_1 A\cos\omega_1 t + \alpha_3 A^3 \cos^3\omega_1 t$

$$= \left(\alpha_1 A + \frac{3}{4}\alpha_3 A^3\right)\cos\omega_1 t + \frac{1}{4}\alpha_3 A^3 \cos 3\omega_1 t$$

$$= \left(V_1^{(1)} + V_1^{(3)}\right)\cos\omega_1 t + V_3^{(3)} \cos 3\omega_1 t$$

Desired Signal
linear effect

In-band Distortion
3rd-order effect

Out-of-band Distortion
3rd-order effect
3rd harmonic

Intercept points !

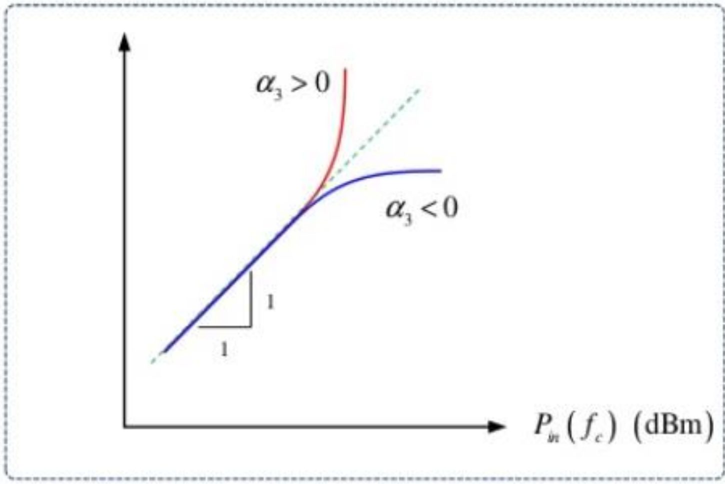
Third-Order Nonlinear Effect (II)

- ***Gain Compression or Enhancement***:

  At f_1, the amplified linear-term signal has been mixed with the third-order term

$$v_{out}(f_1) = \left(\alpha_1 A + \frac{3}{4}\alpha_3 A^3\right)\cos\omega_1 t$$

If $\alpha_3 < 0$, the linear gain is compressed, otherwise, it is enhanced



HIGH-POWER AMPLIFIERS

Intercept points !

Third-Order Nonlinear Effect (III)

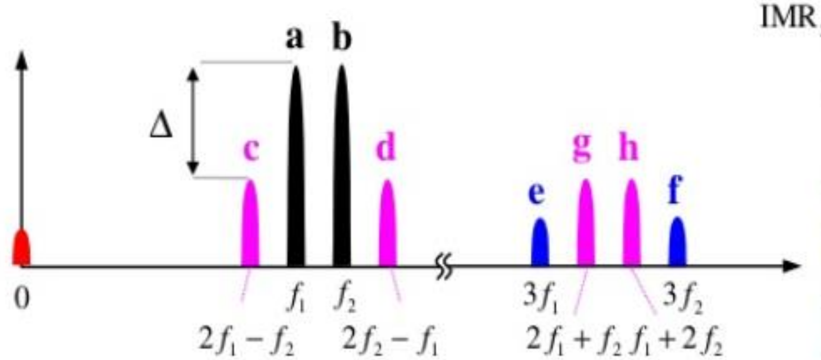
- **Two-tone excitation:** $v_{in}(t) = A\sin\omega_1 t + B\sin\omega_2 t,\ \omega_1 < \omega_2$

$$v_{out}(t) = \alpha_1 v_{in}(t) + \alpha_3 v_{in}^3(t)$$

$$= \overset{\mathbf{i}}{\left(\frac{3}{2}\alpha_3 A^2 B + \frac{3}{2}\alpha_3 AB^2\right)} + \overset{\mathbf{a}}{\left(\alpha_1 A + \boxed{\frac{9}{4}\alpha_3 A^3}\right)\cos\omega_1 t} + \overset{\mathbf{b}}{\left(\alpha_1 B + \boxed{\frac{9}{4}\alpha_3 B^3}\right)\cos\omega_2 t}$$

$$+ \overset{\mathbf{c}}{\frac{3}{4}\alpha_3 A^2 B\cos(2\omega_1 - \omega_2)t} + \overset{\mathbf{d}}{\frac{3}{4}\alpha_3 AB^2\cos(2\omega_2 - \omega_1)t} + \overset{\mathbf{e}}{\frac{1}{4}\alpha_3 A^3\cos 3\omega_1 t} + \overset{\mathbf{f}}{\frac{1}{4}\alpha_3 B^3\cos 3\omega_2 t}$$

$$+ \overset{\mathbf{g}}{\frac{3}{4}\alpha_3 A^2 B\cos(2\omega_1 + \omega_2)t} + \overset{\mathbf{h}}{\frac{3}{4}\alpha_3 AB^2\cos(\omega_1 + 2\omega_2)t}$$



$$\mathrm{IMR}_{2\text{-tone}} = \Delta = 2(IIP3 - P_{in}) = 2(OIP3 - P_{out})$$

a, b : linear term(desired signal) +inband distortion

c , d : IM3, adjacent band distortion

e, f : 3rd harmonics

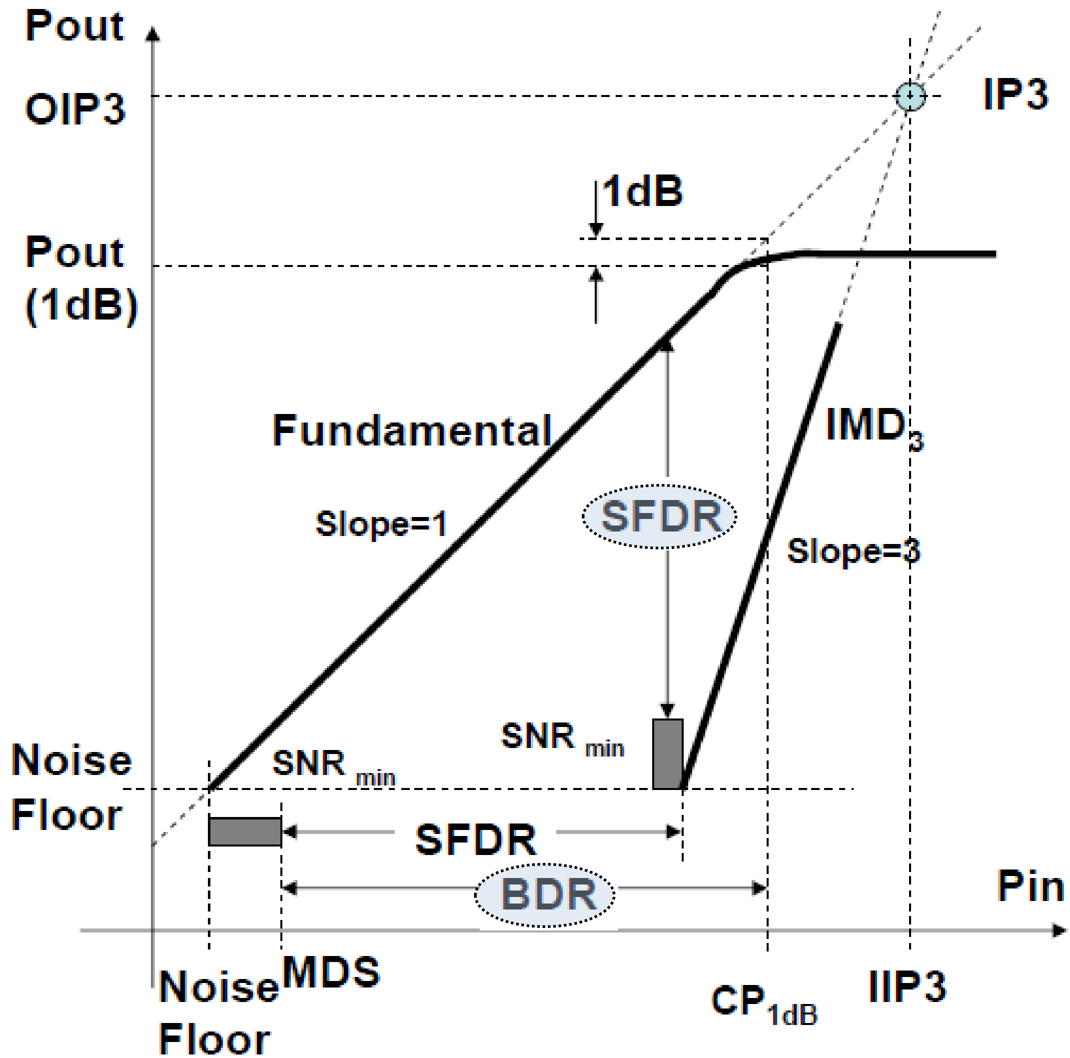
g, h : out of band distortion

i : DC term

HIGH-POWER AMPLIFIERS

Dynamic range !

Engineers measure dynamic range through various techniques. Ideally, it is the difference between noise floor (weakest signals) and clipping level. This is the signal-to-noise ratio (SNR). You can amplify/attenuate to keep signals in the Automatic gain control (AGC) sweet spot. This spot is called **Blocking Dynamic Range (BDR).** A more interesting metric is **Spurious Free Dynamic Range (SFDR)**, a measure relative to the fundamental carrier (dBc) or **Full-Scale Clipping Level** (dBFS).



RF circuit design: Basics, Akira Matsuzawa, Tokyo Institute of Technology
http://play.fallows.ca/wp/radio/software-defined-radio/sdr-receiver-performance-overview/

HIGH-POWER AMPLIFIERS

Dynamic range !

BDR: Blocking Dynamic Range

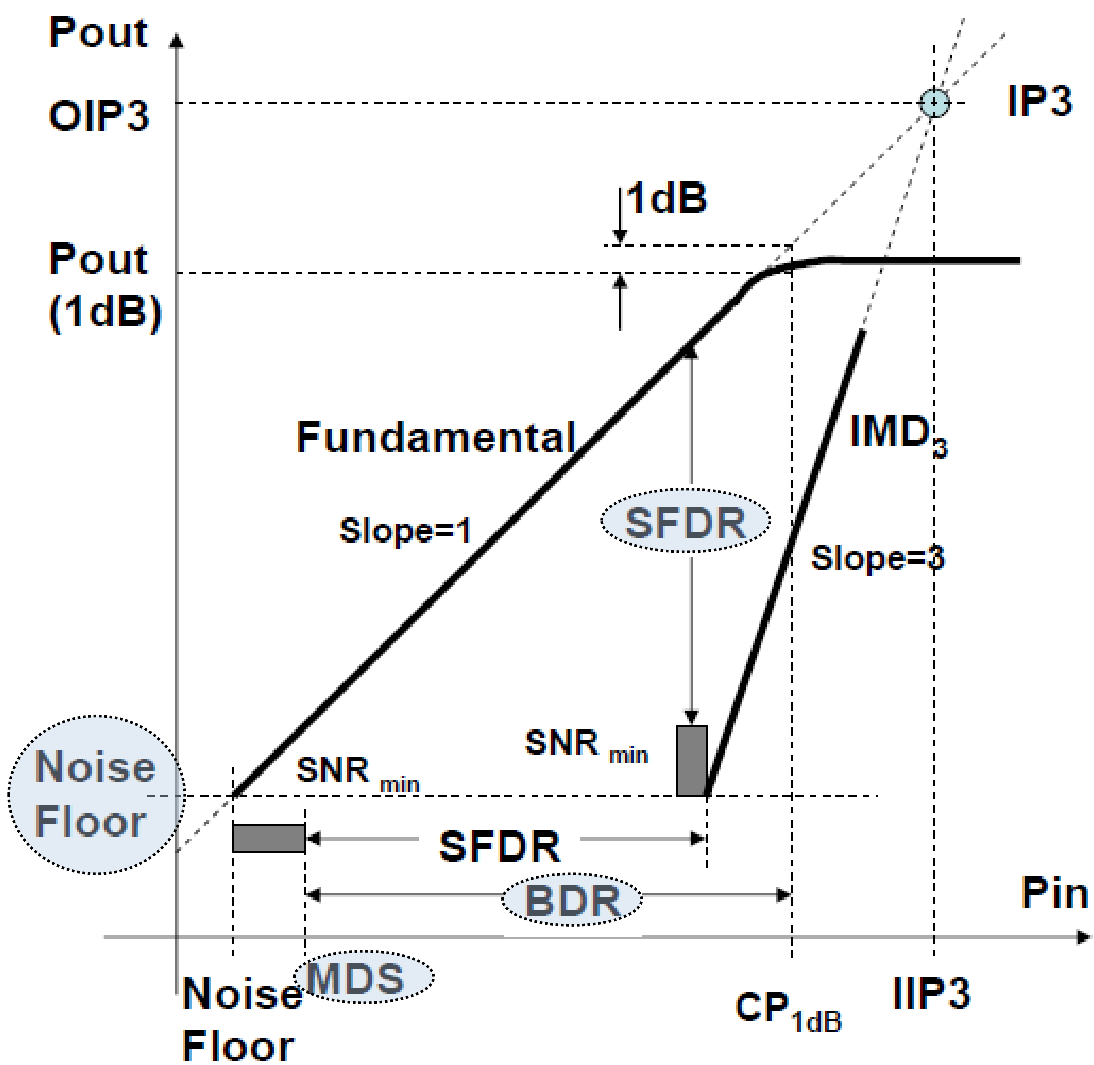
$$BDR = P_{1dB} - Noise\ Floor - SNR_{\min}$$

SFDR: Spurious-Free Dynamic Range

$$SFDR = \frac{2}{3}(IIP_3 - Noise\ Floor) - SNR_{\min}$$

MDS: Minimum Detectable Signal level
(= *Noise Floor* + *SNR* $_{\min}$)

$$Noise\ Floor = \underbrace{-174dBm}_{\text{kT limitation}} + NF + 10\log \underbrace{BW}_{\text{Bandwidth}}$$



RF circuit design: Basics, Akira Matsuzawa, Tokyo Institute of Technology
http://play.fallows.ca/wp/radio/software-defined-radio/sdr-receiver-performance-overview/

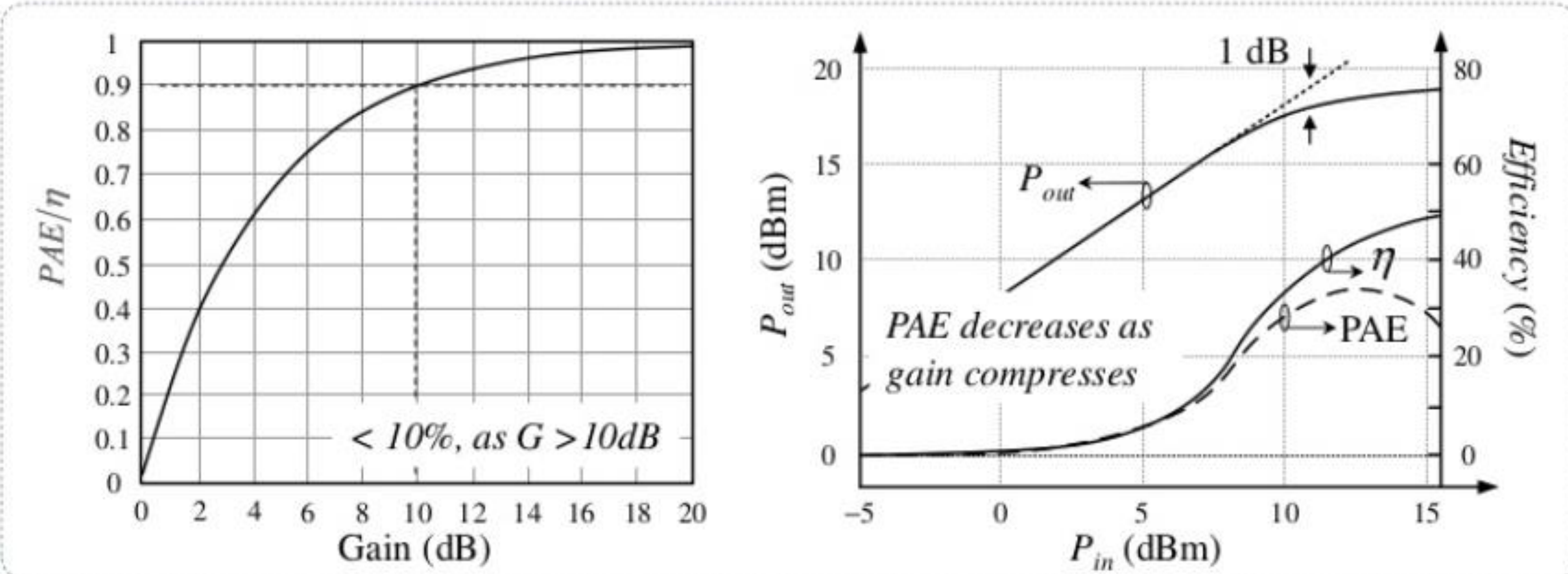
HIGH-POWER AMPLIFIERS

Power-Added Efficiency (PAE) !

- *Power-added efficiency* (PAE) takes the gain of the amplifier into account as follows:

$$PAE = \frac{P_{out} - P_{in}}{P_{dc}} = \frac{P_{out} - P_{out}/G}{P_{dc}} = \eta\left(1 - \frac{1}{G}\right)$$

For a high gain amplifier, PAE is the same as dc-to-RF efficiency η .



Thank you !

End of Chapter 4-1

Chapter 4-2

Microwave Amplifiers

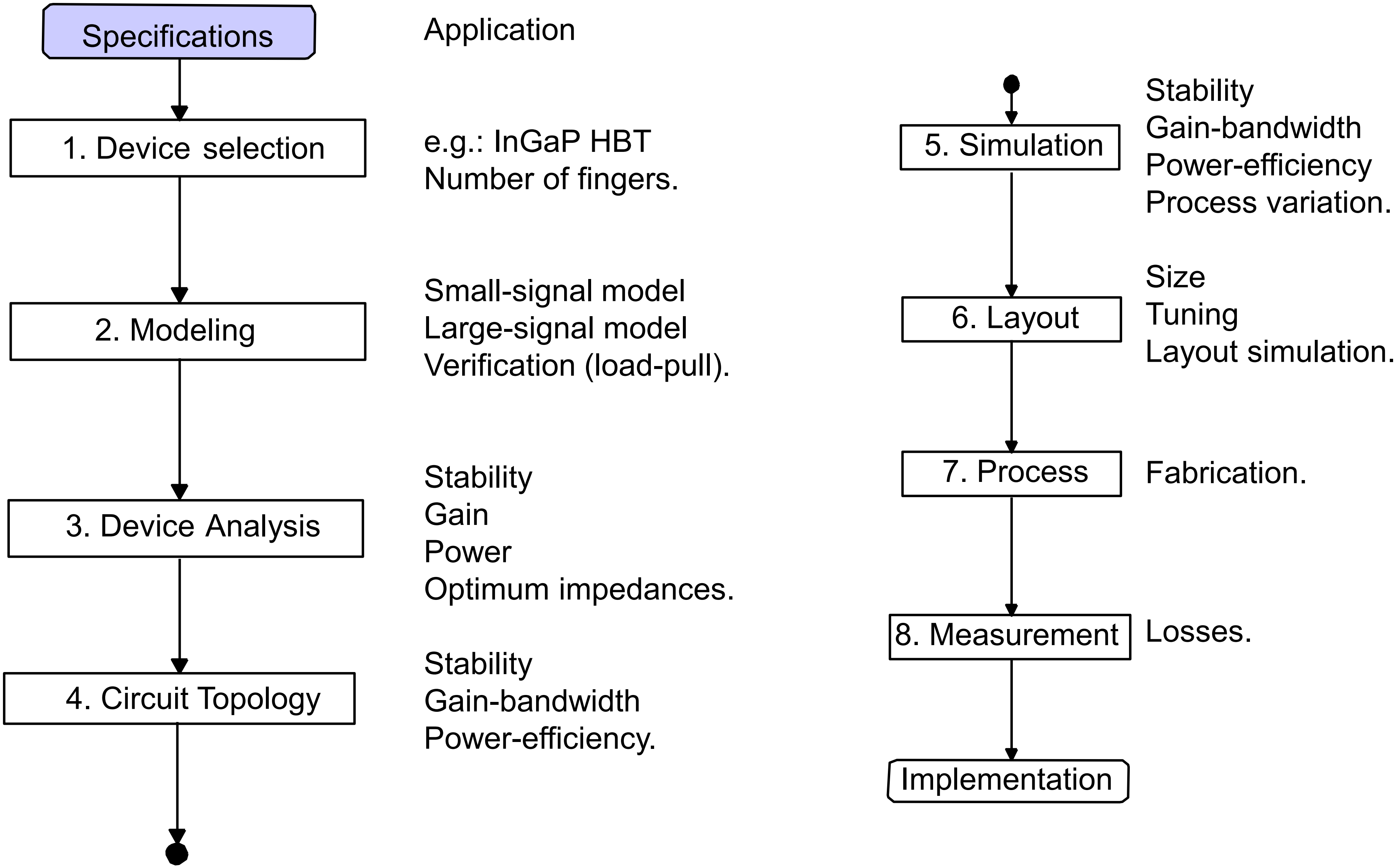
From theory to design:

Designing a low-noise amplifier (linear amplifier)

Design flow

Some slides are taken from

- Institute of Electronics (National Chiao Tung University)
- Fundamentals of RF and Microwave Transistor Amplifiers, Book by Inder Bahl (Wiley)
- Institut für Hochfrequenztechnik und Elektronik (Universität Karlsruhe)
- Dept. of Electric Eng., National Taipei University of Technology
- Keysight Technologies (ADS white papers)


Specifications
Application
1. Device selection
e.g.: InGaP HBT
Number of fingers.
2. Modeling
Small-signal model
Large-signal model
Verification (load-pull).
3. Device Analysis
Stability
Gain
Power
Optimum impedances.
4. Circuit Topology
Stability
Gain-bandwidth
Power-efficiency.
5. Simulation
Stability
Gain-bandwidth
Power-efficiency
Process variation.
6. Layout
Size
Tuning
Layout simulation.
7. Process
Fabrication.
8. Measurement
Losses.
Implementation

Specifications

Application

Why the application is important?

- By selecting an application, you select the specifications linked with the application that you should meet (frequency range, input power levels, maximum acceptable return loss, maximum noise figure, standards, costs, technical/environmental constraints …)

How to select the specifications?

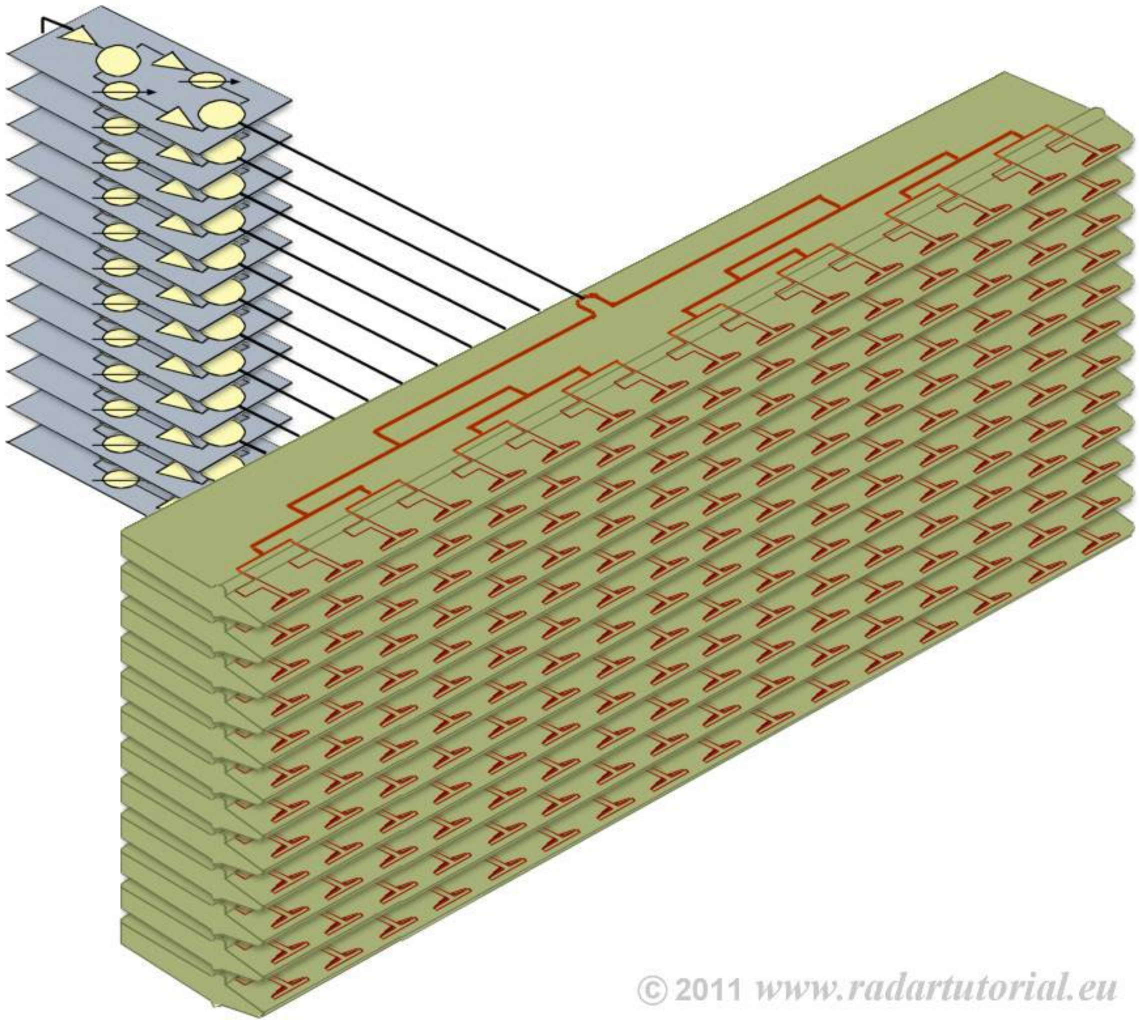
- Read papers which talk about power amplifiers (and in particular power amplifiers designed for the specific application you retained).
- Consult company sites (read their technical notes, white papers).
- Consult company products and technical application notes (if they sell devices that fit with your application: compare the specs).

JUSTIFY YOUR CHOICES via well-known technical references … !

Specifications

- Assess goals
- Set priorities
- Explore possible design configurations
- Consider design partitioning
- Allocate circuit specifications

Example: Design a Phased Array

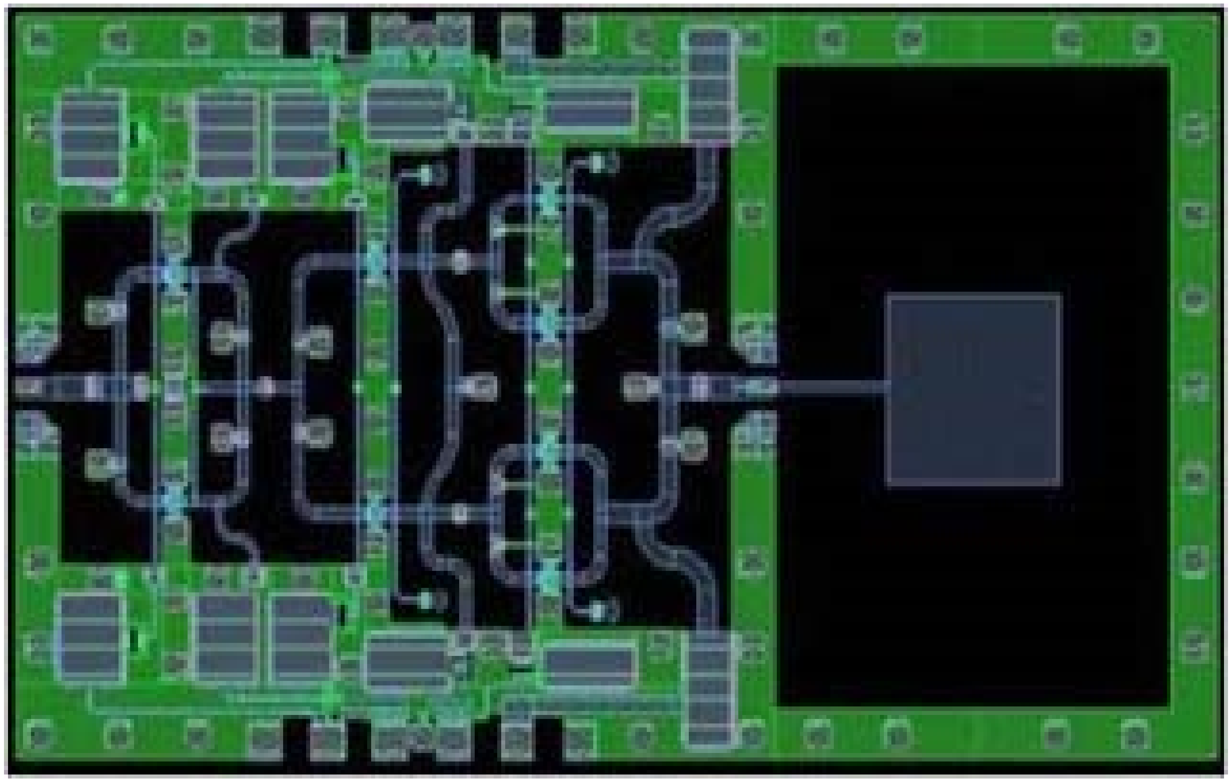


Applications (obsolete or in demand?):

Phased Array (when directive radiation is needed) which is usually used in military, Automotive or Weather Radar; Satellite telecommunication, Spacecraft Communication, Radio Frequency Identification (RFID) which usually work in X, Ku, Ka band or beyond.

Target (in demand?):

Integrating the different RF blocks of a Transmitter or a Trans-receiver including at least a Power Amplifier with an Antenna, regard to new scientific trends and industry requirements. The Antenna may radiate in free space or in a waveguide.



The image above shows an integrated millimetre wave power amplifier and patch antenna on a semiconductor substrate, developed for an experimental array-based wideband communications system.

***Advantages** of the proposed design ?*

- Achieving small-size and low-weight transmitter system.
- Low cost especially when many similar transmitters are required.
- Simplicity to provide power supply for transistor biasing thus, avoiding complicated wiring.
- Removing the RF connectors between the different blocks and decreasing the length of transmission line.
- Reducing loss as well as dispersive and parasitic effects
- Delivering the max available power to Antenna with min loss.
- Simplicity of disassembling a system for maintenance and repairing.

Worth to design !

***Challenges** of the proposed design ?*

- Creating on-chip Antenna usually results low efficiency (about 20%) therefore Gain of these Antenna usually is less than 1.
- There are limitations to implement many kinds of Antennas in chip technology and then to have a particular radiation pattern.
- Thermal conductivity, especially for high power amplifiers, is a considerable issue.
- We can’t use any component (such as Isolators and Circulators) which use ferrite materials.
- Cascading different blocks of a system make it potential to oscillate.
- Surface wave and Mutual Interaction between the components without enough housing probably affect their performance

***Can be addressed** ?*

Some resources: technical papers in well-known journals,

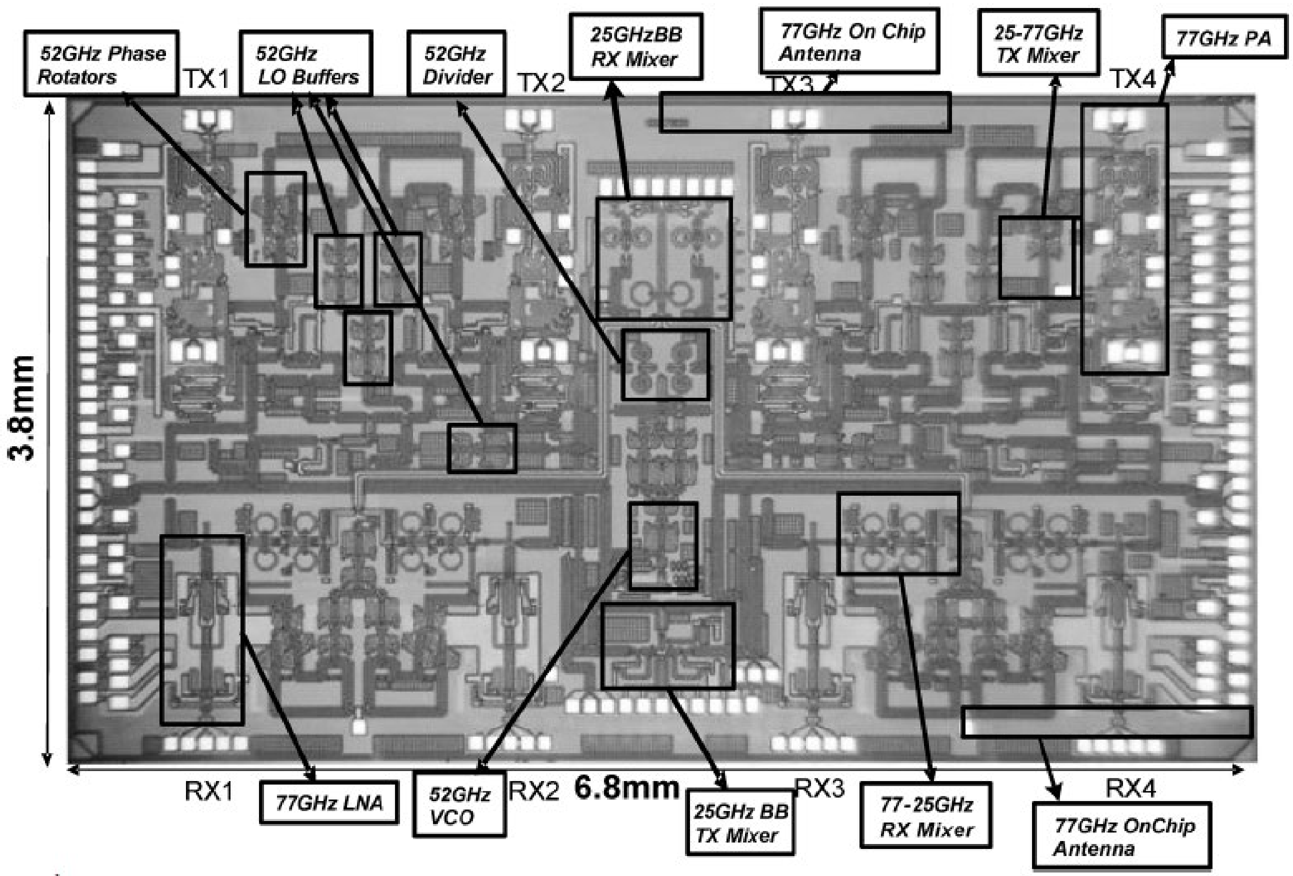
existing designs,

Ph.D. Theses (from top universities) …

IEEE JOURNAL OF SOLID-STATE CIRCUITS, VOL. 41, NO. 12, DECEMBER 2006 2795

A 77-GHz Phased-Array Transceiver With On-Chip Antennas in Silicon: Receiver and Antennas

Aydin Babakhani, *Student Member, IEEE*, Xiang Guan, *Member, IEEE*, Abbas Komijani, *Student Member, IEEE*, Arun Natarajan, *Student Member, IEEE*, and Ali Hajimiri, *Member, IEEE*



MICROWAVE INTEGRATED PHASED-ARRAY TRANSMITTERS IN SILICON

Thesis by

Abbas Komijani

In Partial Fulfillment of the Requirements

For the Degree of

Doctor of Philosophy

CALIFORNIA INSTITUTE OF TECHNOLOGY

Pasadena, California

2006

(Defended August 22, 2006)

Commercial devices: Hittite Productions

Hittite MICROWAVE CORPORATION v00.1112

RoHS✓

HMC6000LP711E

MILLIMETERWAVE TRANSMITTER 57 - 64 GHz

Typical Applications

The HMC6000LP711E is ideal for:

- WiGig Single Carrier Modulations
- 60 GHz ISM Band Data Transmitter
- Multi-Gbps Data Communications
- High Definition Video Transmission
- RFID

General Description

The HMC6000LP711E is a complete mmWave transmitter IC and low profile antenna integrated in a plastic 7x11 mm surface mount package. The transmitter provides 23.5 dBm of EIRP operating over 57 - 64 GHz with 1.8 GHz of modulation bandwidth. An integrated synthesizer provides tuning in 500 or 540 MHz step sizes depending on the choice of external reference clock. Support for a wide variety of modulation formats is provided through a universal analog baseband IQ interface. Together with the HMC6001LP711E, a complete transmit/receive chipset is provided for multi-Gbps operation in the unlicensed 60 GHz ISM band.

Features

Support for IEEE Channel Plan

EIRP: 23.5 dBm

Output Power: 16 dBm

Antenna Gain: 7.5 dBi

Max Gain: 38 dB

Gain Control Range: 17 dB

Integrated Frequency Synth

Integrated Image Reject Filt

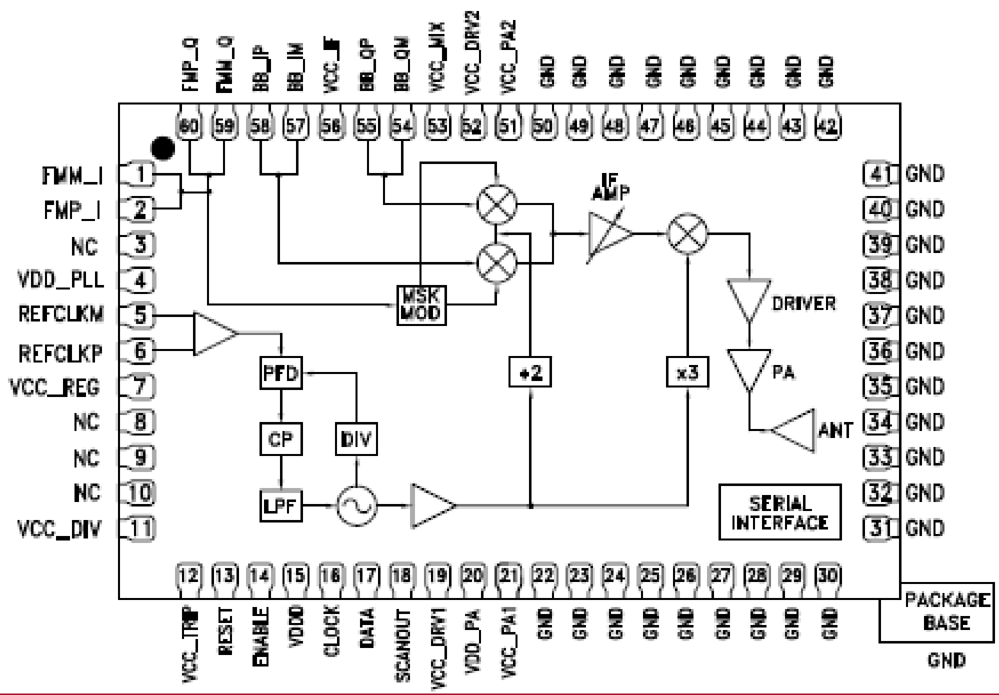
Programmable IF Gain Bloc

Universal Analog I/Q Baseb

Three-Wire Serial Digital Int

7x11mm QFN Package: 77m

Functional Diagram



For price, delivery and to place orders: Hittite Microwave Corporation, 2 Elizabeth Drive, Chelmsford, MA 01824
Phone: 978-250-3343 Fax: 978-250-3373 Order On-line at www.hittite.com
Application Support: Phone: 978-250-3343 or txrx@hittite.com

But before the circuit level …

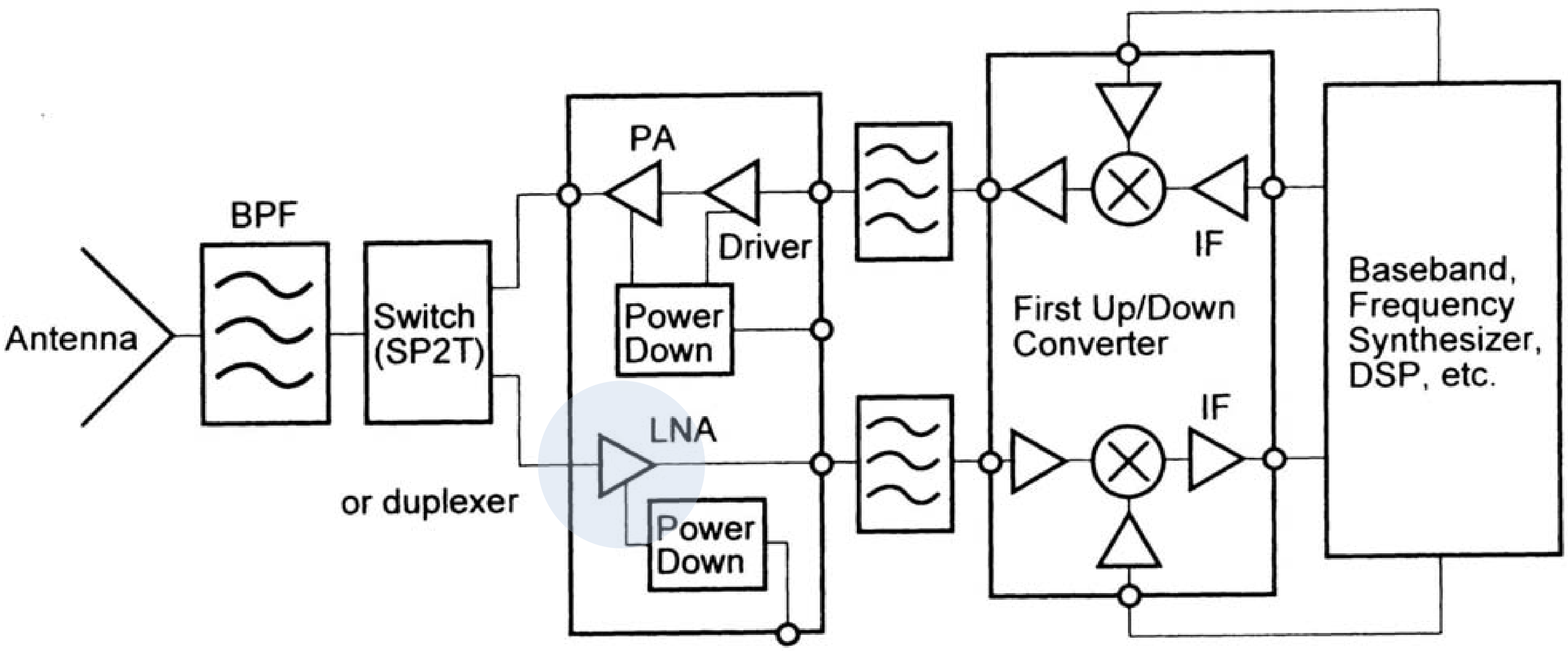
Better to start at the system level (for a larger picture)

… and because designers often have the specs for the system as a whole.

So they have to extract the specs for each element of the system !

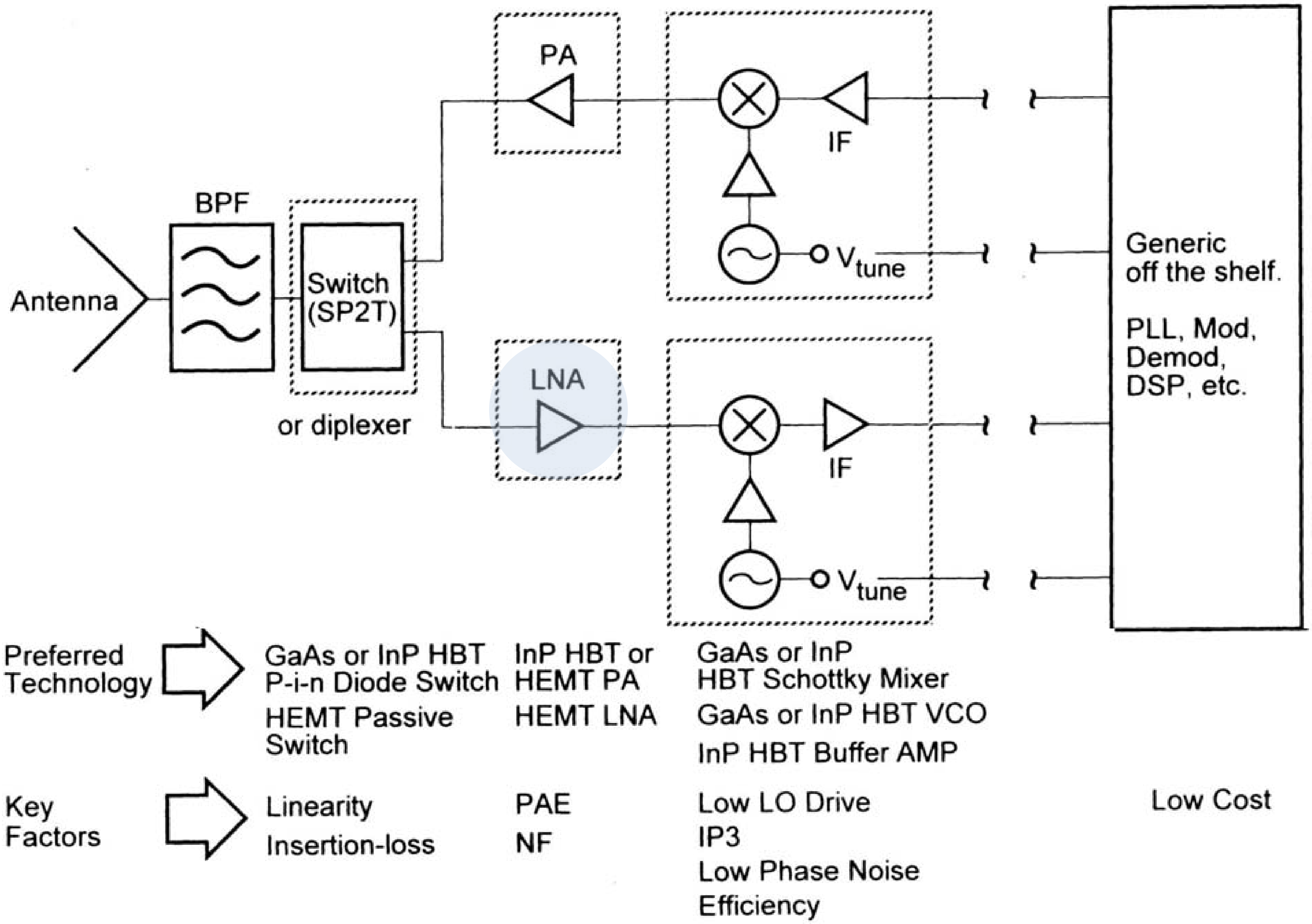
First step: the specifications

Start at the system level (transmitter) - an example -



Key Factors ⇨	Low dc power	Performance Cost Single-supply idle current	Performance Low voltage operation	Low-Cost
Preferred Technology ⇨	GaAs MESFET	GaAs or InP HBT GaAs MESFET **SiGe HBT**	GaAs or InP HBT GaAs MESFET Si-BJT Bi-CMOS, MOS	Low-Cost Silicon Technology

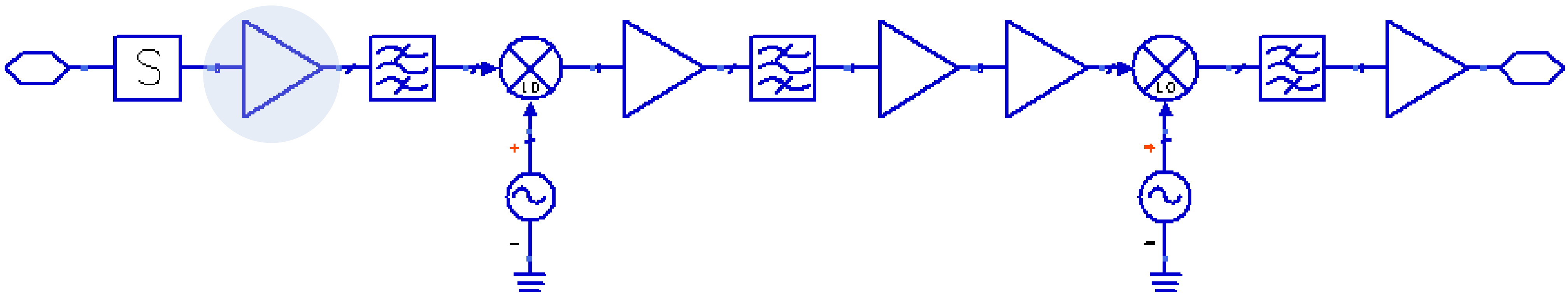
Start at the system level (transmitter) - another example -



Specifications (System Level)

- **Operating Frequency** and **Bandwidth**
- **Output Power** of Power Amplifier or **EIRP** of Antenna
- **Gain**, **Polarization** and **Other Relative Parameters** of Antenna
- **Preferable Technology**, GaN, GaAs or CMOS also using **single** Technology or **combination** of multiple Technologies (e.g., GaN with MEMS)
- The elements that should be **designed** (or **purchased**): Antenna, Filters, High Power Amplifier, Amplifier Drivers, Mixers, VCO or may be other receiver sections, …

System level : Budget Analysis



Desired specs for the whole receiver:

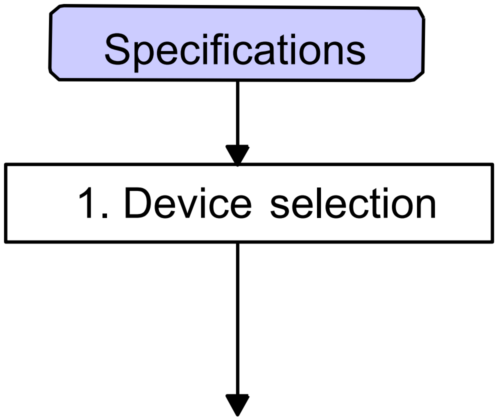
Gain : 102.1 dB (targeting 95 dB min)

NF: 52.8 dB (targeting 60 dB max)

Output power: -19.9 dBm max (targeting -20 dBm max)

Desired specs for the LNA?

Element	Duplexer	LNA	BPF	Mixer	AMP #1	BPF	AMP #2	AMP #3	Mixer	BPF	AMP
Gain (dB)	-3	20	-2.7	-7	20	-17.4	30	30	-6	-6	45
Noise Figure (dB)	2.7	1.5	2.7	7.5	4	17.4	4	6	6	6	5
NFO (dB)	2.7	4.2	4.3	4.7	5.1	5.3	5.6	5.6	5.6	5.6	5.6
Power (dBm)	-125.3	-103.9	-107	-114.4	-94.1	-111.5	-81.5	-53.2	-58.9	-65	-19.9
SNR (dB)	6.5	4.9	4.9	4.5	4	3.9	3.6	3.6	3.6	3.6	3.6
TOI @ Output (dBm)		3.2	0.5	-7.5	8	-9.4	9.1	24.2	15.8	9.8	20



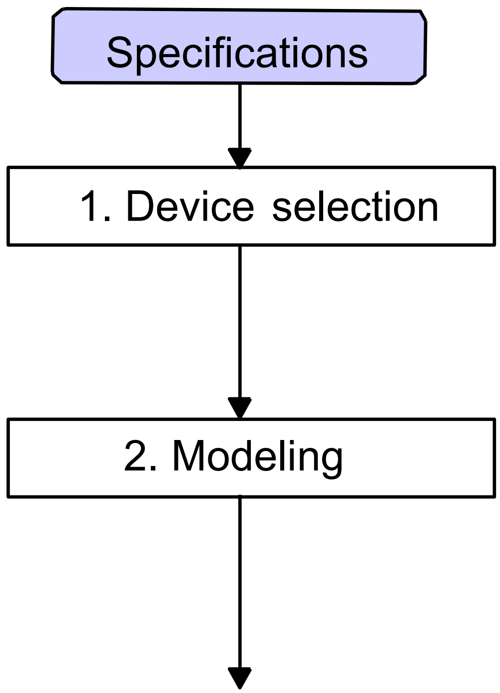
- Based on the application retained, select the most suitable active device (usually a transistor) that can fulfill the defined requirements.

  - Type of transistor (BHT, FET, HEMT, …)
  - Technology (GaAs, Si, GaN …)

How?

- Use datasheets, technical papers ...

JUSTIFY YOUR CHOICE!


Specifications
1. Device selection
2. Modeling

Investment in Models

Dedicate modeling team

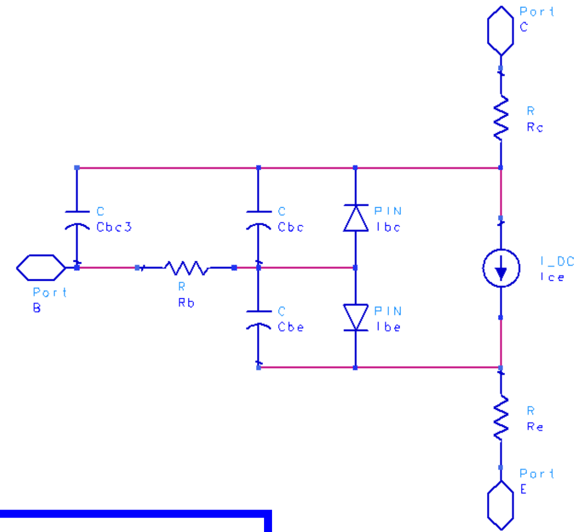
Use available parts libraries

Measure individual parts

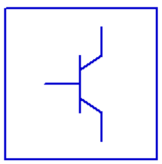
Consider different operating conditions

Improve existing models

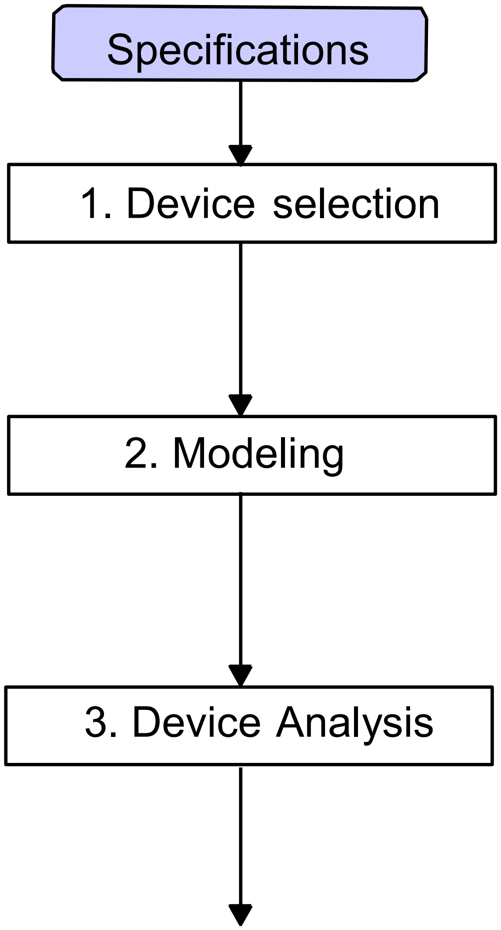
- At the equation level
- At the software level
- At the implemented equivalent circuit level



$$Ibe=(IBbif(\exp(Vbe/NbfVT)-1.0))+Ise(\exp(Vbe/(NexVt))-1.0)$$



!Freq.[Hz]	MagS11[dB]	PhaseS11[DEG]	MagS21[dB]	PhaseS21[DEG]	MagS12[dB]	PhaseS12[DEG]		
300000	-5.986E-07	-1.151E-02	-7.394E+01	8.997E+01	-7.394E+01	8.997E+01	-5.986E-07	-1.151E-02
315229	-6.384E-07	-1.210E-02	-7.351E+01	8.997E+01	-7.351E+01	8.997E+01	-6.384E-07	-1.210E-02
331231	-6.812E-07	-1.271E-02	-7.308E+01	8.997E+01	-7.308E+01	8.997E+01	-6.812E-07	-1.271E-02
348046	-7.273E-07	-1.336E-02	-7.265E+01	8.997E+01	-7.265E+01	8.997E+01	-7.273E-07	-1.336E-02
365714	-7.769E-07	-1.403E-02	-7.222E+01	8.997E+01	-7.222E+01	8.997E+01	-7.769E-07	-1.403E-02
384279	-8.303E-07	-1.475E-02	-7.179E+01	8.997E+01	-7.179E+01	8.997E+01	-8.303E-07	-1.475E-02
403787	-8.879E-07	-1.550E-02	-7.136E+01	8.997E+01	-7.136E+01	8.997E+01	-8.879E-07	-1.550E-02
424285	-9.501E-07	-1.628E-02	-7.093E+01	8.997E+01	-7.093E+01	8.997E+01	-9.501E-07	-1.628E-02
445823	-1.017E-06	-1.711E-02	-7.050E+01	8.997E+01	-7.050E+01	8.997E+01	-1.017E-06	-1.711E-02
468455	-1.090E-06	-1.798E-02	-7.007E+01	8.997E+01	-7.007E+01	8.997E+01	-1.090E-06	-1.798E-02
492235	-1.168E-06	-1.889E-02	-6.964E+01	8.997E+01	-6.964E+01	8.997E+01	-1.168E-06	-1.889E-02
517223	-1.252E-06	-1.985E-02	-6.921E+01	8.997E+01	-6.921E+01	8.997E+01	-1.252E-06	-1.985E-02
543479	-1.344E-06	-2.086E-02	-6.878E+01	8.997E+01	-6.878E+01	8.997E+01	-1.344E-06	-2.086E-02

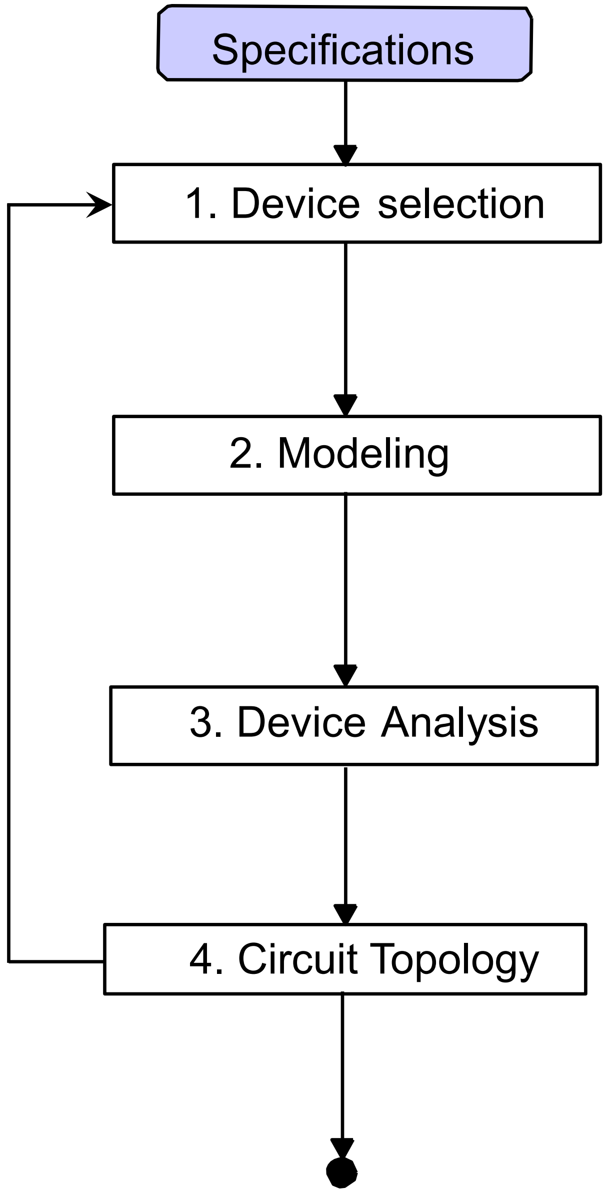


Verify the performance of the retained device

WHY?

- Implement the device into a circuit simulator. Plot the I-V curves. Verify if they match with the expected I-V curves (from a trusted source like datasheets).
- Compile the small-signal S-parameters. Is the small-signal performance as expected?
- Compile the large-signal parameters such as large-signal gain. Are the large-signal response as expected?
- Verify its nonlinear model (e.g., the drain current model to use)
- When possible, measure the transistor characteristics (S-parameters, load pull contours …) and compare with datasheets.

JUSTIFY YOUR CHOICE!



Based on the application and related specifications, select the most suitable configuration that can fulfill the requirements.

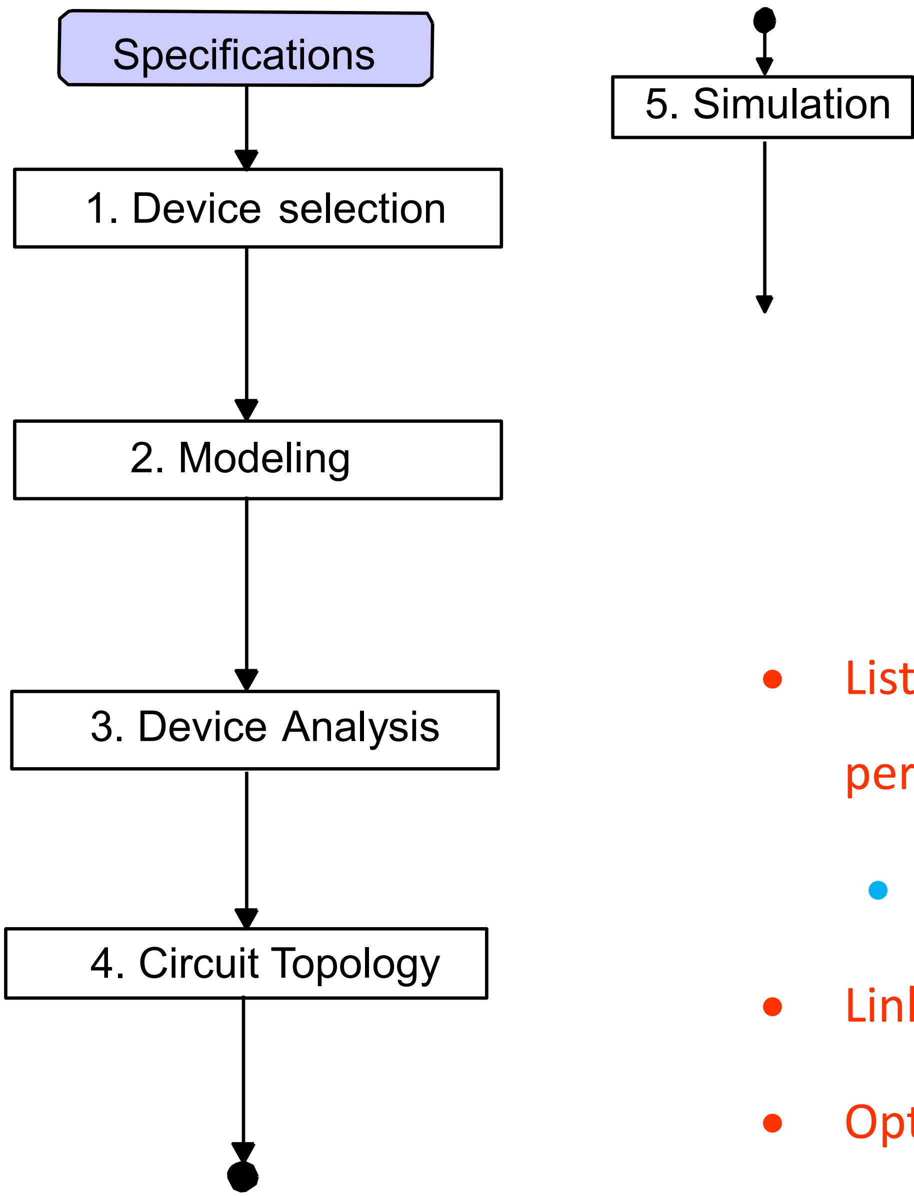
How?

- Review all existing configurations proposed for the targeted application.
- List the parameters the designers prioritized.
- Then, from that list, select the configuration type that should help reaching the desired objectives (for instance if gain AND bandwidth are both important, see which configuration can best fit with these two design constraints).
  - Cascode, Cascade, distributed, differential …
  - Common drain, common source …
- If needed (not able to meet specs), change the transistor !

JUSTIFY YOUR CHOICE!

The most important : **Why and How**!

- Check the specifications:
  - Desired ranges of frequency, gain, noise figure, and ***how*** to meet them.
- Select the key parameters to consider based on the application:
  - What are the most sensitive design parameters? ***Why?***
- From that … what are the possible configurations you may select?
  - List their advantages/limitations. ***Justify***.
- Why you selected this particular configuration?
  - How it fits with the specifications? ***Justify***.



- List the different simulations you have to perform and why you performed them.
  - Rationale behind each simulation.
- Link with the specifications.
- Optimization round?

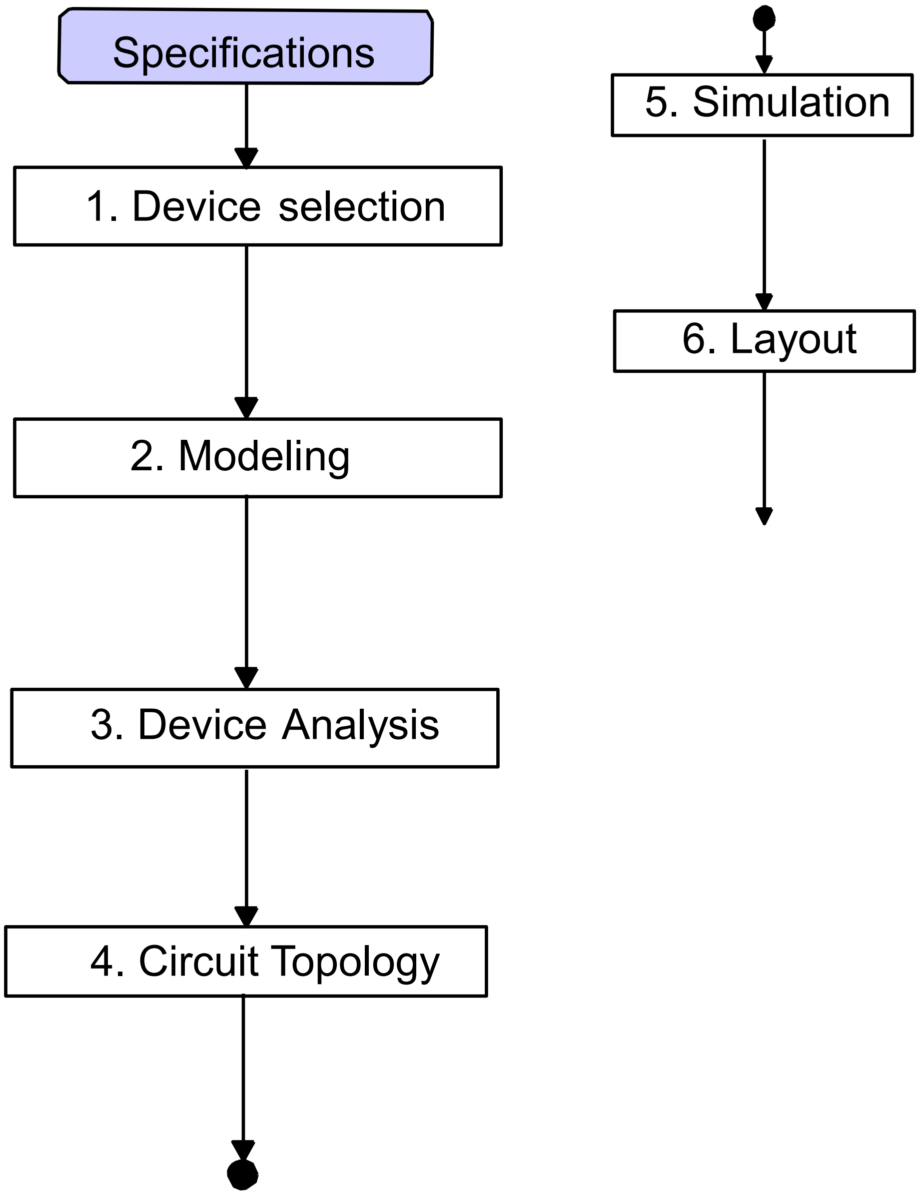
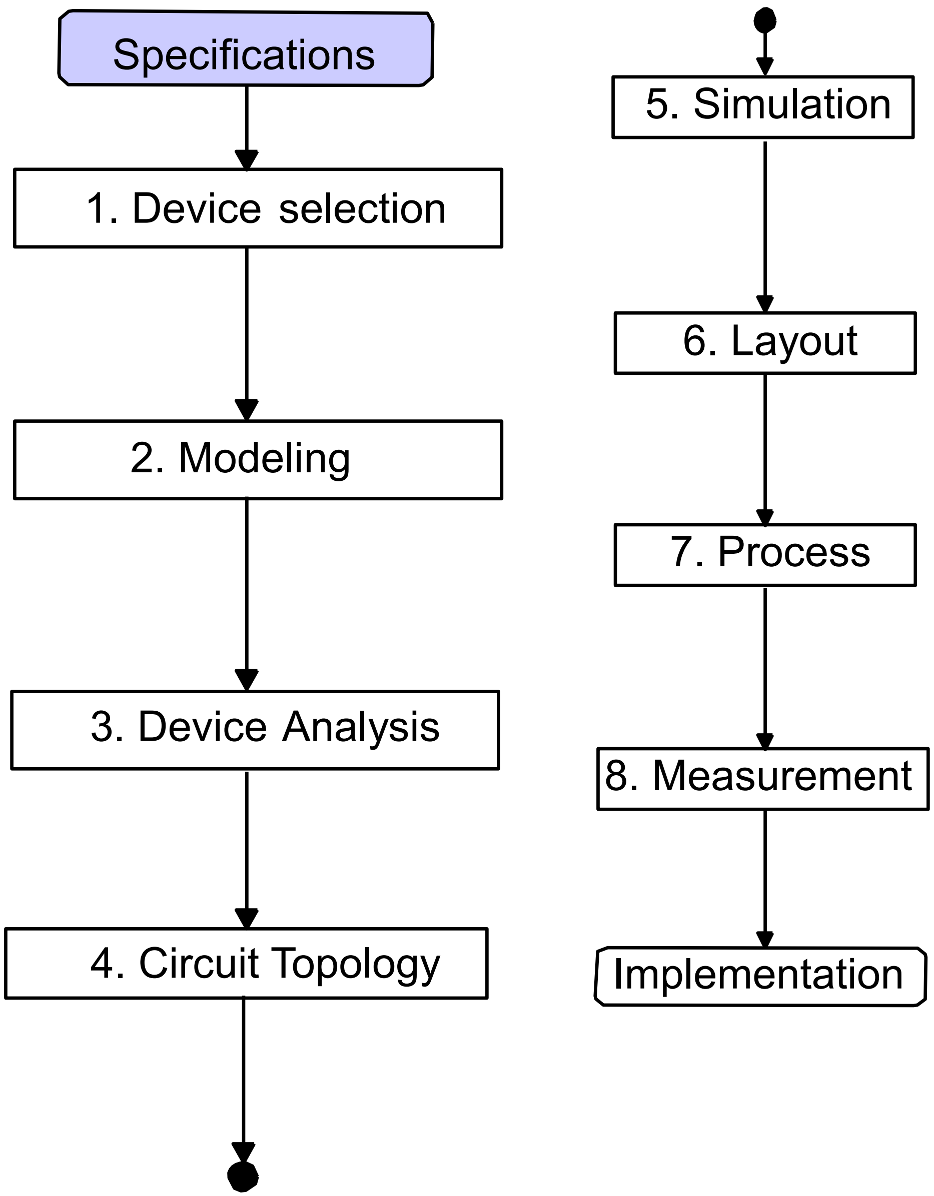

Specifications
1. Device selection
2. Modeling
3. Device Analysis
4. Circuit Topology
5. Simulation
6. Layout

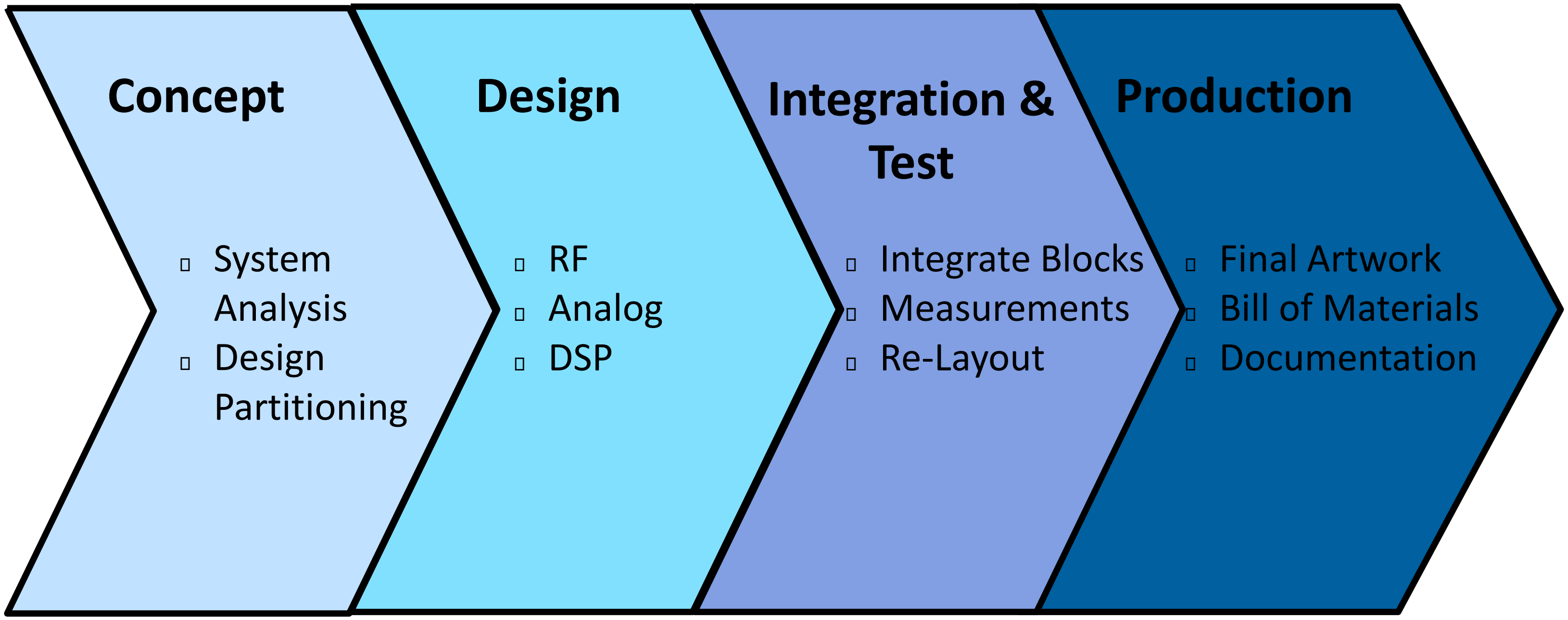
TABLE 3.1. Comparison of Monolithic Integrated-Circuit Substrates: Pure Materials at Room Temperature

Property	Silicon	SiC	GaAs	InP	GaN
Semi-insulating	No	Yes	Yes	Yes	Yes
Resistivity (Ω cm)	10^3-10^5	$>10^{10}$	10^7-10^9	$\sim 10^7$	$>10^{10}$
Dielectric constant	11.7	40	12.9	14	8.9
Electron mobility (cm^2/V sec)	1450	500	8500	6000	800
Saturation electrical velocity (cm/sec)	9×10^6	2×10^7	1.3×10^7	1.9×10^7	2.3×10^7
Radiation hardness	Poor	Excellent	Very good	Good	Excellent
Density (g/cm^3)	2.3	3.1	5.3	4.8	—
Thermal conductivity (W/cm °C)	1.45	4.3	0.46	0.68	1.3
Operating temperature (°C)	250	>500	350	300	>500
Energy gap (eV)	1.12	2.86	1.42	1.34	3.39
Breakdown field (kV/cm)	≈300	≥2000	400	500	≥5000

RF and Microwave Circuit and component Design for Wireless Systems; K. Chang, I. Bahl and V. Nair


Specifications
1. Device selection
2. Modeling
3. Device Analysis
4. Circuit Topology
5. Simulation
6. Layout
7. Process
8. Measurement
Implementation

The Complete Design Process



- Integrated Simulators
- Faster Simulators
- Optimizers
- Instrument I/O
- Parts Libraries

- Co-Simulation
- System Simulation

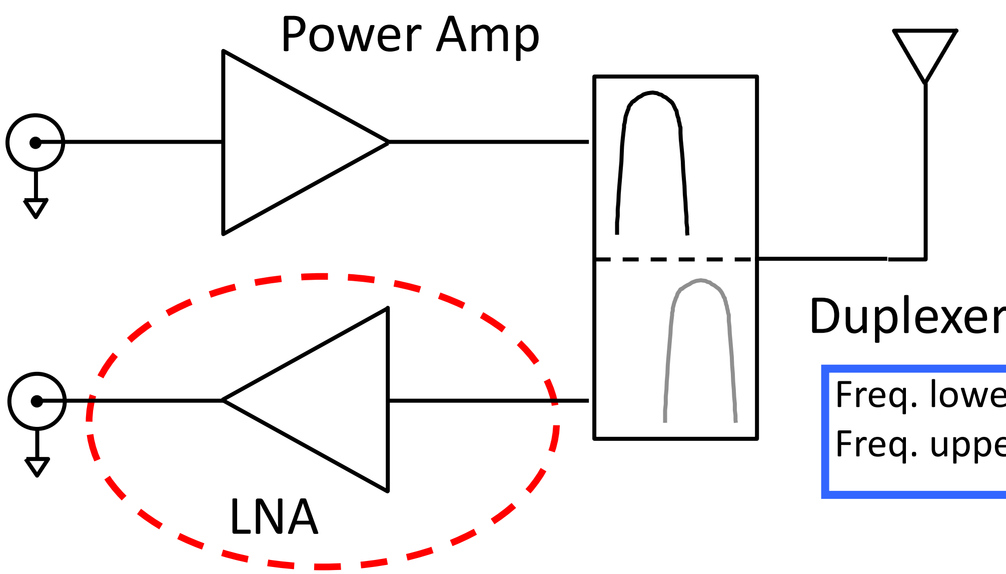
- Layout
- EM Simulation
- Parts Libraries

- Artwork Generation

Low-Noise Amplifier Design

Concept: System-Level Design

Freq. = 1880 MHz +/- 50 MHz min.
Pout (1 dB) = +25 dBm min
Pout = +27 dBm min
Gain = + 24 dB min



Freq. lower band = 1880 MHz +/- 30 MHz
Freq. upper band = 1960 MHz +/- 30 MHz

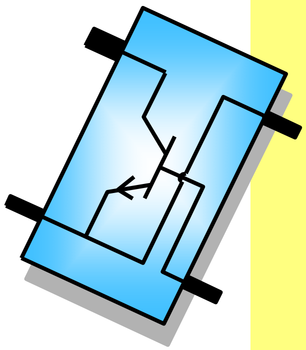
Freq. = 1960 MHz +/- 50 MHz min
Gain = + 25 dB min
NF < 2.5 dB

What Are My Resources?

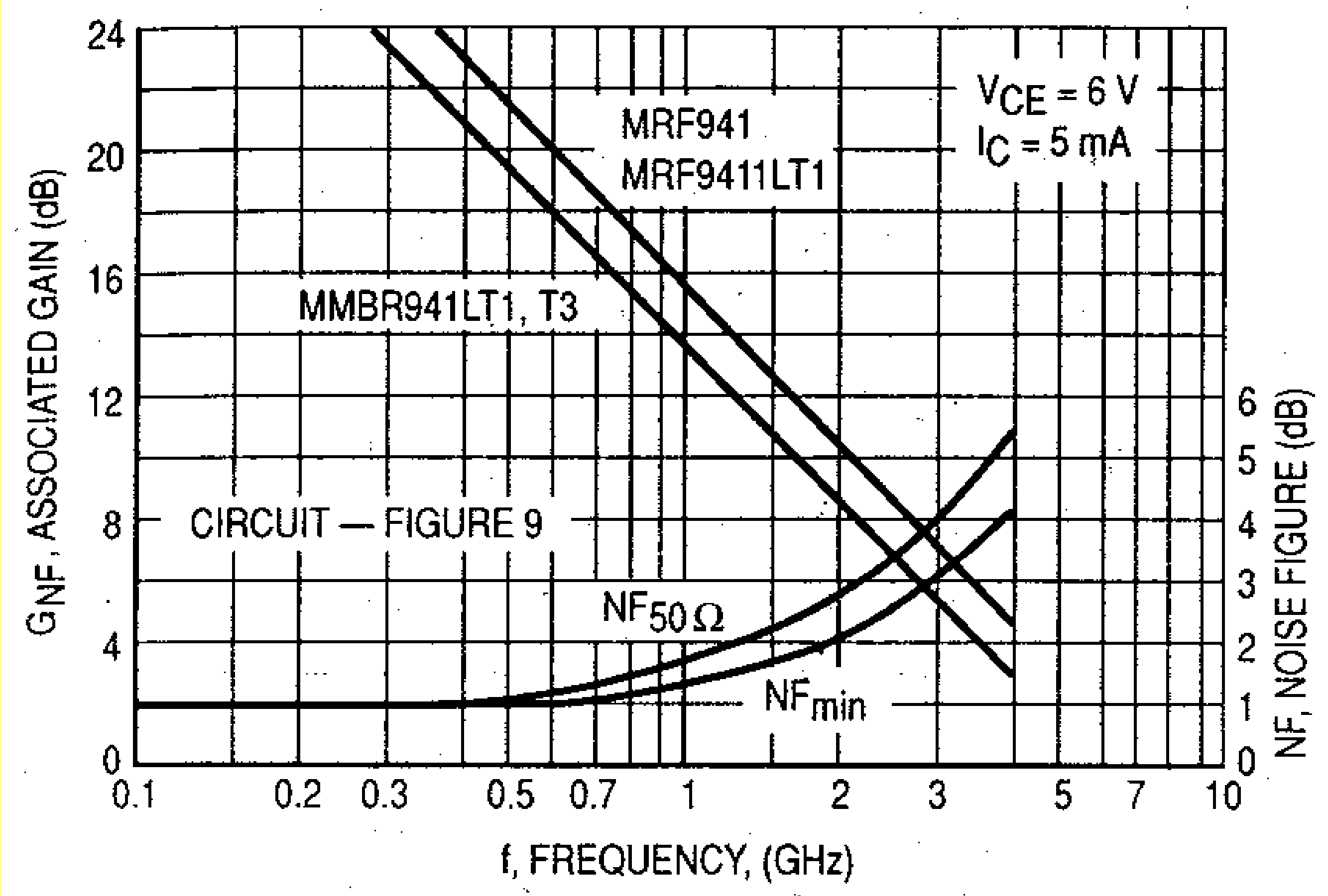
How do I design a low-noise amplifier?

- Talk with experienced people
- Read Textbooks
- Explore Magazines
- Review Commercial Applications Notes
- Purchase from third-party?

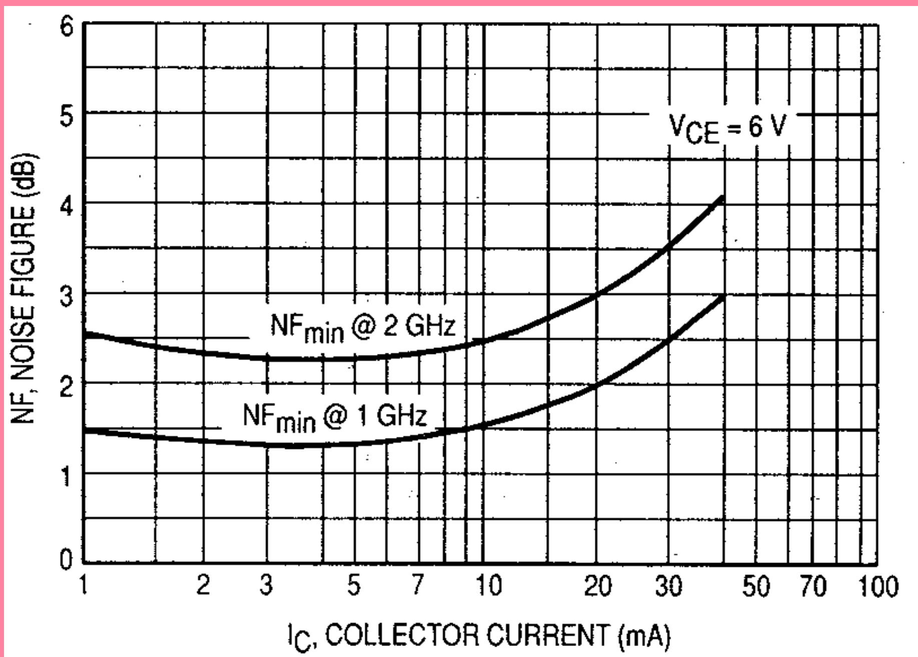
Choosing a Device

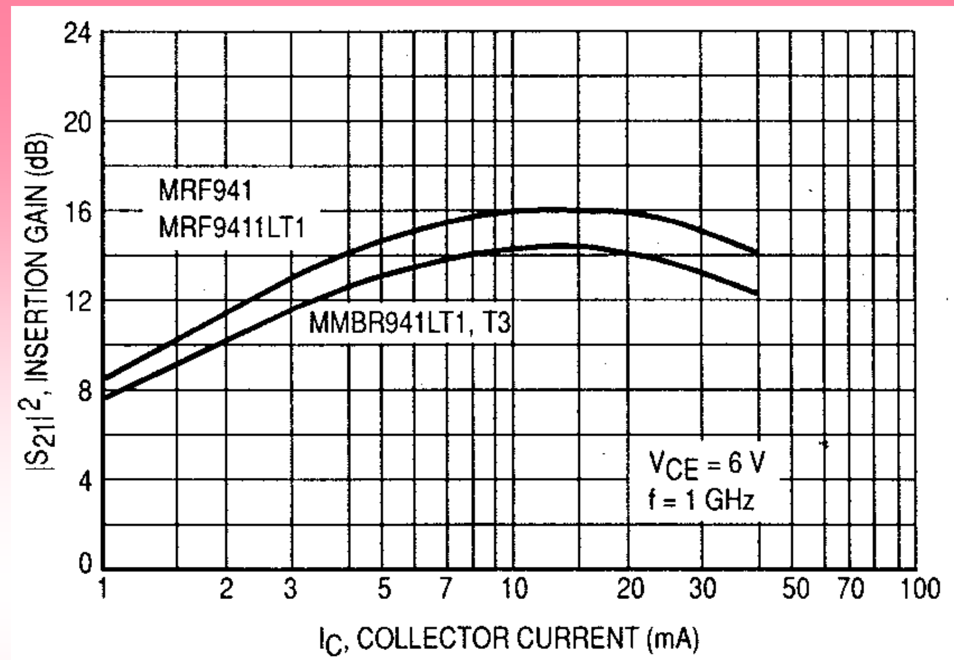


A typical low-noise device



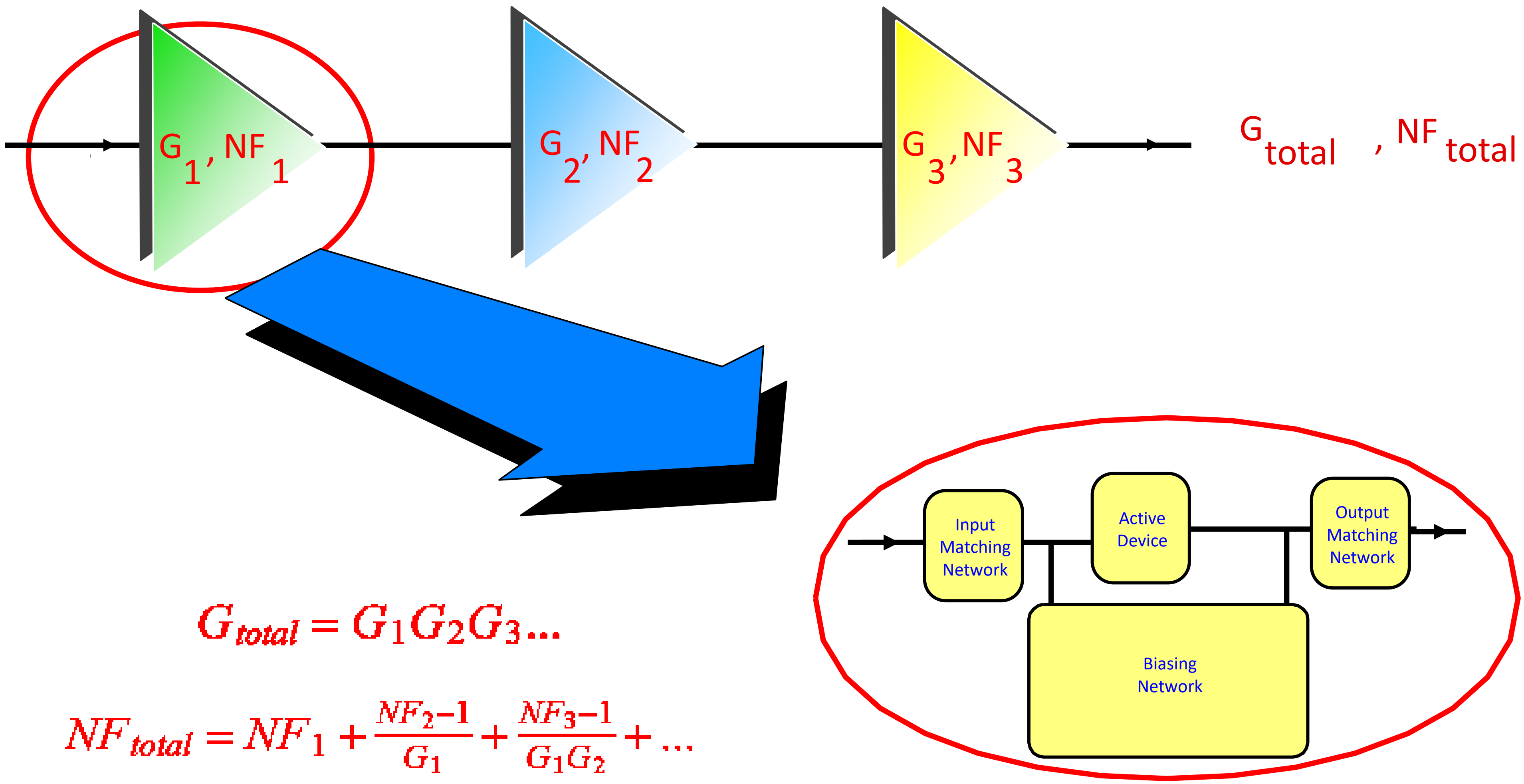
Transistor Biasing





- Decide on bias currents and voltages
- Find values for biasing resistors
- Verify values using DC analysis with the nonlinear model
- Design from available power supplies
- Check power dissipation

Amplifier Stage Design: How many stages?



$$G_{total} = G_1 G_2 G_3 \ldots$$

$$NF_{total} = NF_1 + \frac{NF_2 - 1}{G_1} + \frac{NF_3 - 1}{G_1 G_2} + \ldots$$

STABILITY

Stability and Simultaneous Conjugate Match

$|K| > 1$

$K > 1$

Simultaneous conjugate match can be achieved

$|\Delta| > 1$

Potentially unstable

$|\Delta| < 1$

Unconditionally stable

$$\Gamma_{Ms} = \frac{B_1 - \sqrt{B_1^2 - 4|C_1|^2}}{2C_1} \qquad \Gamma_{ML} = \frac{B_2 - \sqrt{B_2^2 - 4|C_2|^2}}{2C_2}$$

$K < -1$

Simultaneous conjugate match doesn't exist

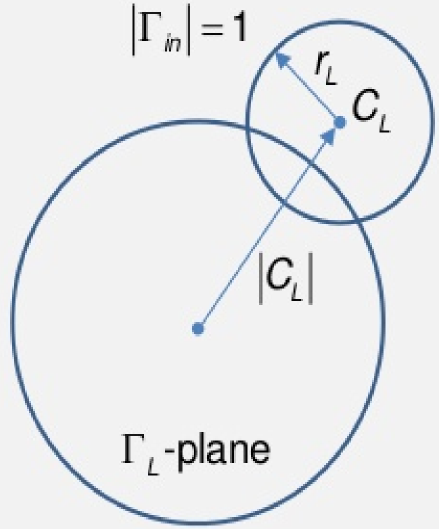
$|K| < 1$

Potentially unstable or Unstable

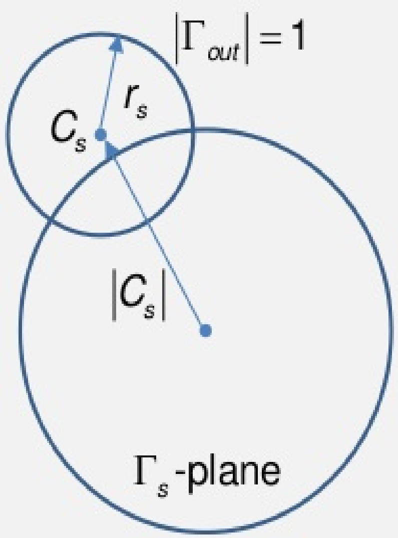
Any reference to a simultaneous conjugate match assumes that the two port network is unconditionally stable.

The Stability Circles

- **Output Stability Circle** (Γ_L values for $|\Gamma_{in}| = 1$)
  - Center $C_L = \dfrac{(S_{22} - \Delta S_{11}^*)^*}{|S_{22}|^2 - |\Delta|^2}$
  - Radius $r_L = \left| \dfrac{S_{12}S_{21}}{|S_{22}|^2 - |\Delta|^2} \right|$

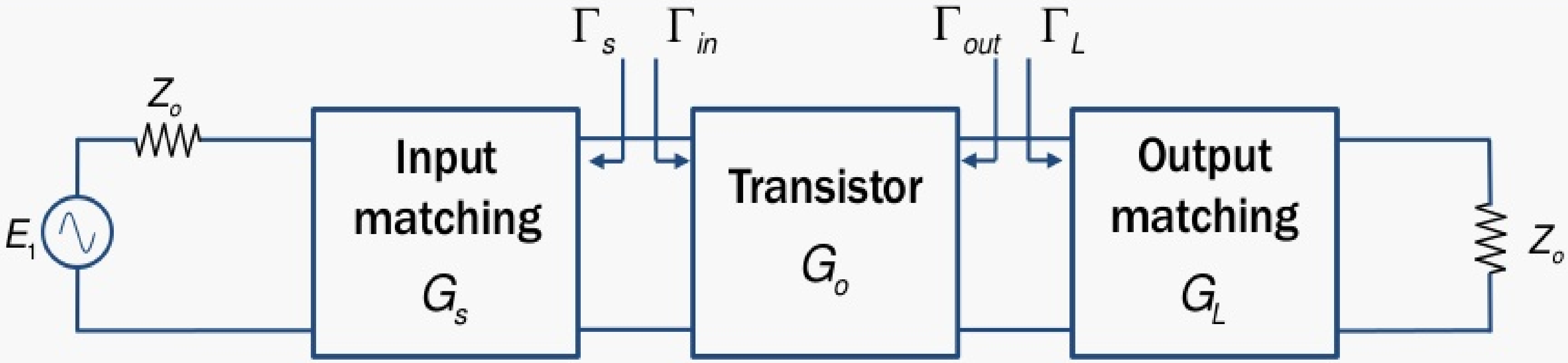


- **Input Stability Circle** (Γ_s values for $|\Gamma_{out}| = 1$)
  - Center $C_s = \dfrac{(S_{11} - \Delta S_{22}^*)^*}{|S_{11}|^2 - |\Delta|^2}$
  - Radius $r_s = \left| \dfrac{S_{12}S_{21}}{|S_{11}|^2 - |\Delta|^2} \right|$



GAIN AND NOISE FIGURE

Maximum Stable and Available Gain



- **Maximum Simultaneous Conjugate Matched Transducer Power Gain $G_{T,max}$**

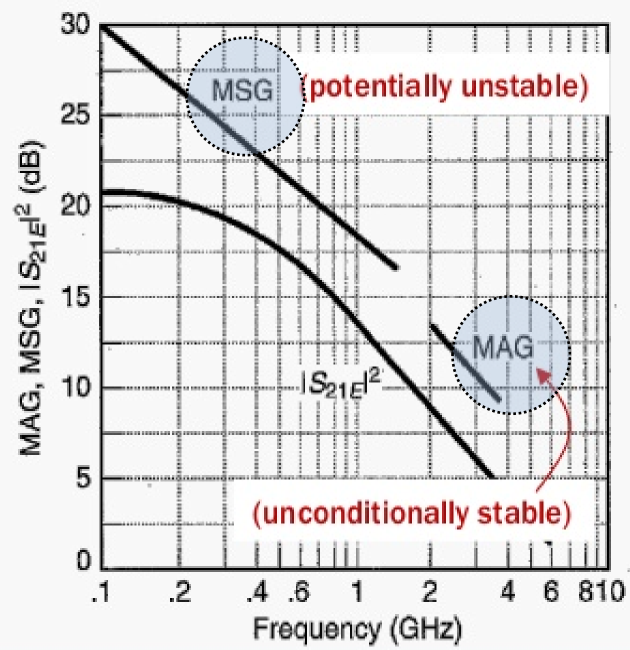
$$\Gamma_s = \Gamma_{in}^* = \Gamma_{Ms} \quad \text{and} \quad \Gamma_L = \Gamma_{out}^* = \Gamma_{ML}$$

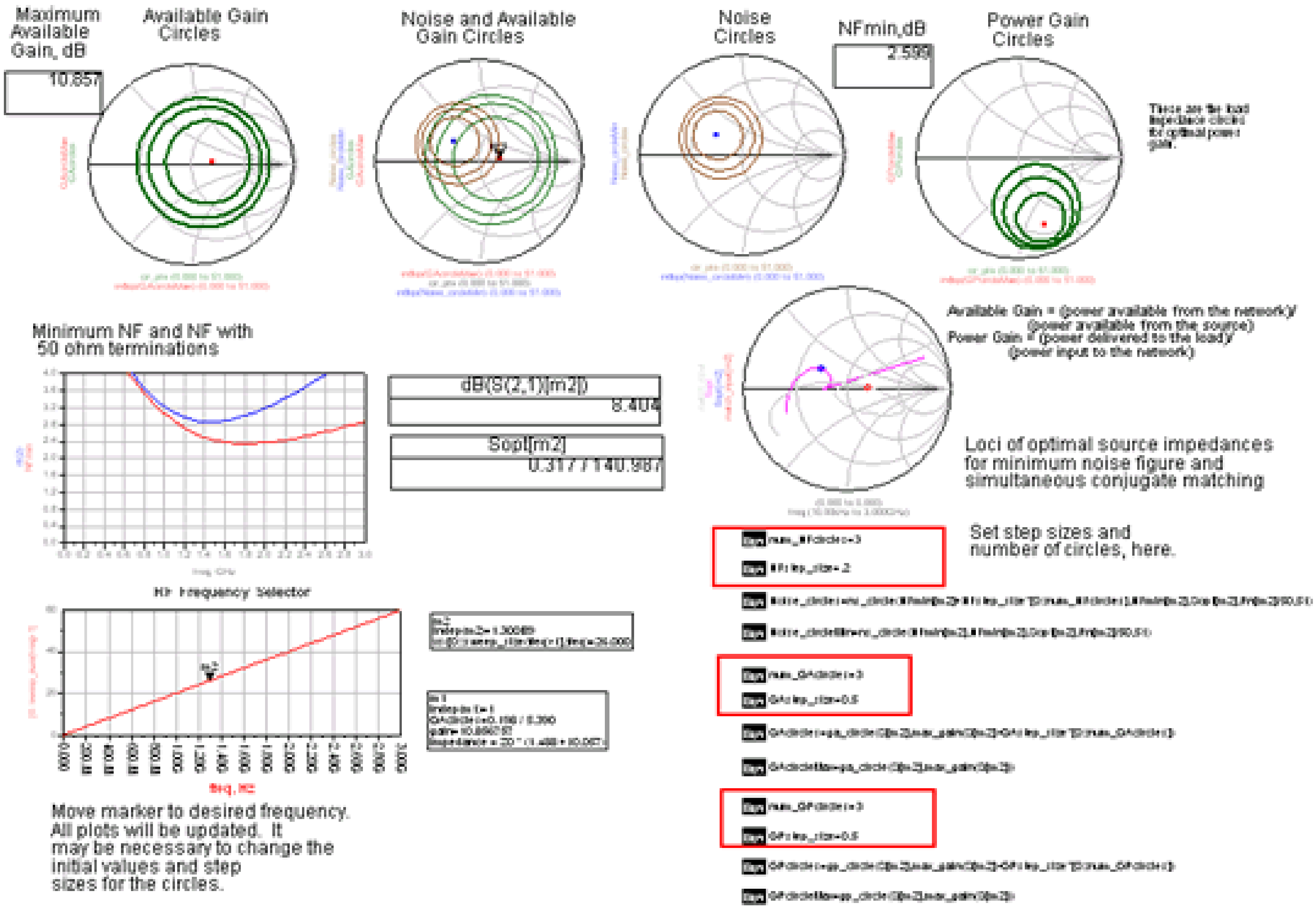
$$G_T = \frac{1-|\Gamma_s|^2}{|1-\Gamma_{in}\Gamma_s|^2}|S_{21}|^2\frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2}$$

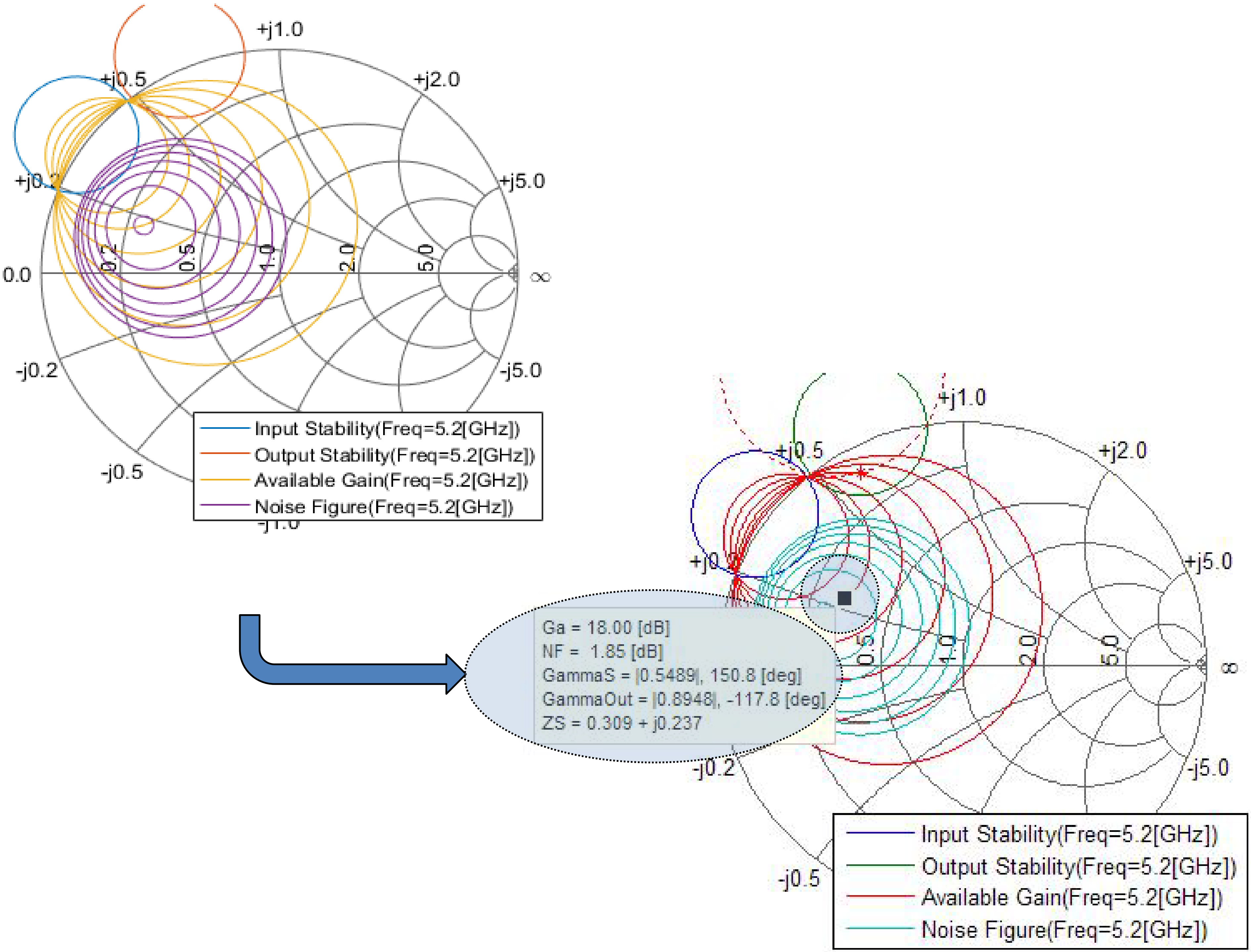
$$\Rightarrow G_{T,\max} = \frac{1}{1-|\Gamma_{Ms}|^2}|S_{21}|^2\frac{1-|\Gamma_{ML}|^2}{|1-S_{22}\Gamma_{ML}|^2} = \frac{|S_{21}|}{|S_{12}|}\left(K-\sqrt{K^2-1}\right)$$

- **Maximum Stable Gain (MSG) is defined when $K=1$:**

$$G_{MSG} = \frac{|S_{21}|}{|S_{12}|}$$




Maximum Available Gain, dB
10.857
Available Gain Circles
Noise and Available Gain Circles
Noise Circles
NFmin,dB
Power Gain Circles
These are the load impedance circles for optimal power gain.
Available Gain = (power available from the network)/(power available from the source)
Power Gain = (power delivered to the load)/(power input to the network)
Minimum NF and NF with 50 ohm terminations
dB(S(2,1)[m2])
8.404
Sopt[m2]
0.317 / 140.987
Loci of optimal source impedances for minimum noise figure and simultaneous conjugate matching
Set step sizes and number of circles, here.
Move marker to desired frequency.
All plots will be updated. It may be necessary to change the initial values and step sizes for the circles.

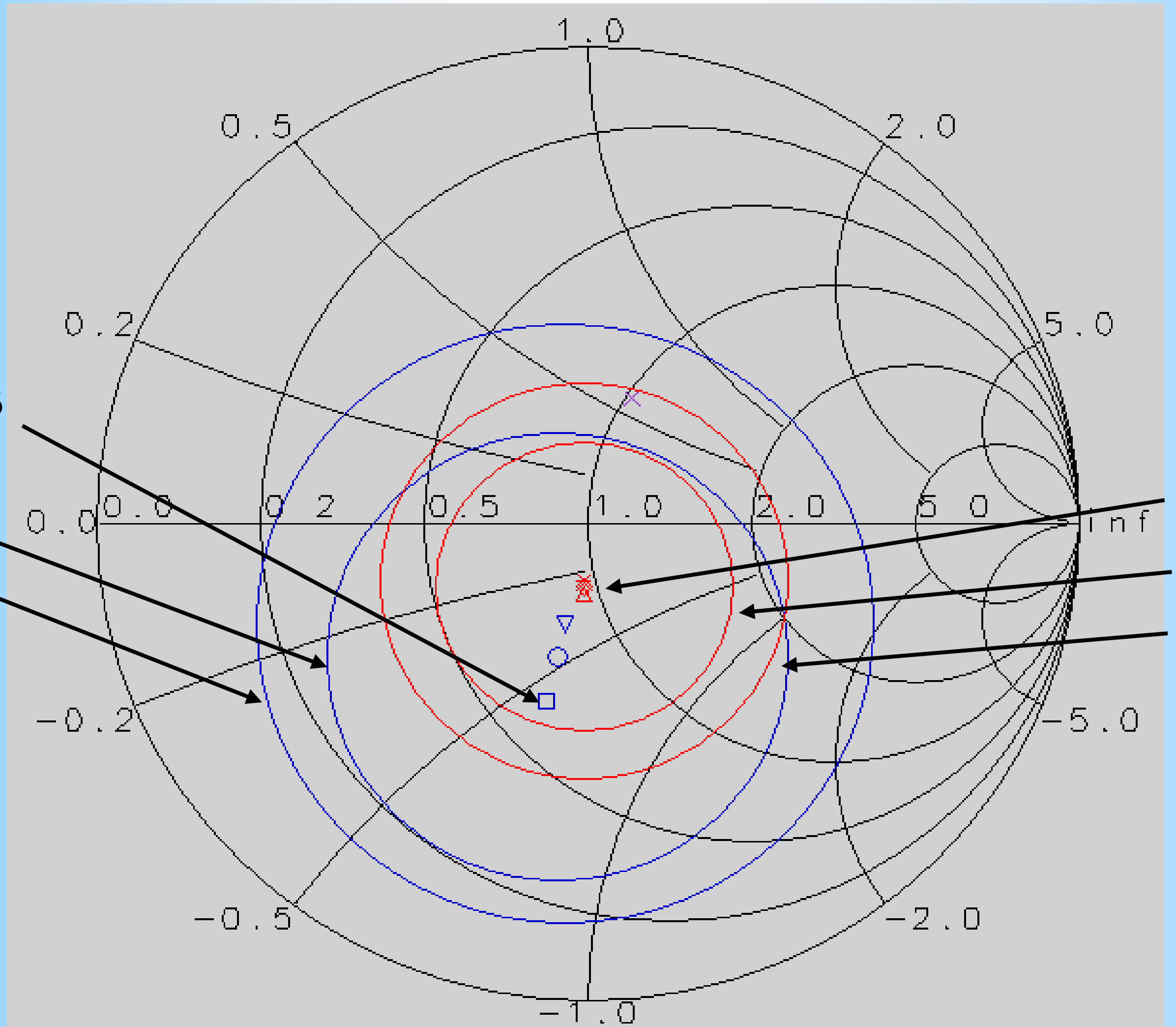

+j1.0
+j0.5
+j2.0
+j0.2
+j5.0
0.0
0.2
0.5
1.0
2.0
5.0
∞
-j0.2
-j5.0
-j0.5
-j1.0
Input Stability(Freq=5.2[GHz])
Output Stability(Freq=5.2[GHz])
Available Gain(Freq=5.2[GHz])
Noise Figure(Freq=5.2[GHz])
Ga = 18.00 [dB]
NF = 1.85 [dB]
GammaS = |0.5489|, 150.8 [deg]
GammaOut = |0.8948|, -117.8 [deg]
ZS = 0.309 + j0.237
+j1.0
+j0.5
+j2.0
+j5.0
0.5
1.0
2.0
5.0
∞
-j0.2
-j5.0
-j0.5
Input Stability(Freq=5.2[GHz])
Output Stability(Freq=5.2[GHz])
Available Gain(Freq=5.2[GHz])
Noise Figure(Freq=5.2[GHz])

MATCHING

Input Matching Network

Gain Circles:

Max. Gain: 8.6 dB
-1.0 dB
-2.0 dB

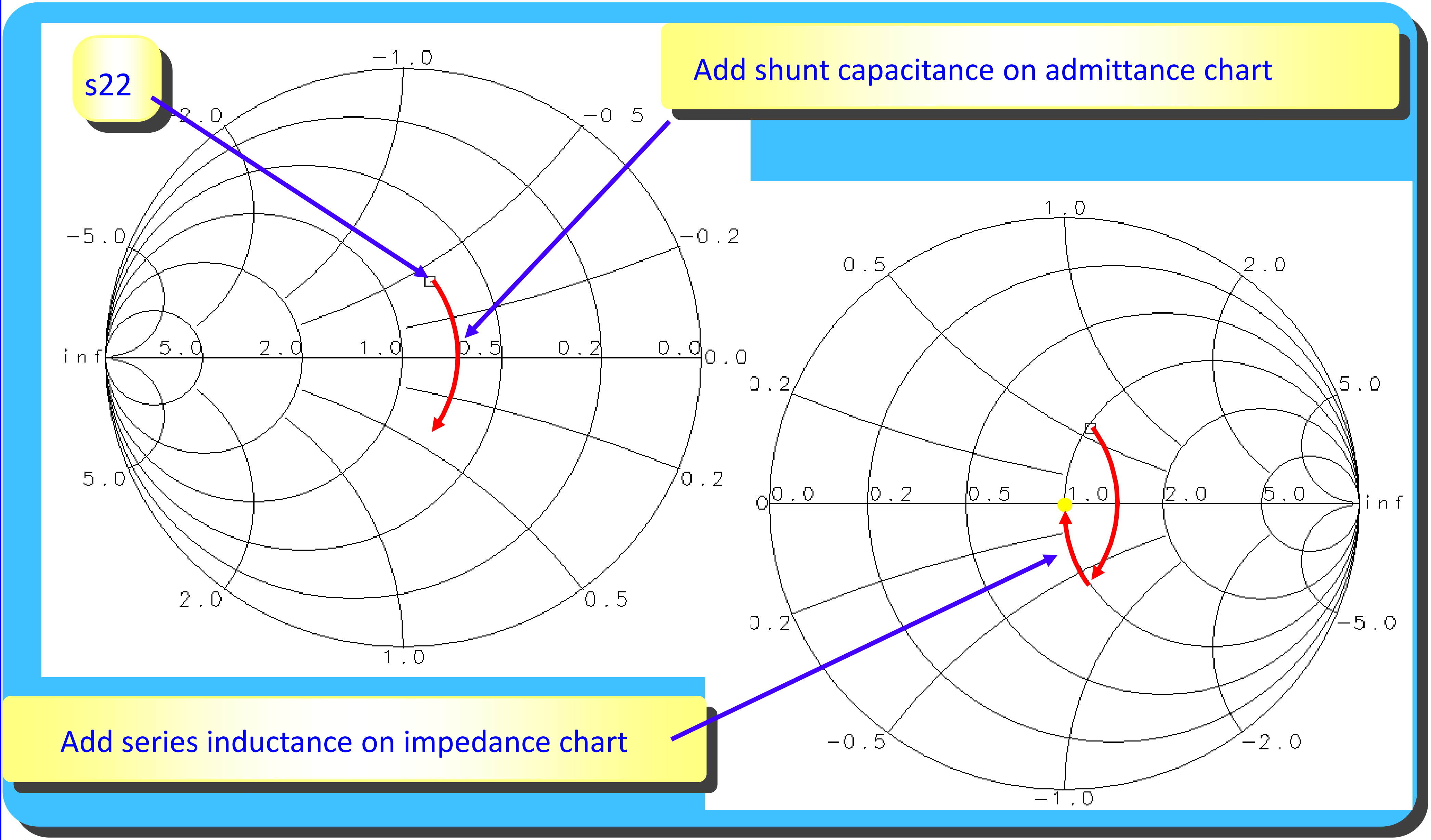


Noise Circles:

Min. Noise: 1.8 dB
+0.25 dB
+0.50 dB

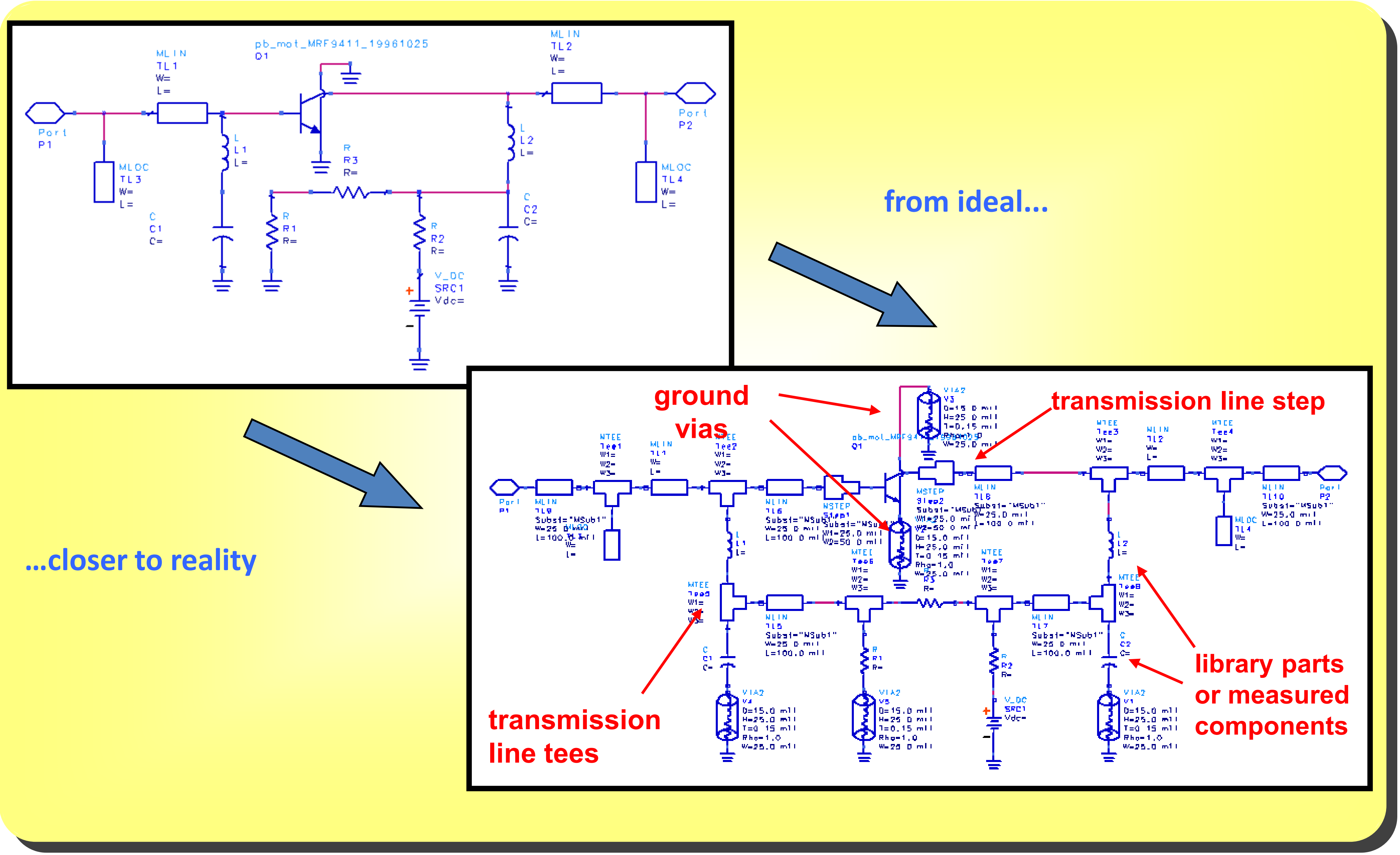
Notice that the match for maximum gain is not the same as the match for minimum noise figure

Output Matching Network

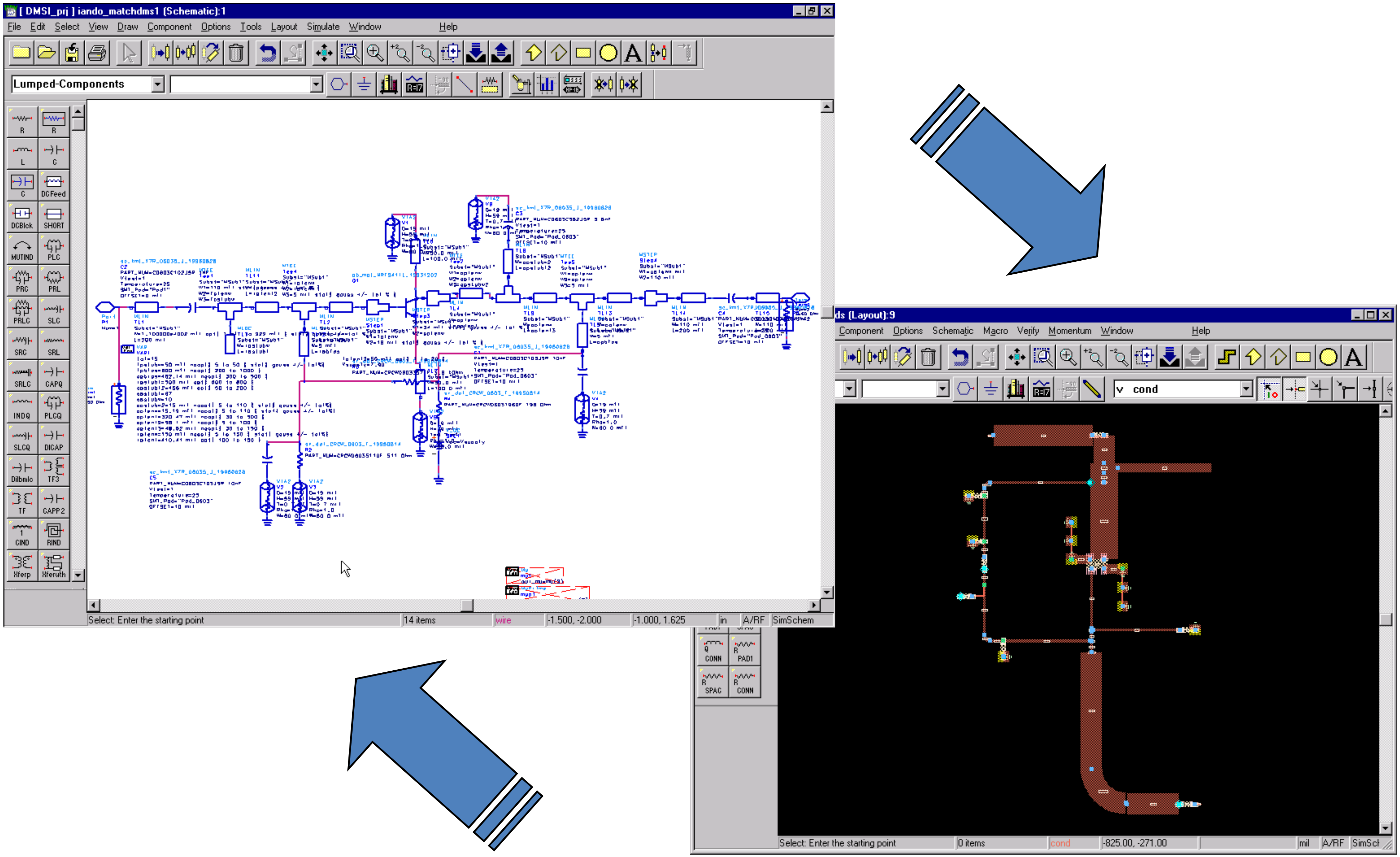


LAYOUT: CO-SIMULATION

Refining the Design: Layout



Layout & Design Synchronization

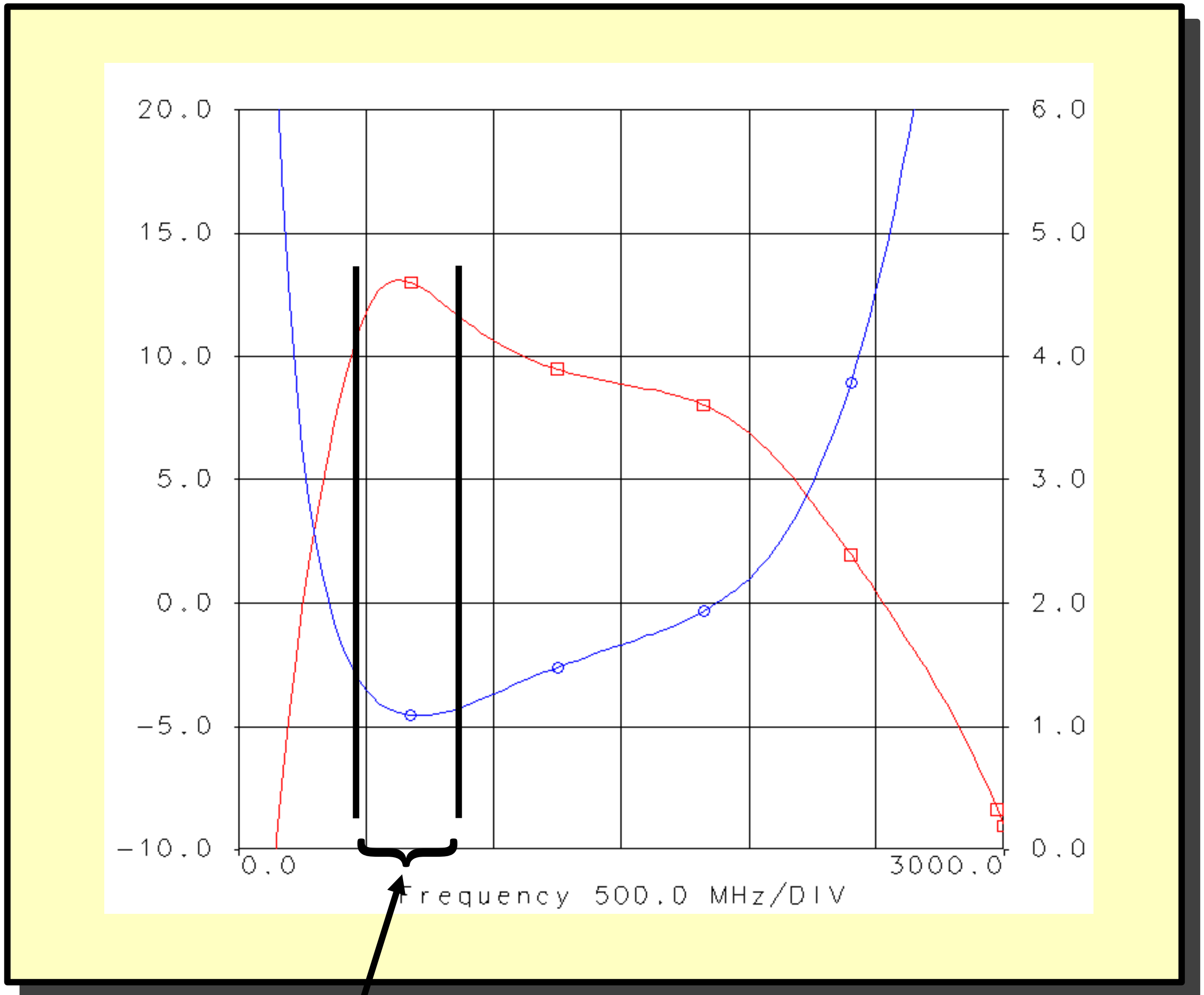


Problems with the Breadboard?

Response:

Things to look for when troubleshooting:

- Stability at <u>all frequencies</u>
- Biasing problems
- System interactions



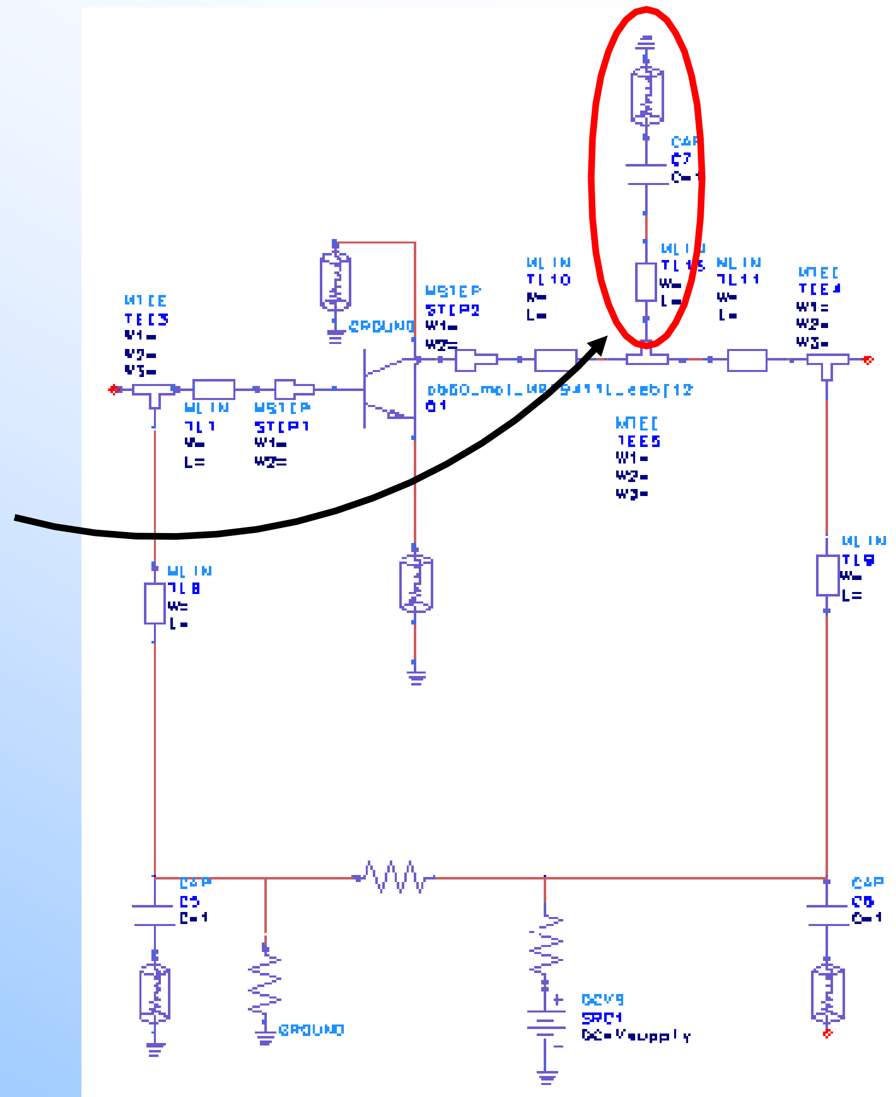
region of possible oscillation

Removing the Oscillation

Hypothesis: collector impedance is too high at lower frequencies

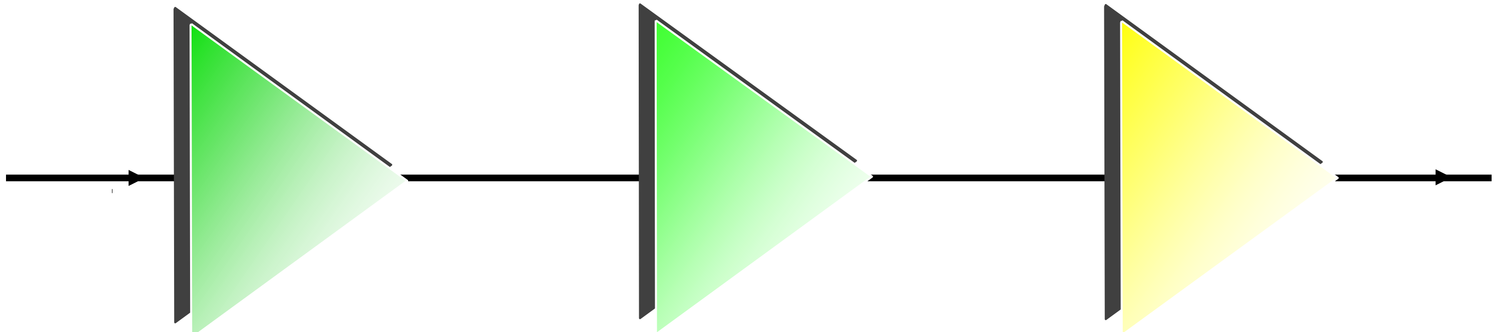
- Remove oscillation by lowering collector impedance at problem frequencies (while maintaining correct impedance in the desired band)

- Shorted stub on the collector



LAYOUT: INTEGRATION

Combined Breadboard

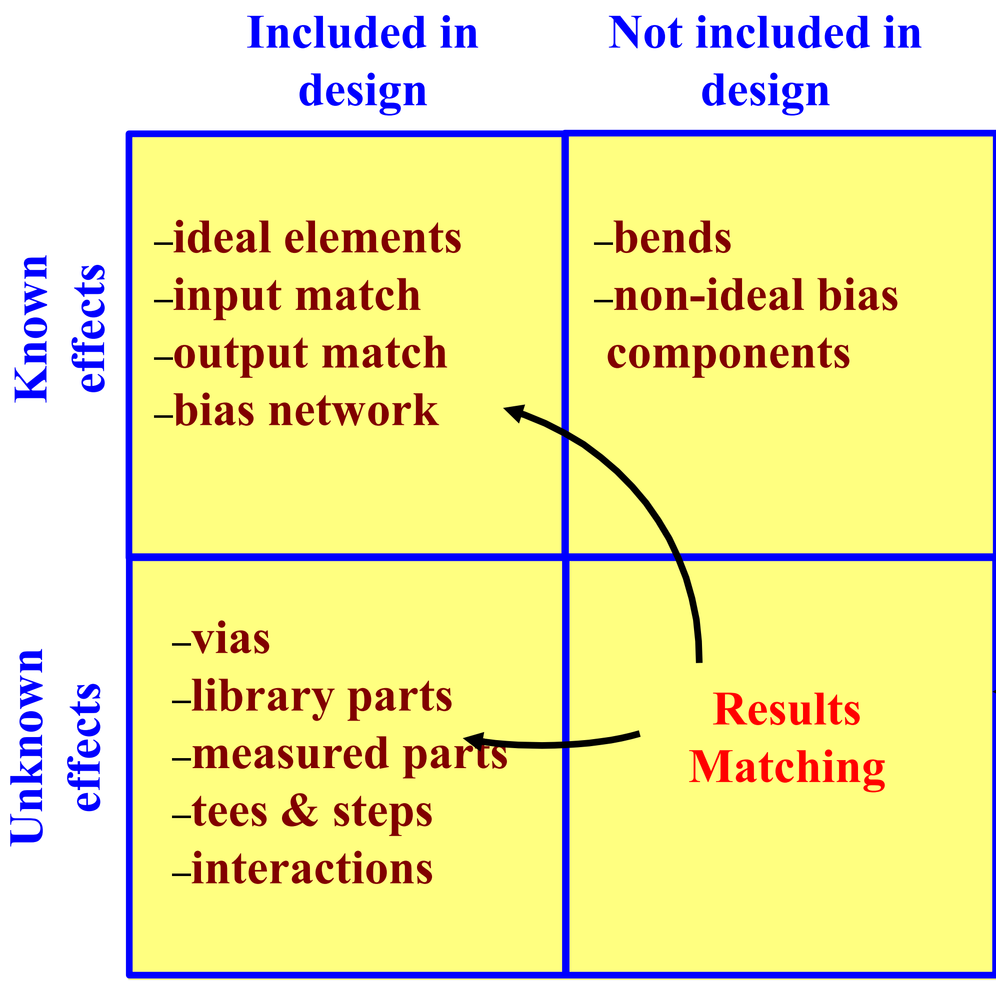


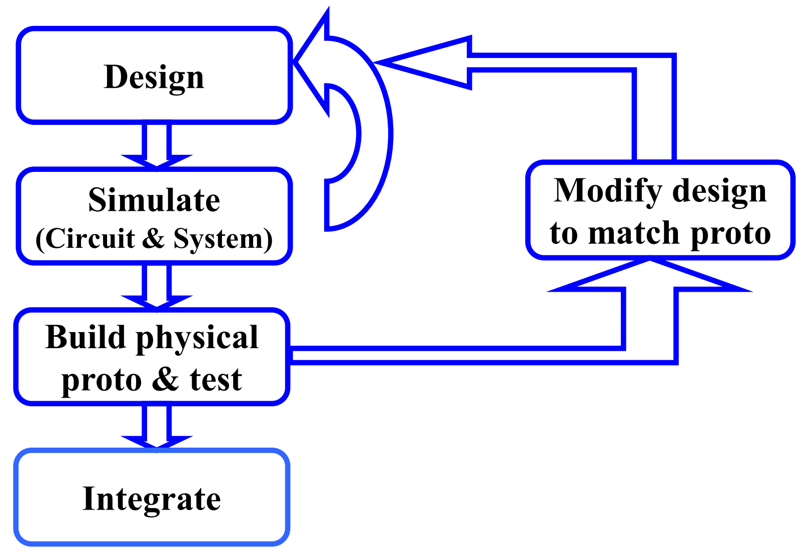
- Three stages
  - Stages 1 & 2: low-noise (identical stages)
  - Stage 3: supply the rest of the desired gain

Combining Issues

- Stability: if even one stage is conditionally stable, circuit may oscillate
- Matching: small mismatches individually, can become worse collectively

Modify Circuit Design to Match Proto



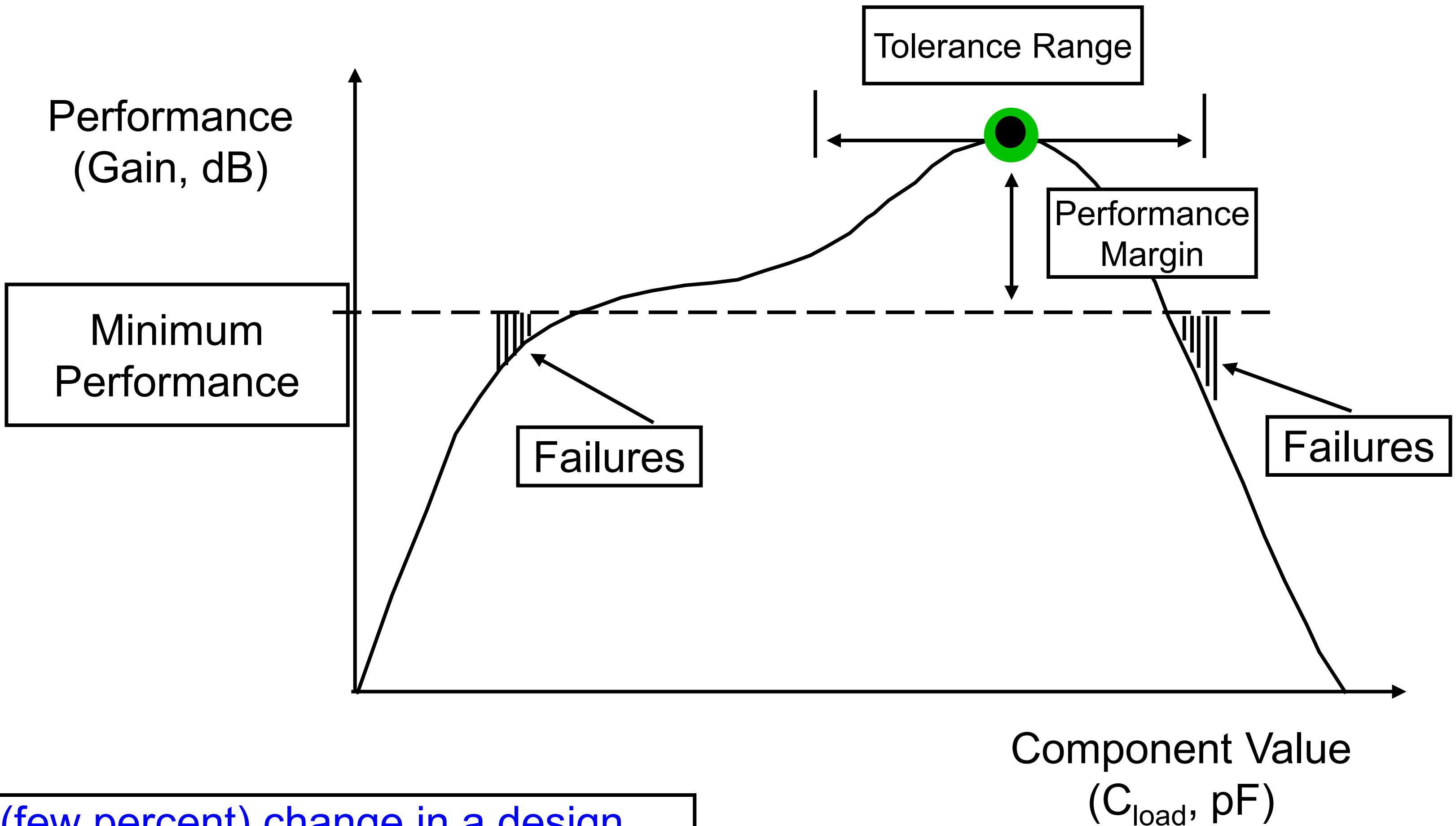


Things discovered in the matching step should be added to the circuit design

LAYOUT: OPTIMIZATION

Performance Optimization Weakness

Maximum performance margin can be near a cliff



A small (few percent) change in a design parameter (like C_{load}, for instance) could have a large effect on the response (many dBs)

Thank you !

End of Chapter 4-2

Chapter 4-3

Microwave Amplifiers

From theory to design:

Designing a power amplifier

Some slides are taken from

- Keysight Technologies (ADS white papers)
- ELG6369 Students Projects

SPECIFICATIONS

1.88 GHz PCS-Band Amplifier (for Personal Communications Service Band)

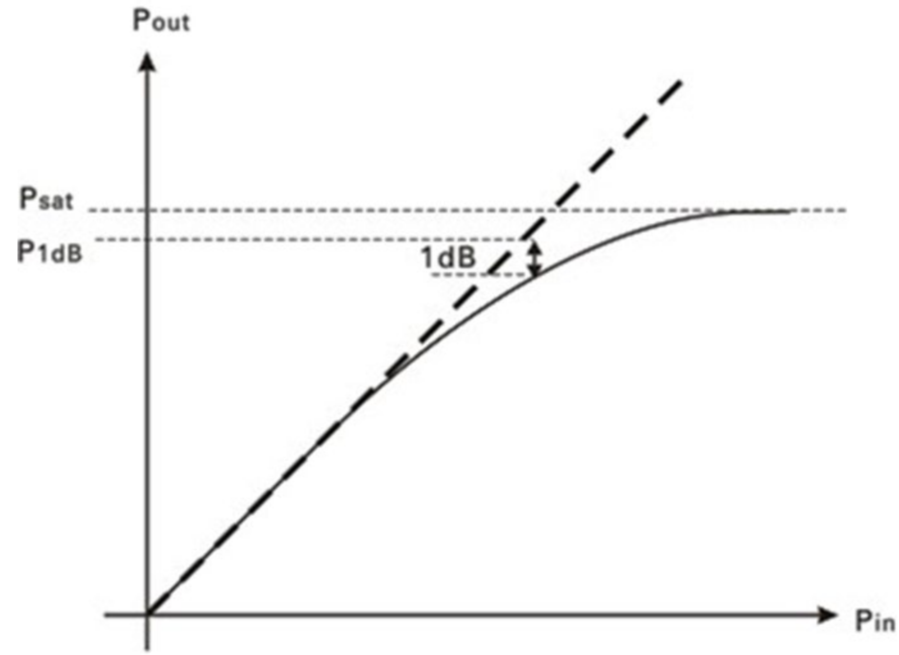
- Gain > +24 dB
- 1-dB-compression > +25 dBm
- Psat > +27 dBm

Psat : Saturation power ?

Saturated output power is the **maximum output power** we can get out from an amplifier.

P_{1dB} is the output power when the amplifier is at the 1 dB compression point.

P_{sat} is the output power when the amplifier is saturated.



https://www.keysight.com/main/editorial.jspx?ckey=1458206&id=1458206&nid=-11143.0.00&lc=eng&cc=CA

SPECIFICATIONS

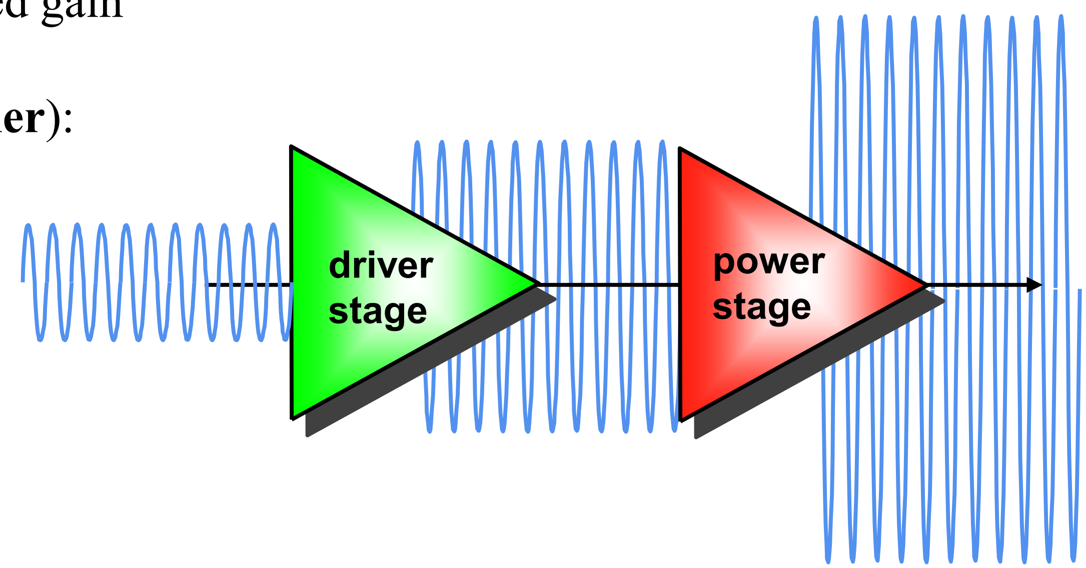
1.88 GHz PCS-Band Amplifier (for Personal Communications Service Band)

- Gain > +24 dB
- 1-dB-compression > +25 dBm
- Psat > +27 dBm

Main challenge: Designing for maximum power output

During pre-design process, decided:

- **Two** stages needed to get desired gain
- Stage one (**Driver - Preamplifier**):
  - silicon transistor
  - passive bias
- Stage two (**Power**):
  - power FET
  - active bias

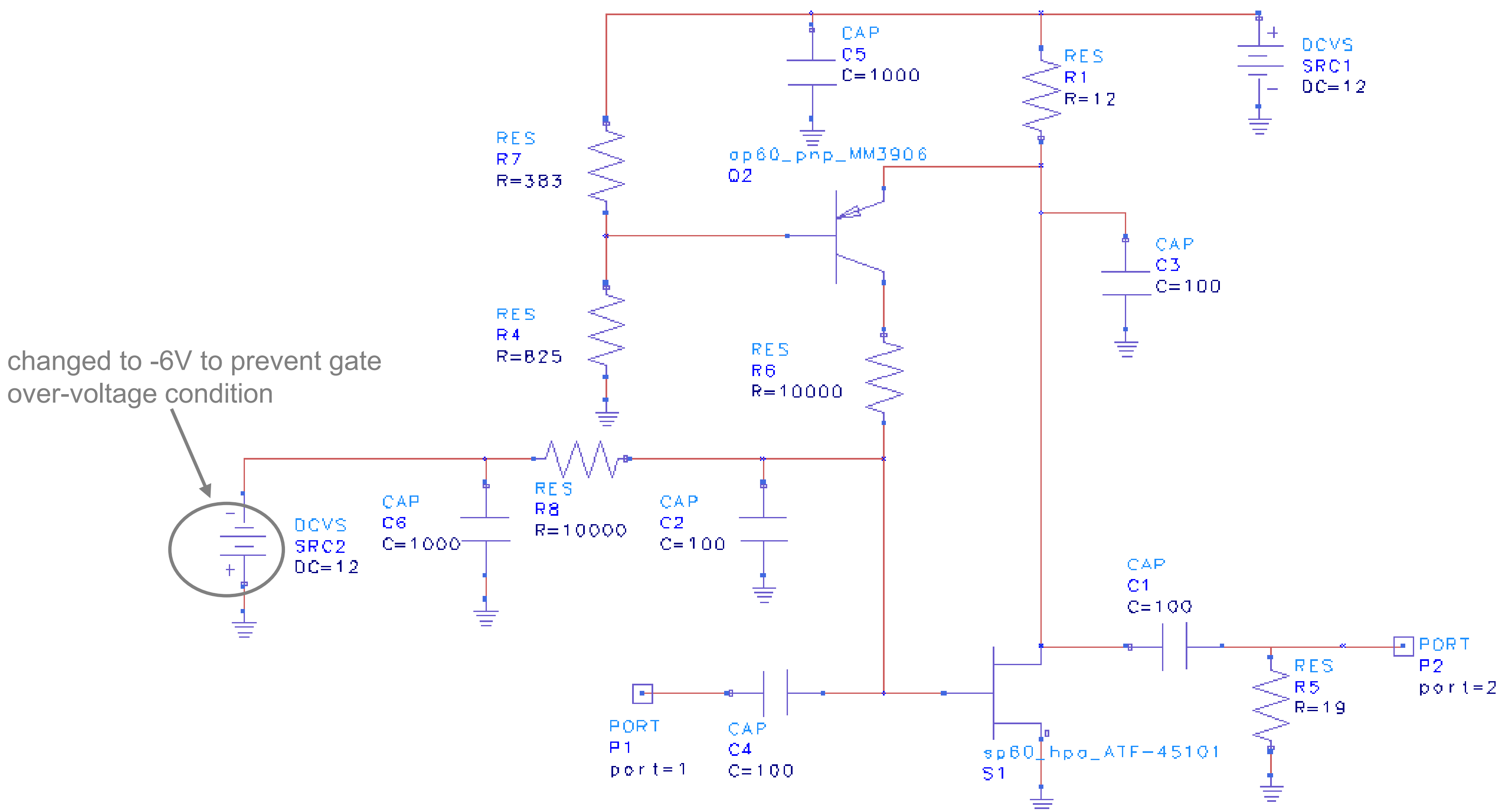


BIAS CONSIDERATIONS - ACTIVE VS. PASSIVE ?

- **Passive** : simpler, less space, less expensive, **BUT** not well controlled.
- **Active** : extra circuitry, more expensive, **BUT** more repeatable:
    - It guarantees a more stable bias current for the power FET without the need for manual adjustments.
    - In fact, Typically, FETs are loosely specified for drain-saturation current (Idss) and gate pinch-off voltage (Vp),
    - So accurate bias using passive circuitry involves adjusting the gate voltage until the desired current is achieved.

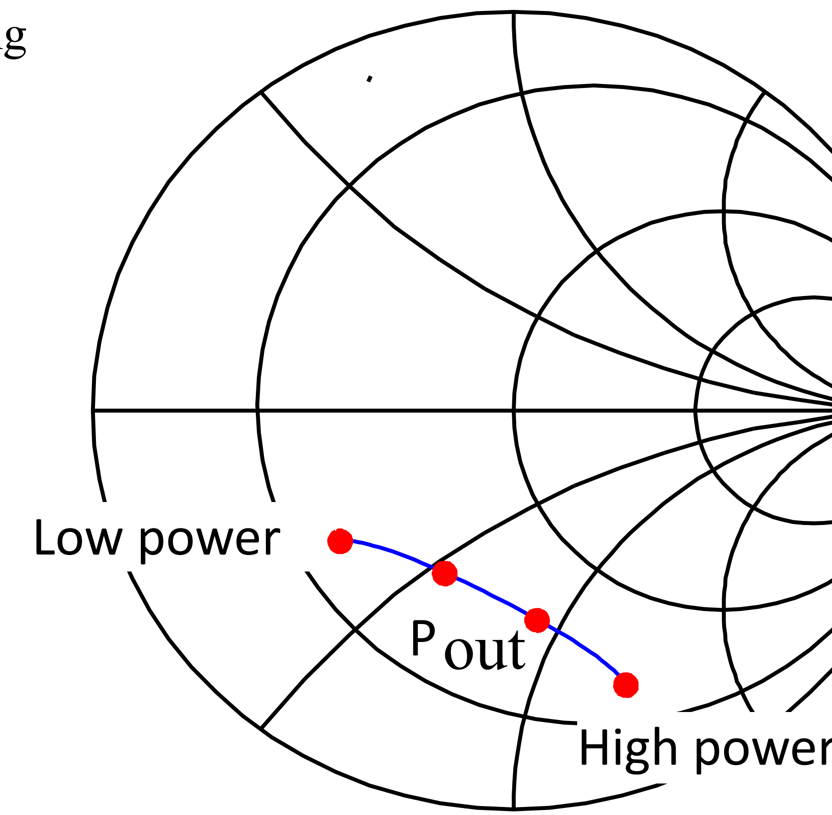
BIAS CONSIDERATIONS

- Hint : watch out for over-biasing junctions (FET gate-bias example)



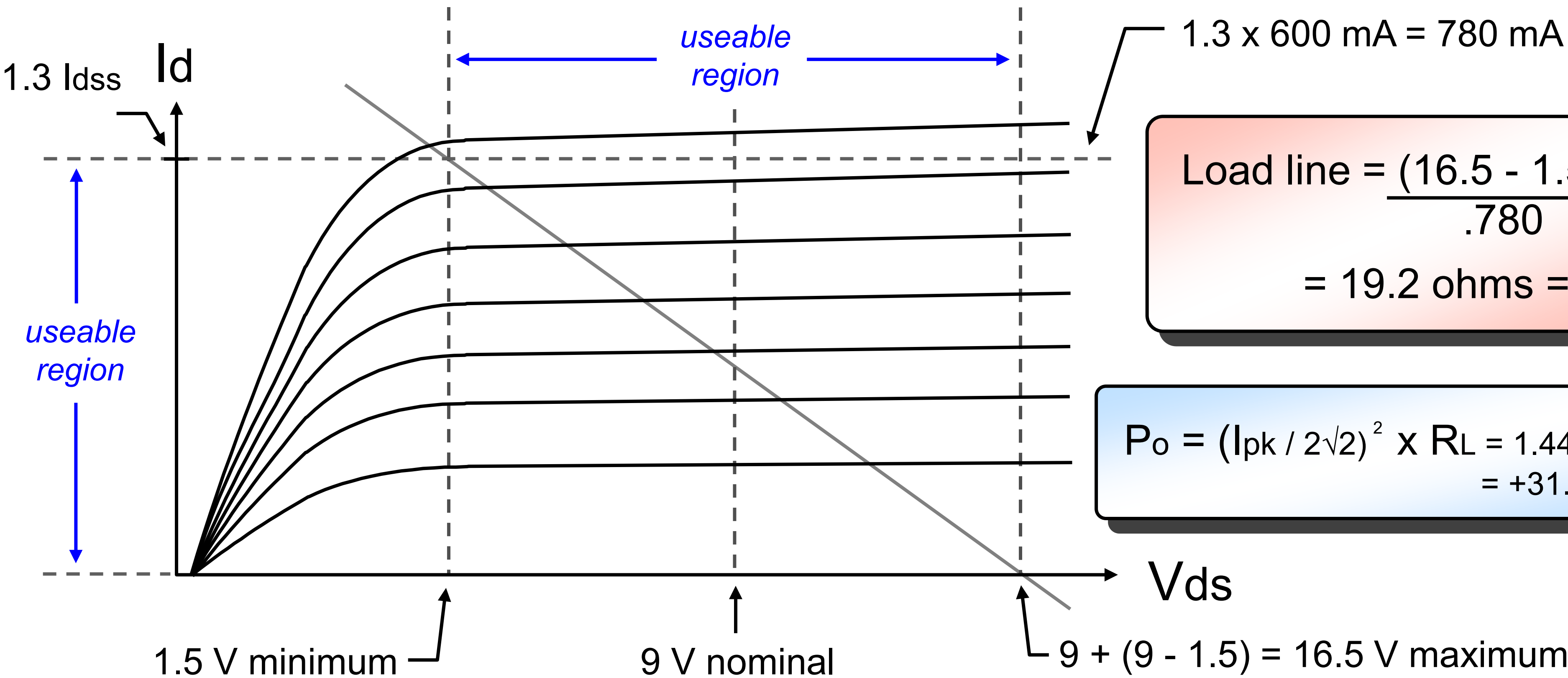
DESIGNING THE OUTPUT STAGE FOR MAXIMUM OUTPUT POWER

- Output impedance varies as function of output power
- Find the best impedance for maximum output power
- Two techniques for output-stage matching
  - load-line analysis
  - load-pull technique



LOAD-LINE ANALYSIS

- Current source of FET needs to be presented with this load
- Should give similar output match as load-pull technique
- Determines resistance that gives highest power:
  - use the I-V curves

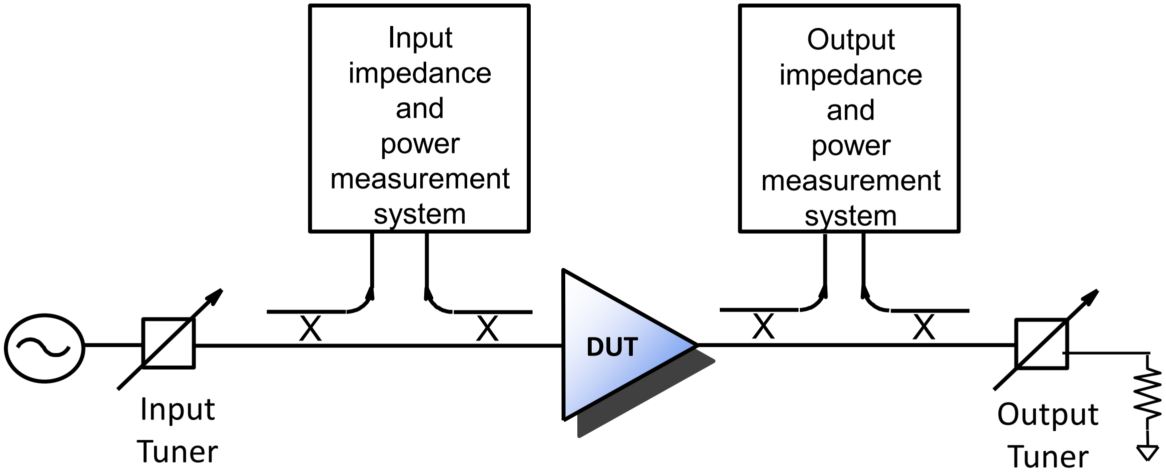


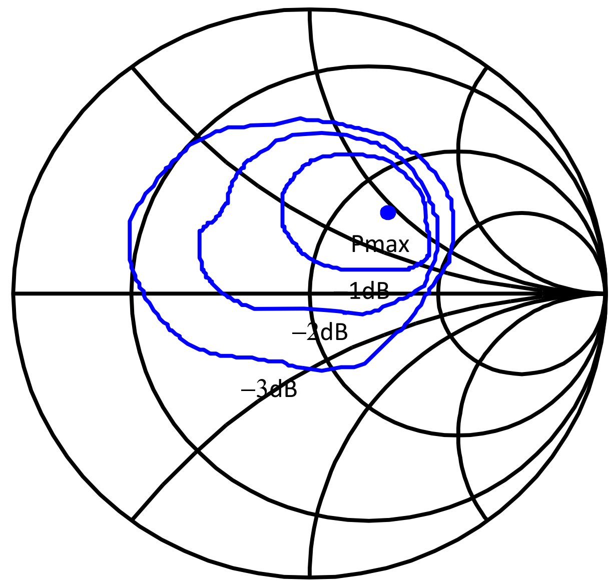
$$\text{Load line} = \frac{(16.5 - 1.5)}{.780} = 19.2 \text{ ohms} = R_L$$

$$P_o = (I_{pk} / 2\sqrt{2})^2 \times R_L = 1.44 \text{ W} = +31.6 \text{ dBm}$$

LOAD-PULL TECHNIQUE

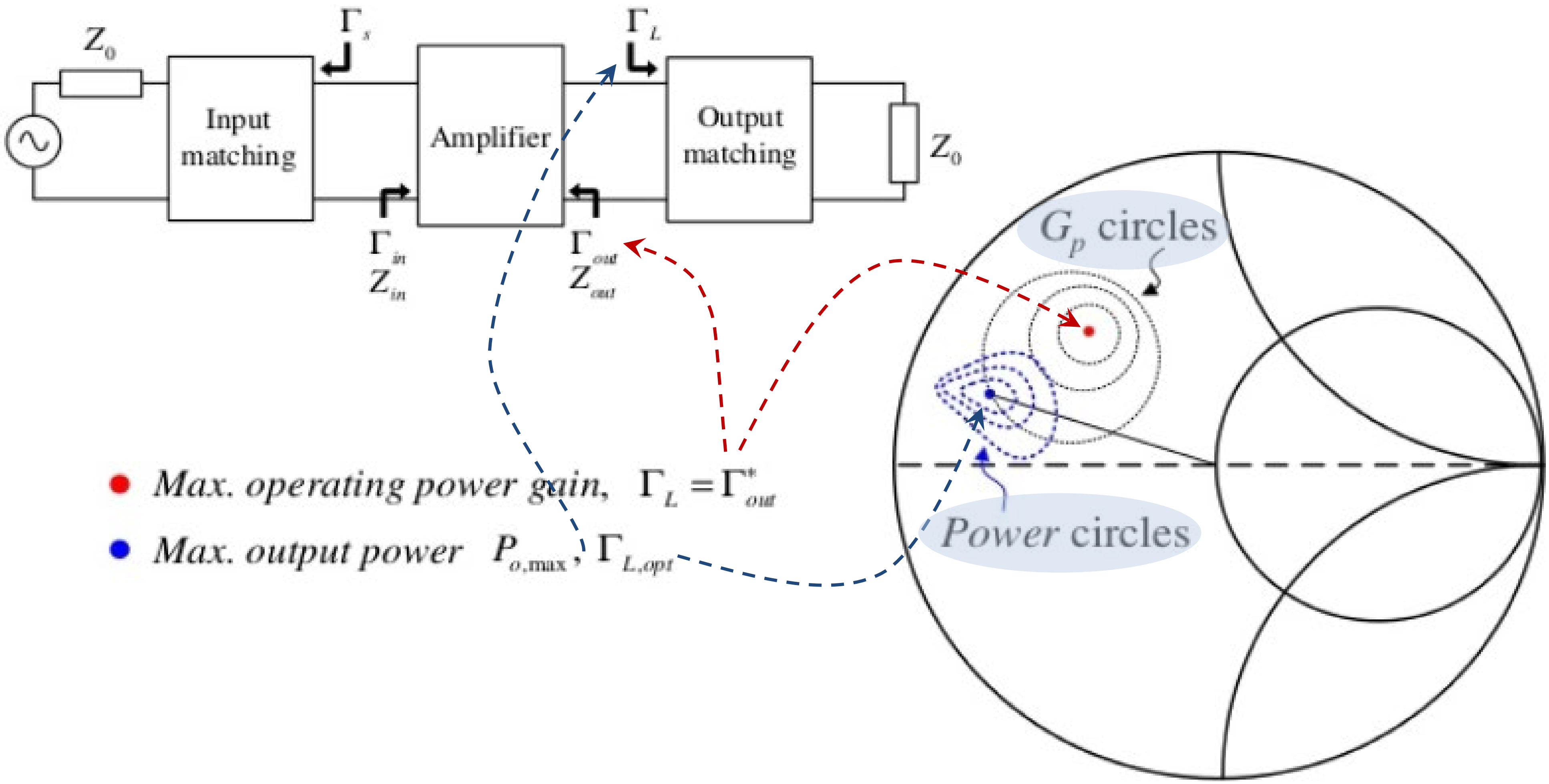
- More precise (link between output impedance and output power)
- Vary magnitude and phase of load presented to circuit to find the power contours
- Power output is measured at each impedance point
- Can use behavioural model (based on measurements)

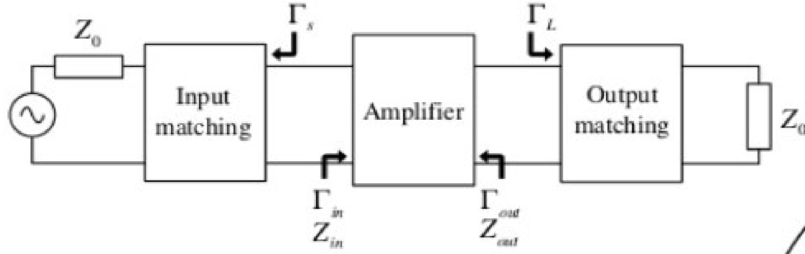


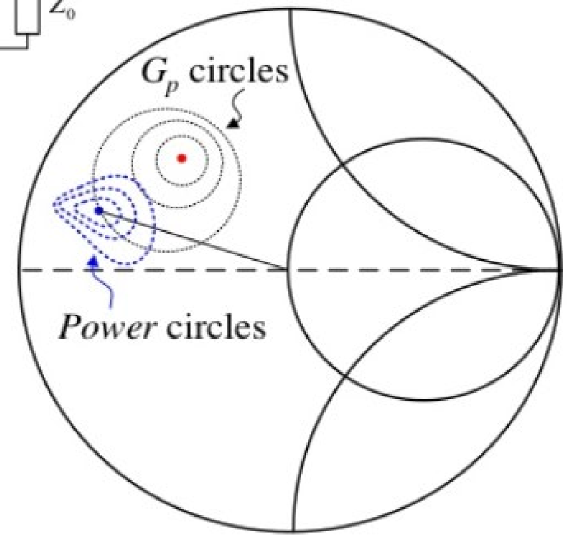


Constant output power contours versus output load impedance (input power constant)

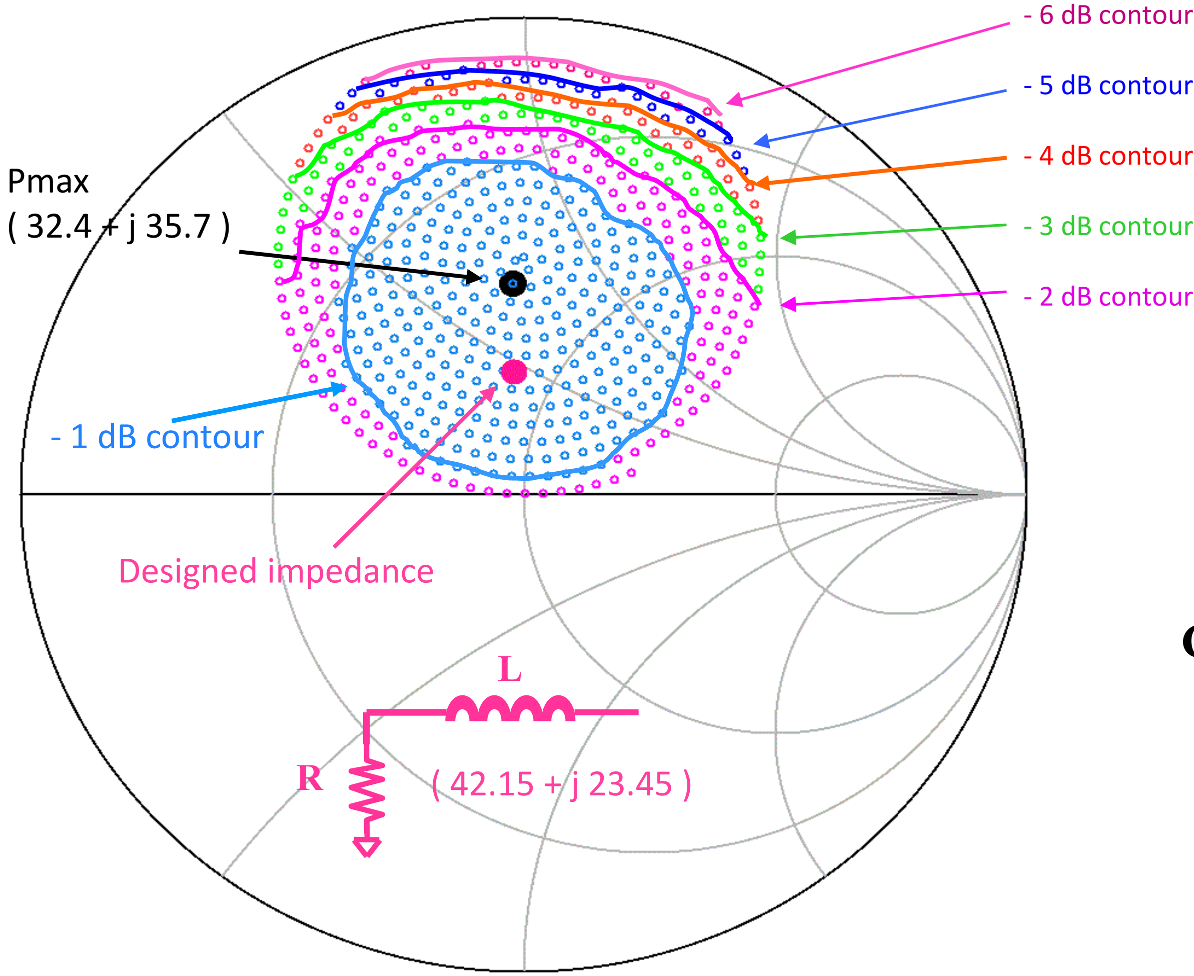
Can be very expensive and time-intensive


Z_0
Input matching
Γ_s
Γ_{in}
Z_{in}
Amplifier
Γ_L
Γ_{out}
Z_{out}
Output matching
Z_0
G_p circles
Power circles
Max. operating power gain, $\Gamma_L = \Gamma_{out}^*$
Max. output power $P_{o,\max}$, $\Gamma_{L,opt}$



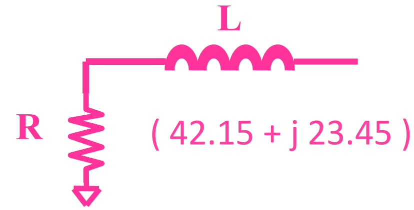


- Max. operating power gain, $\Gamma_L = \Gamma_{out}^*$
- Max. output power $P_{o,\max}$, $\Gamma_{L,opt}$



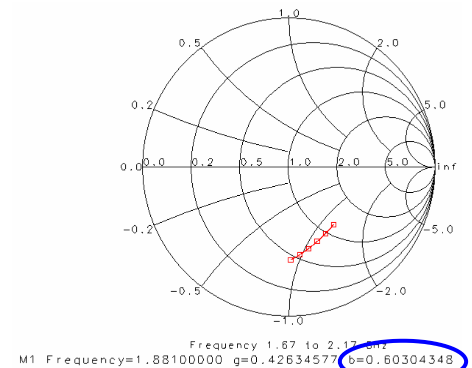
Compromise : Power vs. Stability

Matching to FET Output (Load-line)



- Parasitics must be included in matching circuit
  - include package parasitics (values usually given by manufacturers)
  - assume simple parallel-output capacitance for FET (from S_{22})
  - can use the chip model.

Output match of FET chip



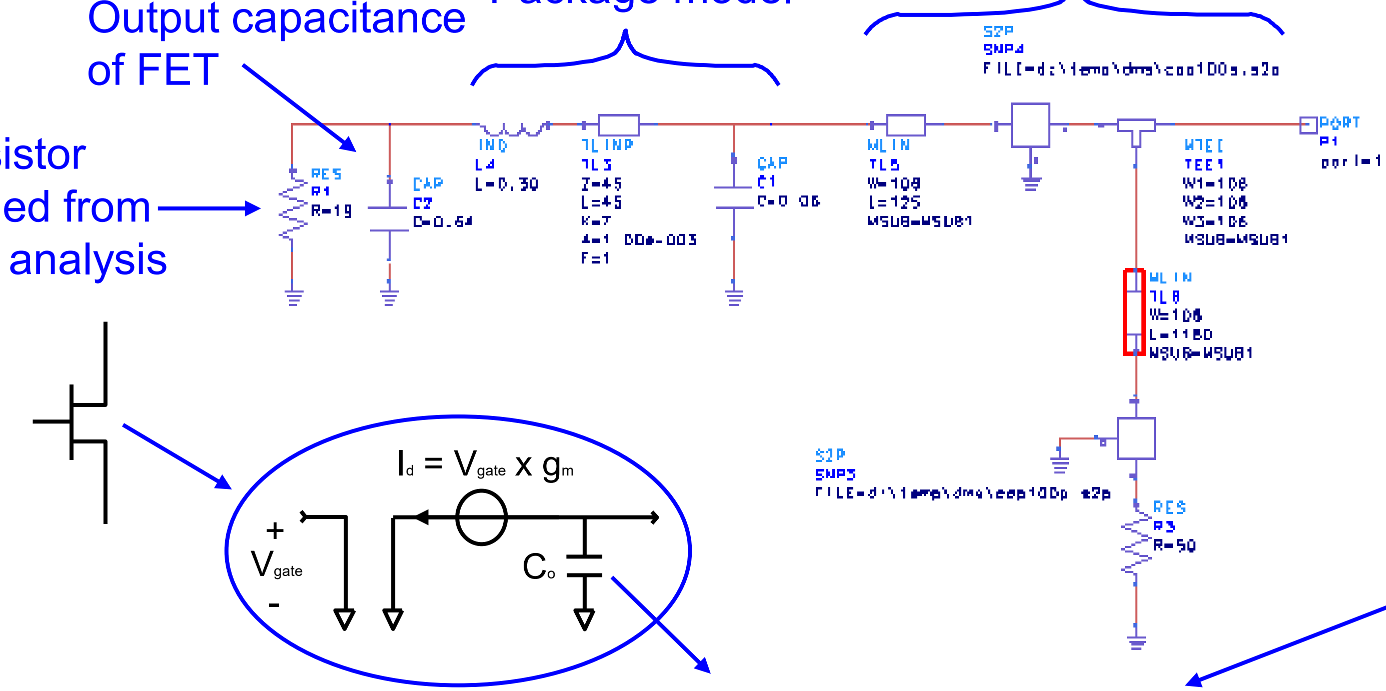
external output matching circuit

Package model

Output capacitance of FET

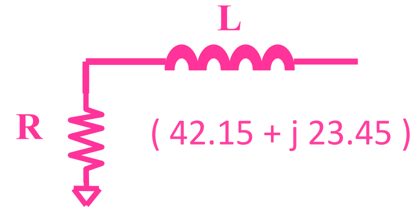
Load resistor determined from load-line analysis

match to 50 ohms here

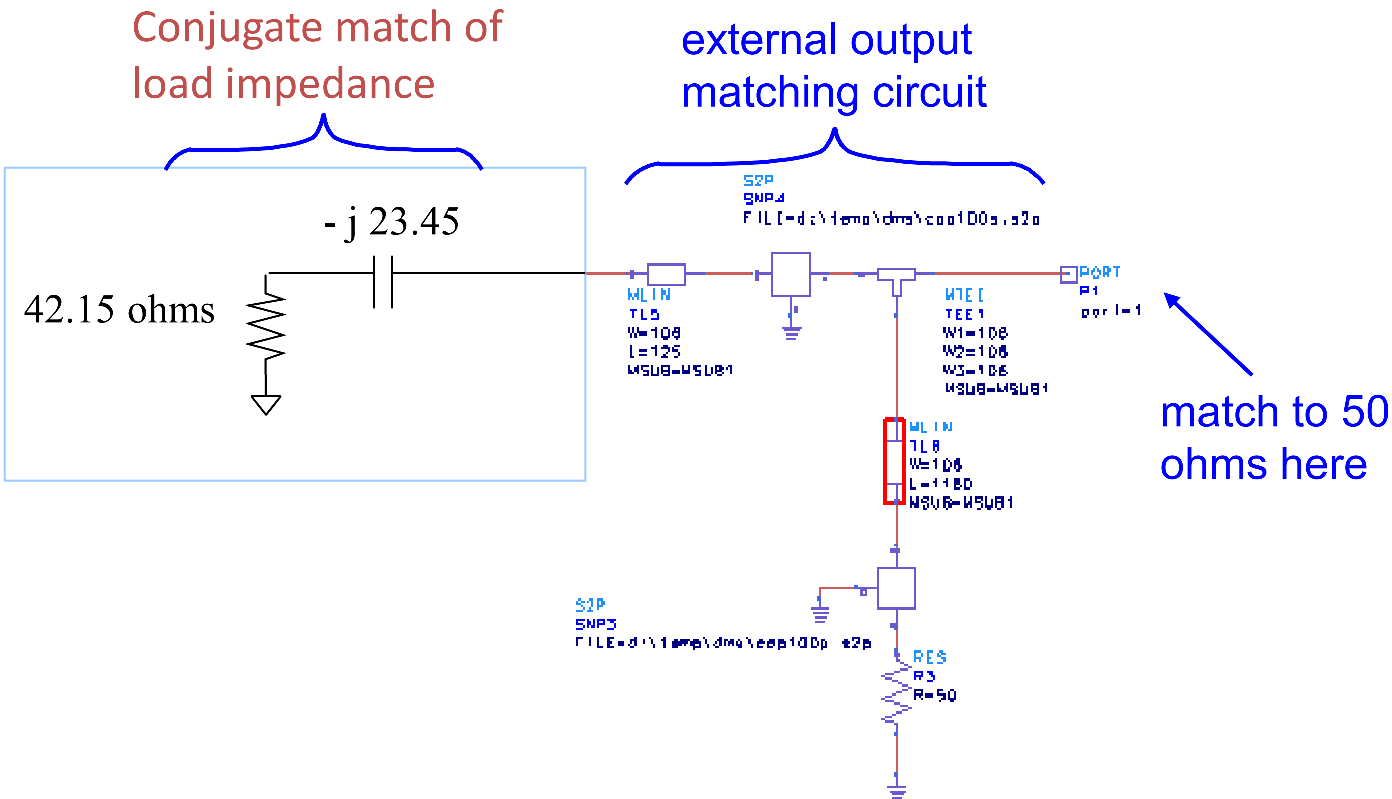


$C_o = y / (2\pi f * 50) = 1.02$ pf

Matching to FET Output (Load-pull)



- No need to model parasitics of FET or package
- Easier and faster BUT much more expensive and time-consuming to setup !
  - Unless we stay on the simulation side (for load-pull)

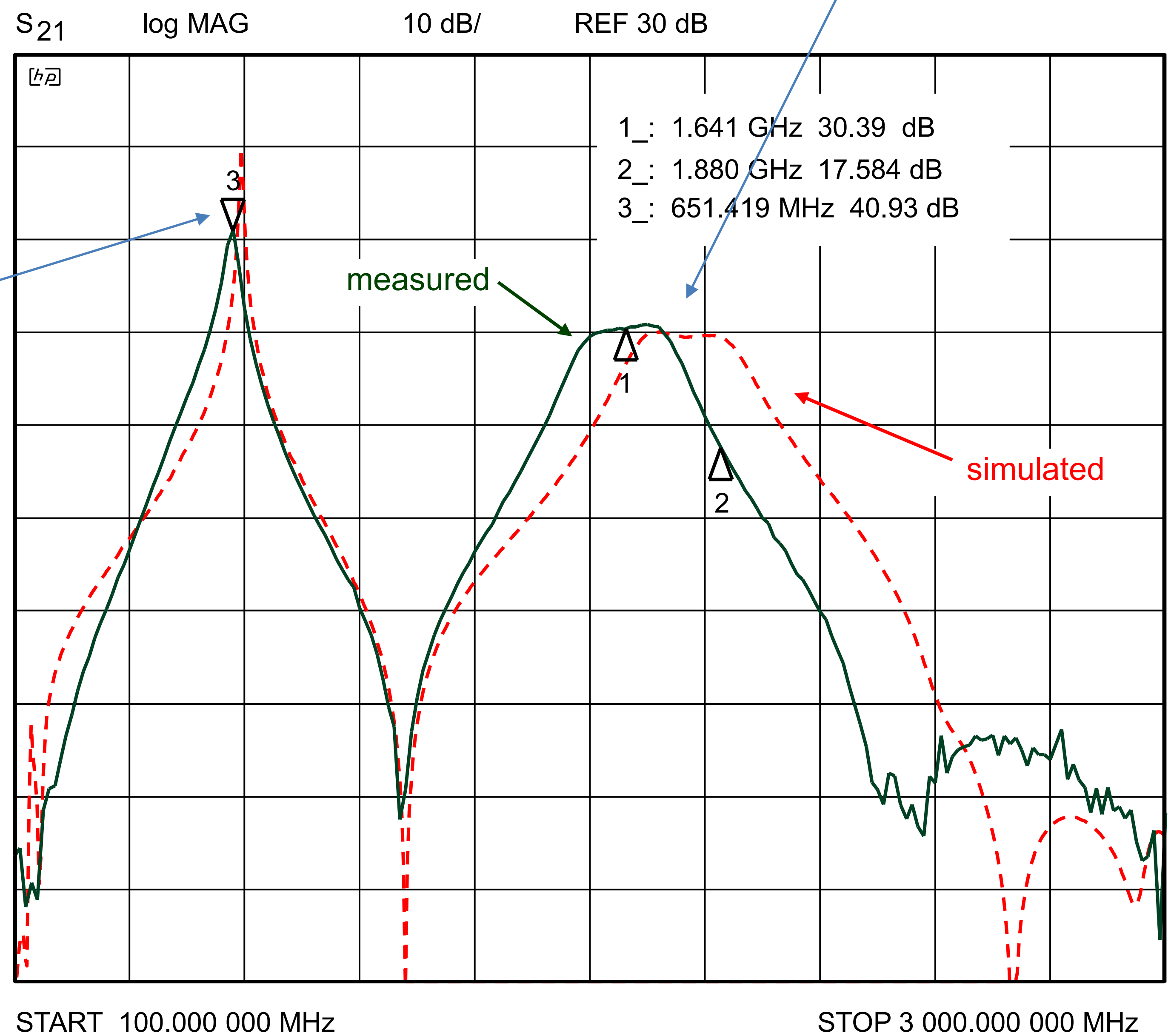


Measured Performance of First Prototype

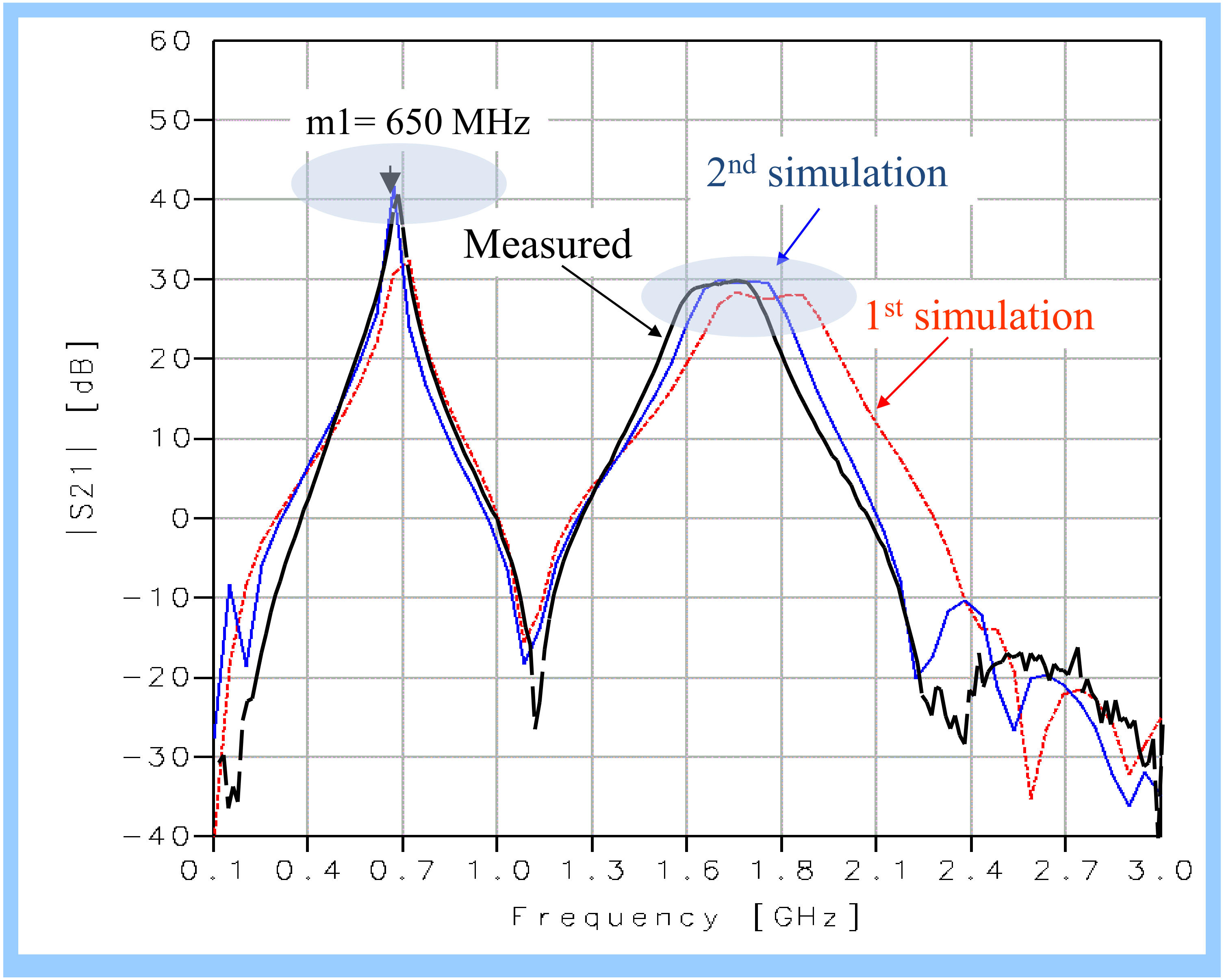
Another problem is that the center frequency of the amplifier is about 240 MHz too low.

Huge high-gain peak: the amplifier will likely oscillate at this frequency.

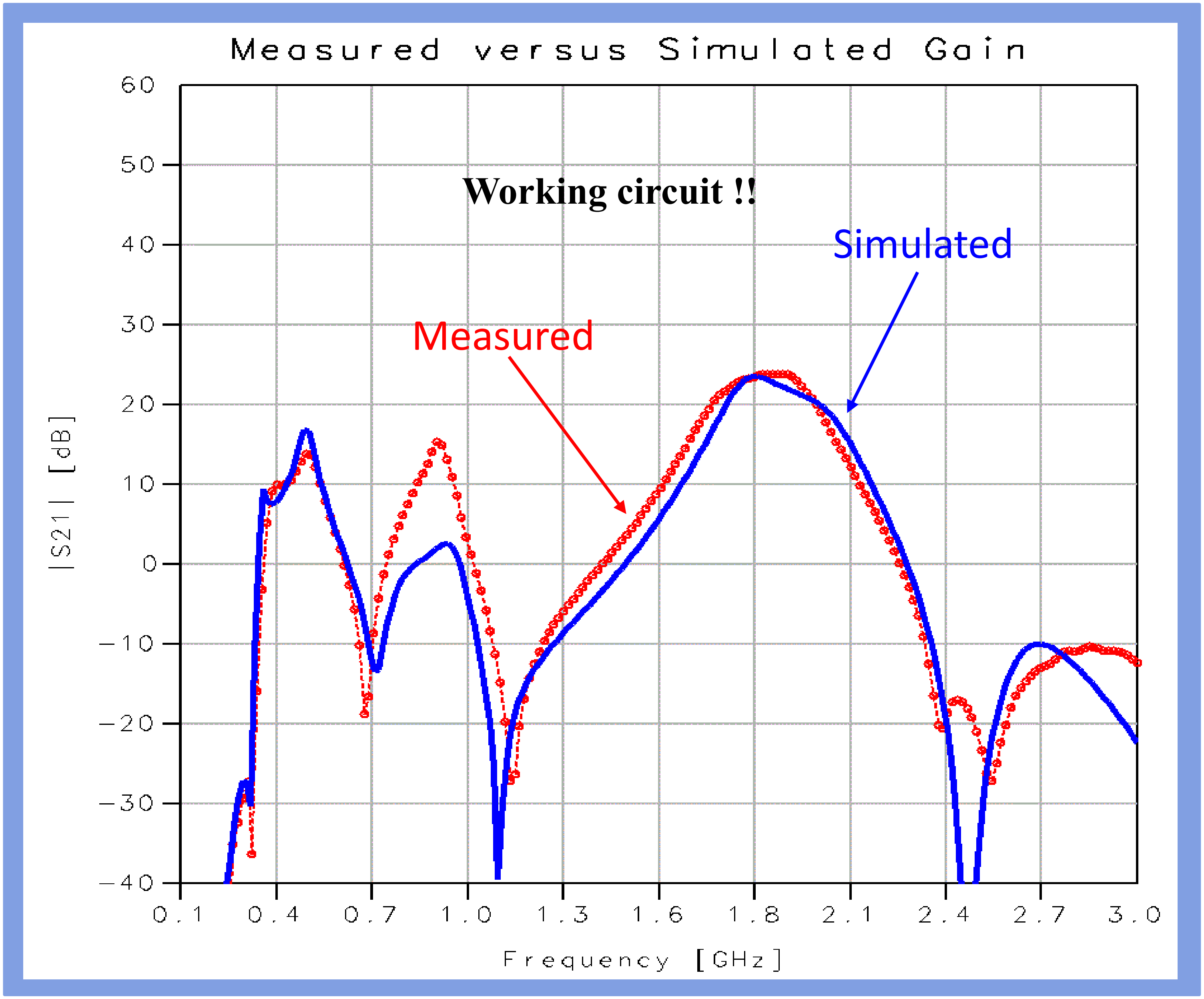
Re-design the circuit !



Modify the circuit design file : 2nd round: change the capacitor models (update the models)



Modify the circuit design file : 3rd round : change the vias dimensions



From theory to design:

Designing a power amplifier

- Second example -

“unusual” Design

DESIGNING A POWER AMPLIFIER FOR A 5G PICO CELL

What is a pico cell?

*"*Femtocells and picocells are a relatively new option to extend cellular network coverage. Media flow between the mobile handset and one of these cells is identical to that which flows between the handset and a standard cellular base station except there is an IP transport segment in between the cell and the cellular cloud […]*"*

(Richard Watson, in Fixed/Mobile Convergence and Beyond, 2009)

WHAT ARE THE CHARACTERISTICS OF A PICO CELL?

Small Cell Type	Cell Radius	Power Level (Watts)	Number of Users
Outdoor DAS (oDAS)	1 mile	20	3,000 per sector
Indoor DAS (iDAS)	Up to 200 feet per antenna	2	2,500 – 3,000 per sector
Microcell	1 mile	10	1,800 per baseband unit
Metrocell	500 – 1,000 feet	5	200
Picocell	750 feet	1	32
Femtocell	50 - 60 feet	0.1	4 - 6
Wi-Fi	50 - 60 feet	0.1	Up to 200 per access point

From: Small cell installations: microcell, metrocell, picocell, femtocell for at&t, verizon, t-mobile, sprint coverage, https://www.signalbooster.com/pages/small-cell-installations-microcell-metrocell-picocell-femtocell

WHAT DO WE WANT TO ACHIEVE?

- High power using low power transistor (2W or below)
- High PAE
- Unconditional stability (if possible – at least : conditional stability)
- Medium gain
- Working at 5G frequencies

DEFINING THE SPECIFICATIONS

- 27.5 to 28.35 GHz
- PAE > 30 % *
- Gain 5 to 10 dB
- Psat > 28 dBm *
- Stability factor > 1

* Specifications to focus on

CHOOSING THE TRANSISTOR

Transistor technology

GaAs	*GaN*
Established technology	Good for high frequencies
Good for high frequencies	High power density
Good for small signal	More thermal conductivity
Low noise	High voltage
More linear	High efficiency
Low voltage	

CHOOSING THE TRANSISTOR

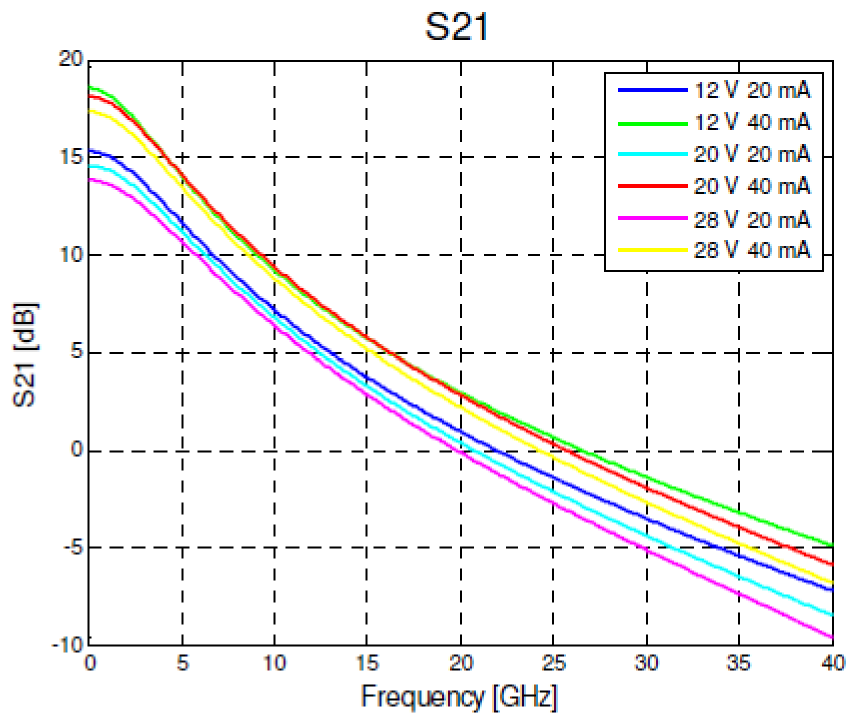
Qorvo TGF2942 is the ONLY transistor in the lab for such frequency range!!

Product Overview

The Qorvo TGF2942 is a 2 W (P_{3dB}) discrete GaN on SiC HEMT which operates from DC to 25 GHz and 28 V supply. The device is constructed with Qorvo's proven QGaN15 process. The device can support pulsed, CW, and linear operations.

Lead-free and ROHS compliant

qorvo
all around you



Key Features

- Frequency: DC to 25 GHz
- Output Power (P_{3dB})[1]: 2.4 W (33.8 dBm)
- Linear Gain[1]: 18 dB
- Typical PAE_{3dB}[1]: 59%
- Typical Noise Figure[1]: 1.2 dB
- Operating Voltage: 28 V
- CW and Pulse capable
- Non-linear & Noise Models available

Note 1: @ 10 GHz

Our Specifications ?

- 27.5 to 28.35 GHz
- PAE > 30 % *
- Gain 5 to 10 dB
- Psat > 28 dBm *
- Stability factor > 1

Highly Risky ??

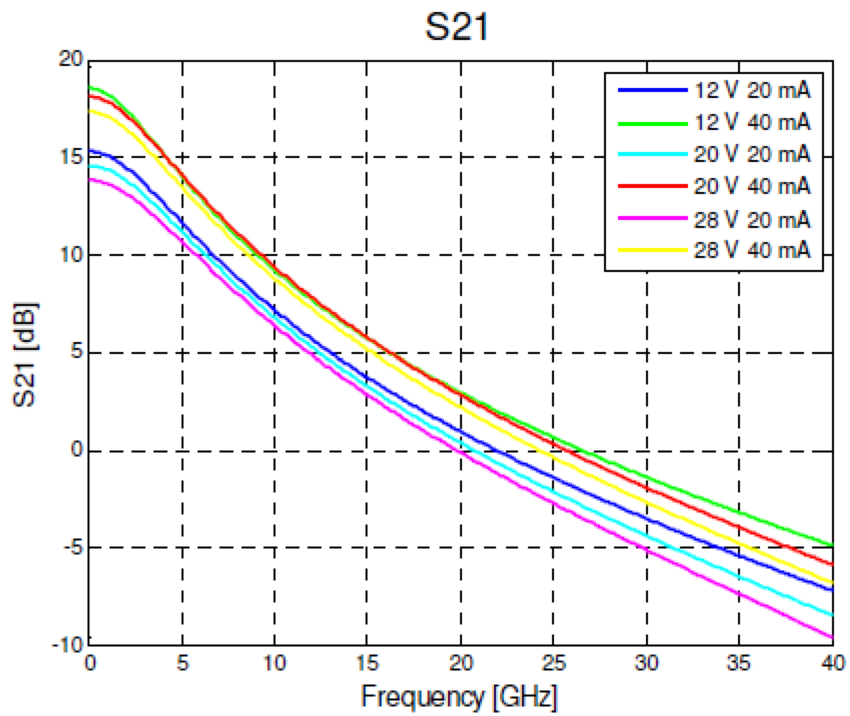
CHOOSING THE TRANSISTOR

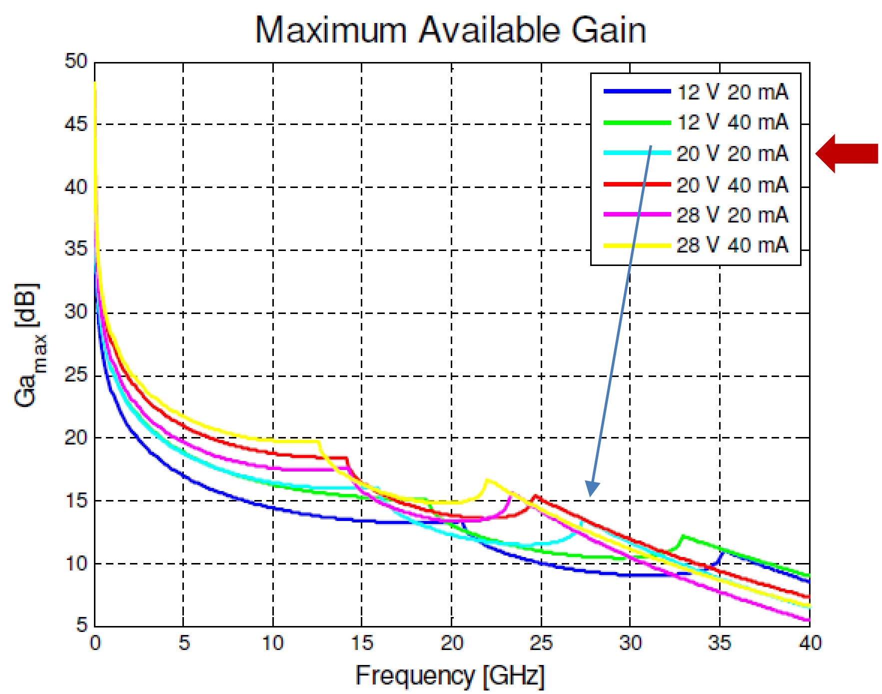
Product Overview

The Qorvo TGF2942 is a 2 W (P_{3dB}) discrete GaN on SiC HEMT which operates from DC to 25 GHz and 28 V supply. The device is constructed with Qorvo's proven QGaN15 process. The device can support pulsed, CW, and linear operations.

Lead-free and ROHS compliant

qorvo
all around you





CHOOSING THE TRANSISTOR

Modelithics®

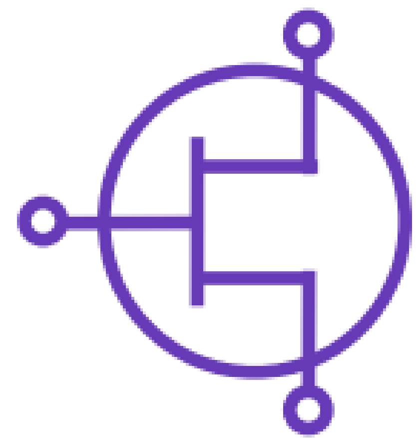
Transistor model?

ADS model provided by Modelithics

Model Features

- Broadband (DC to 40GHz)
- Large-signal model (Angelov-based)
- Optimized Operation: VDS 12V to 28V
- Temperature scalable: (25C to 85C)
- Advanced model feature: enabling intrinsic I-V sensing
- Measurement Validations:
  - Pulsed I-V (25C to 85C)
  - Multi-bias S-parameters (25C to 85C)
  - Multi-bias Noise Parameters (2 to 36GHz)
  - Single Tone Power and Load Pull (10 & 18GHz)

This model was last revised in Modelithics' Qorvo GaN Library ver. 1.9.0.

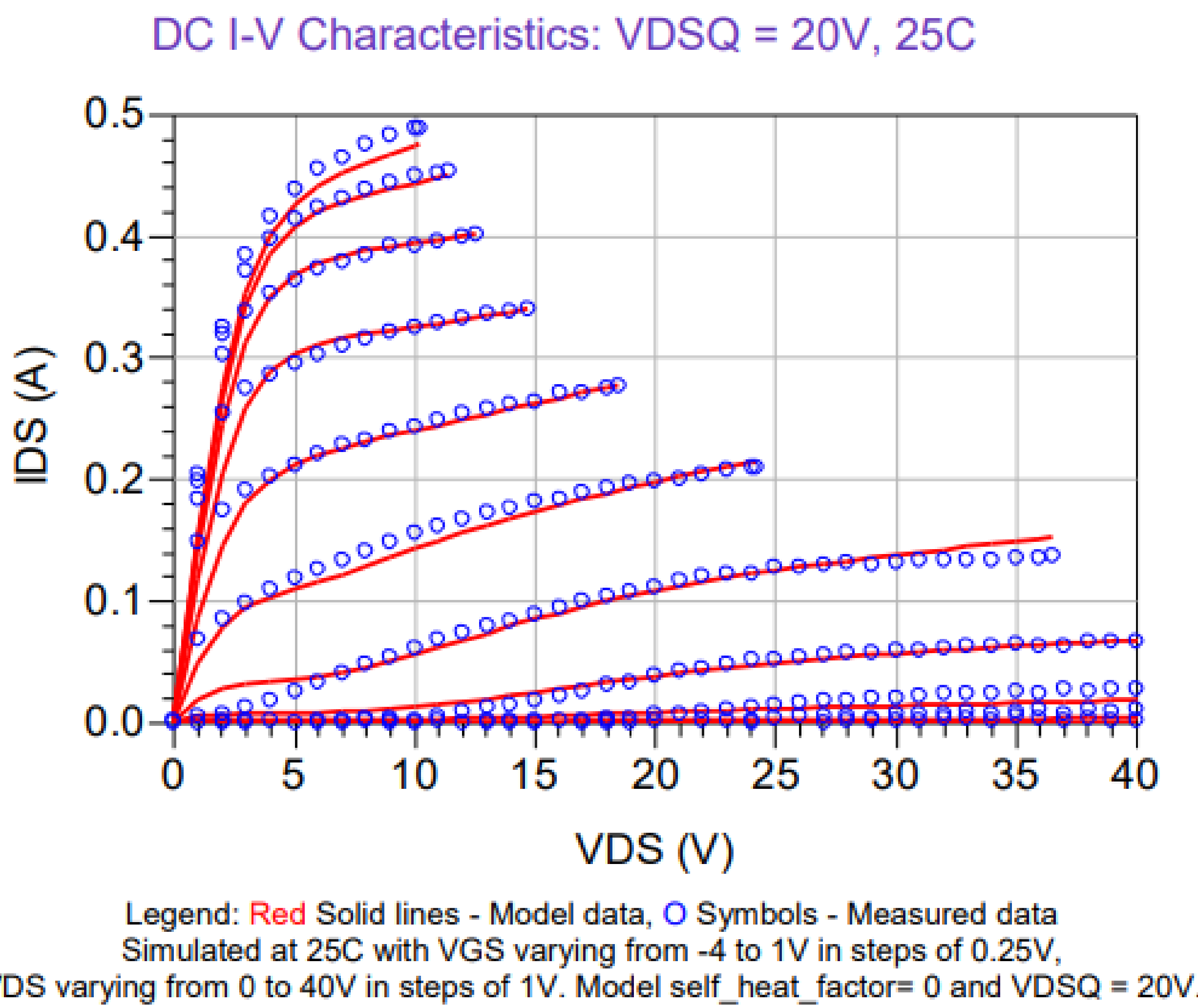


HMT-QOR-TGF2942-001
Qorvo TGF2942
GaN on SiC HEMT

CHOOSING THE TRANSISTOR

Modelithics®

Validation of the model : DC (vs. specifications)



CHOOSING THE TRANSISTOR

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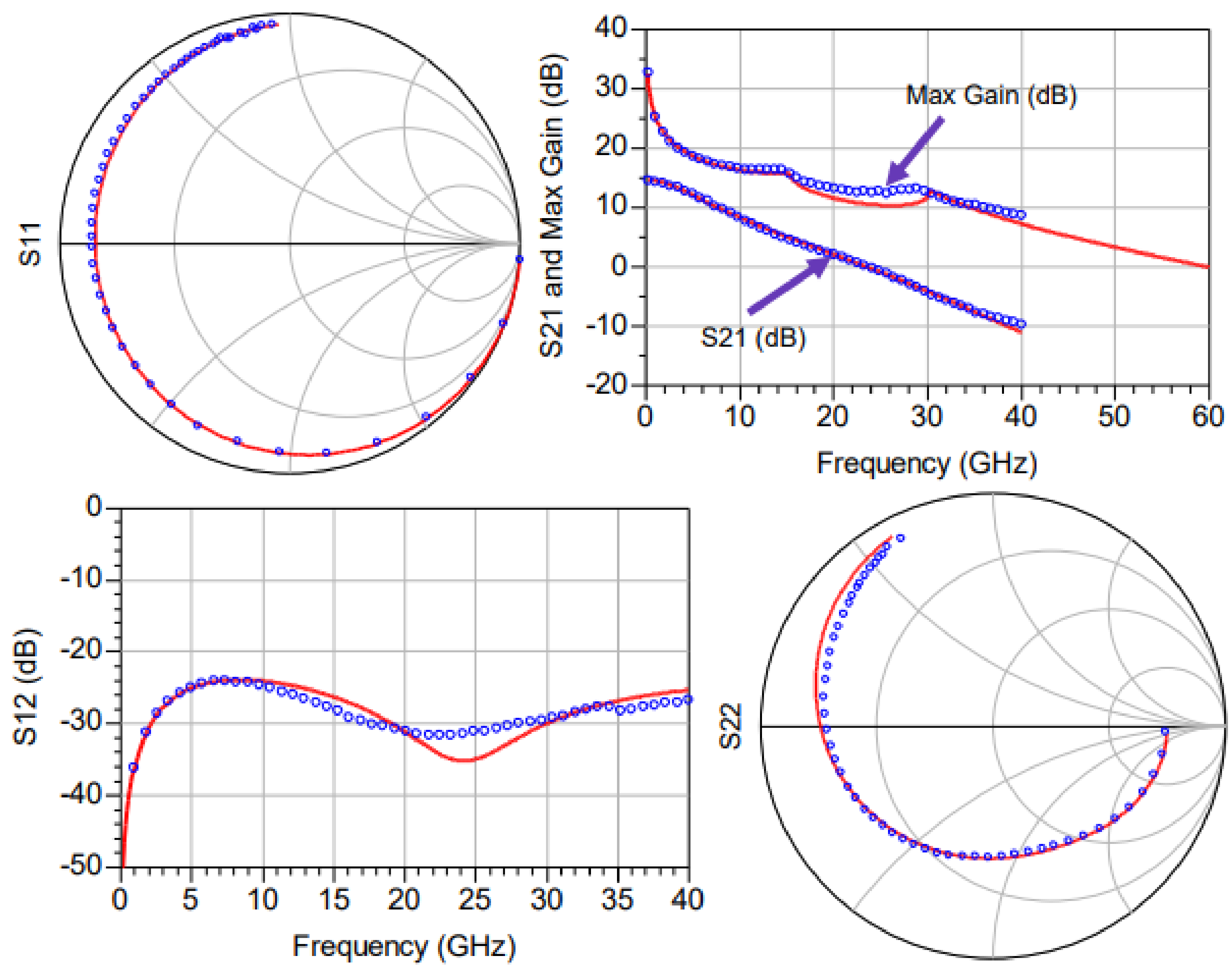
Validation of the model : RF (vs. specifications)

S-Parameters Model vs. Measured: VDS = 20V, IDS = 20mA, 25C

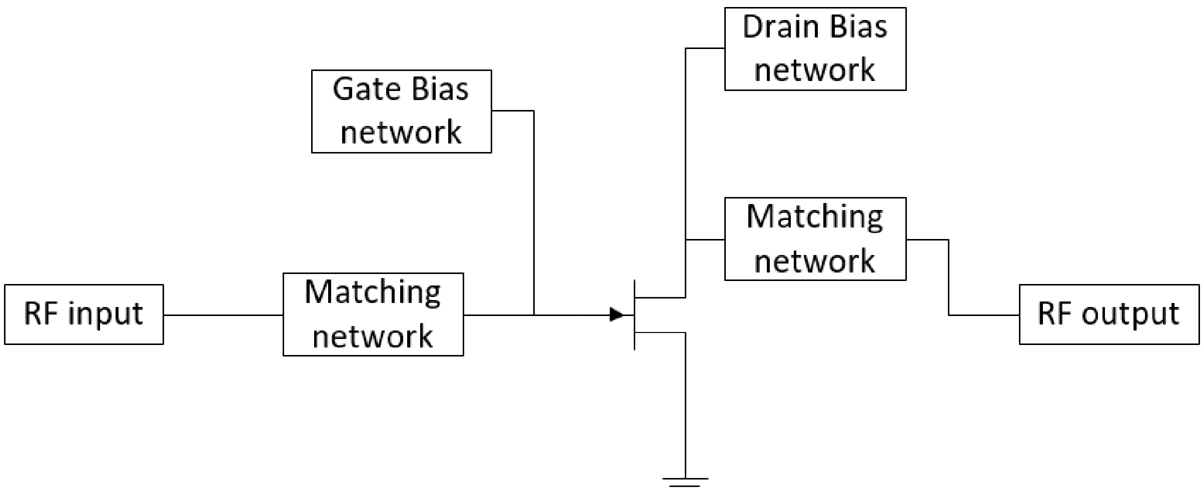


Legend: Red Solid lines - Model data, O Symbols - Measured data
Simulated at 25C with the frequency range from 0.2 - 40GHz. 50Ω Smith Charts

CHOOSING THE TOPOLOGY

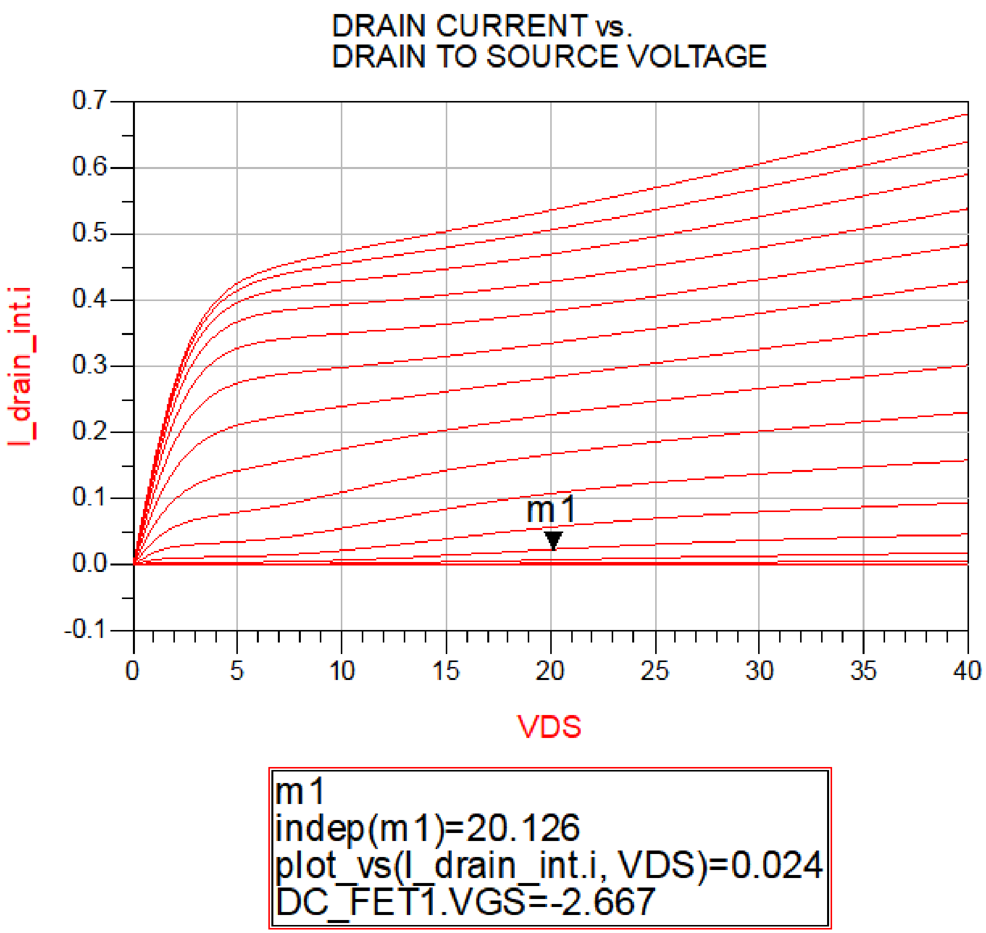
How to achieve the specs?

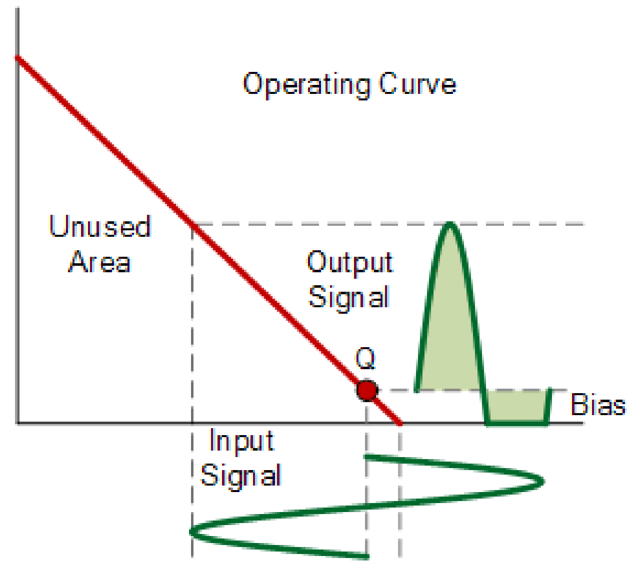
- Class AB
- Single stage
- **Reasons**?
  - Simple
  - Good efficiency
  - No inter-stage matching
  - Good power



PA DESIGN

Bias point

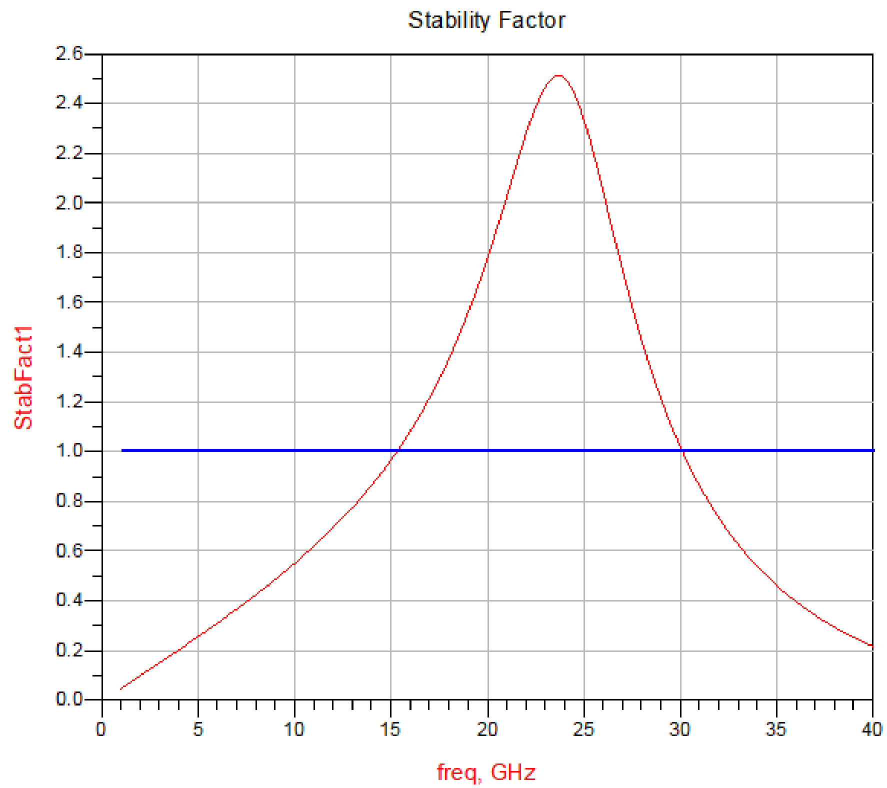


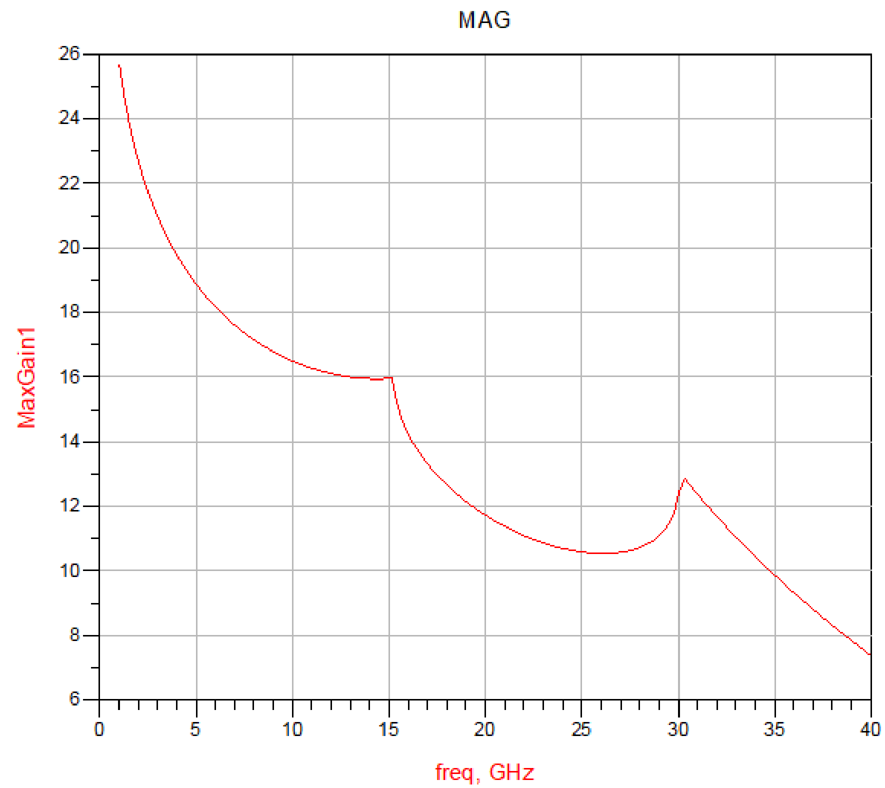


From: Amplifier Classes, https://www.electronics-tutorials.ws/amplifier/amplifier-classes.html

PA DESIGN

Stability and Gain (MAG)

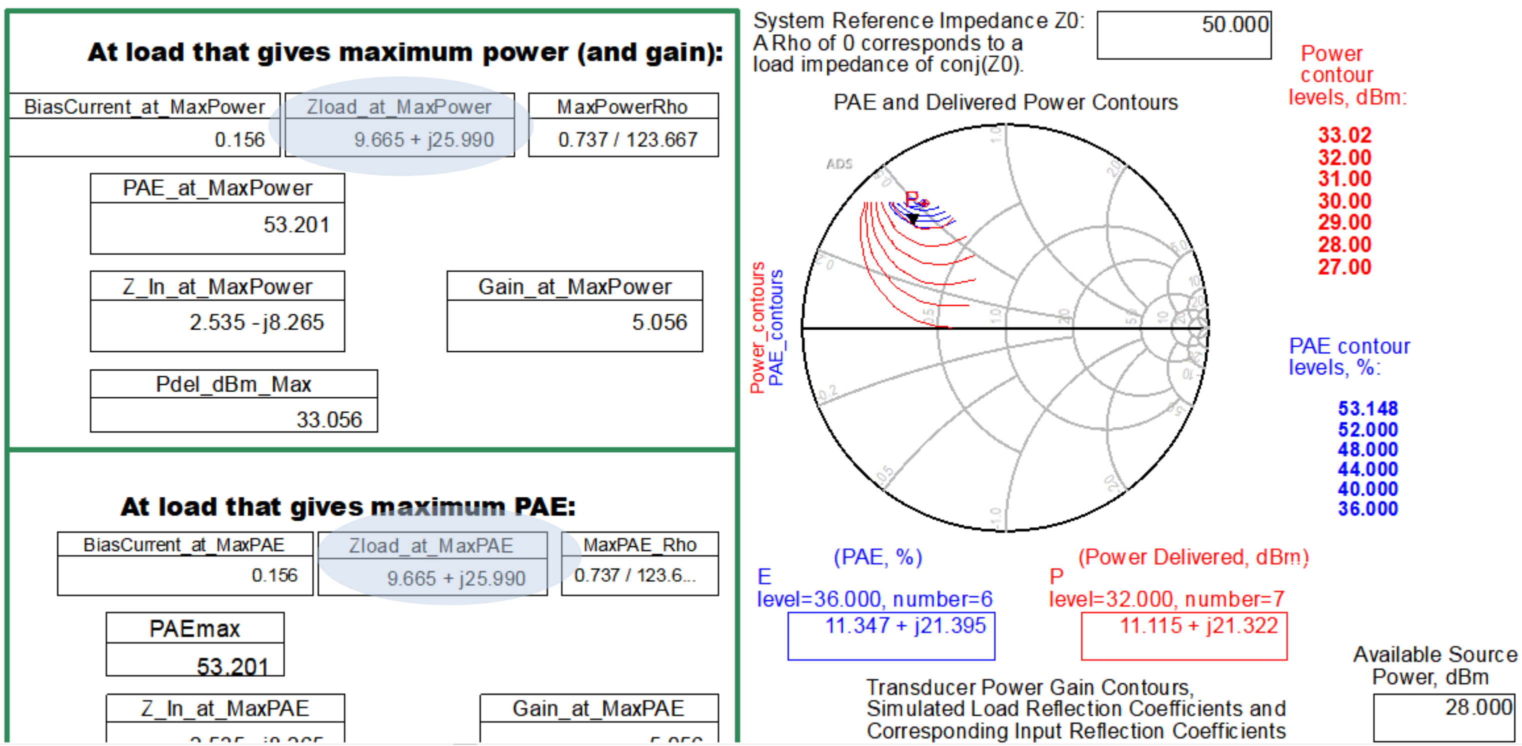




MATCHING NETWORKS

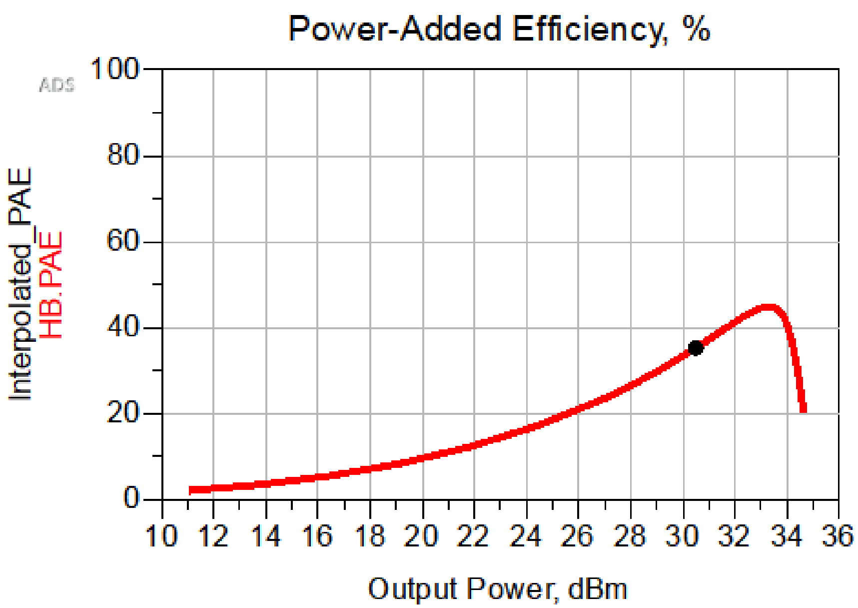
Load-pull simulation

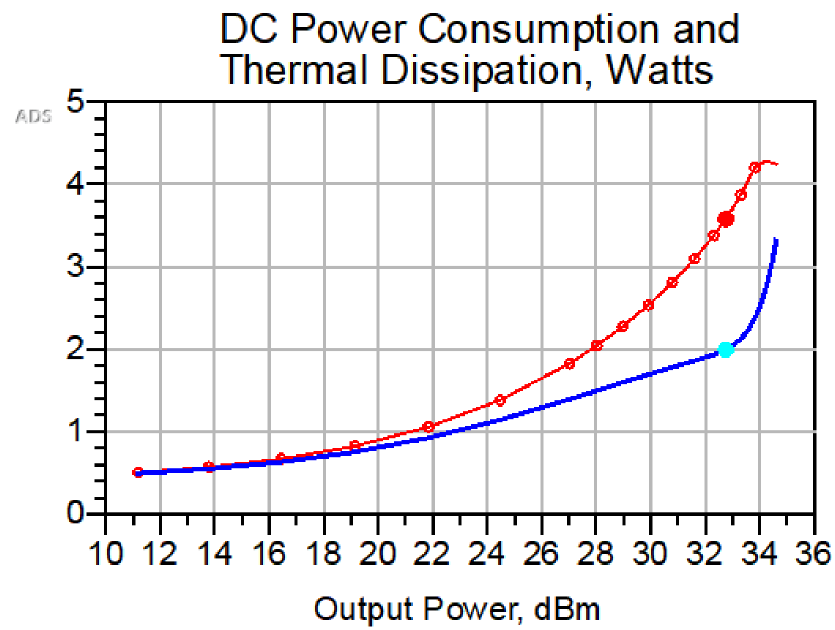
- ADS utility (HB1Tone_LoadPull_ClassAB)
- Input impedance defined at 3 + j 13 (from: Modelithics datasheet, TGF2942)
  - Optimized for 18 GHz
- Output impedance chosen at 11.347 + j 21.395



PA DESIGN

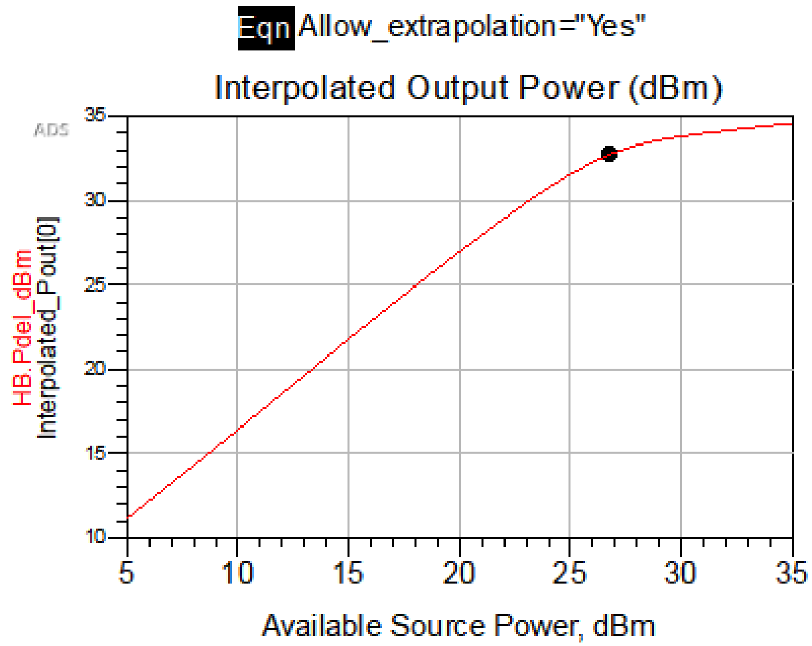
PAE and power consumption





PA DESIGN

1-dB compression point



Interpolated Values at 1.000 dB Gain Compression (approximately):

Gain Compression (dB)	Output Power (dBm)	Transducer Power Gain (dB)	Power- Added Efficiency, %	DC Power Consumpt. Watts	High Supply Current	Thermal Dissipation Watts
1.010	32.760	6.020	44.020	3.578	0.179	1.996

PA DESIGN

Summary !

Goals	*Results*
27.5 to 28.35 GHz	27.5 to 28.35 GHz
PAE > 30 %	PAE of 44 % (P1db)
Gain 5 to 10 dB	Power gain ≈ 6 dB
Psat > 28 dBm	Psat > 33 dBm
Stability factor > 1	K between 1.3 and 1.6 (27.5 to 28.35 GHz)

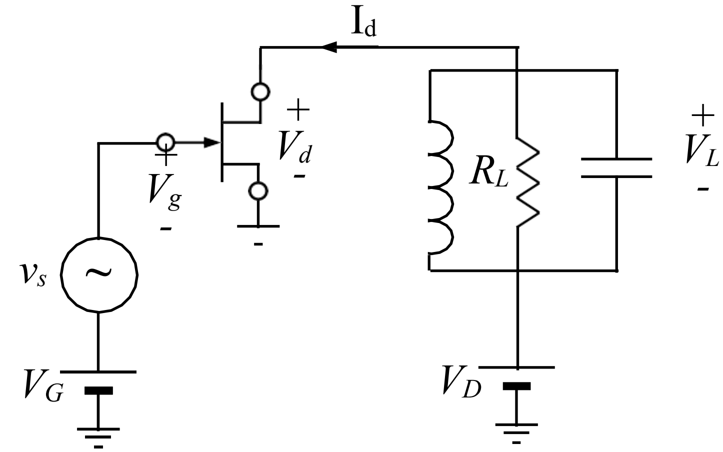
“First-order” design approach

Some fundamental notions

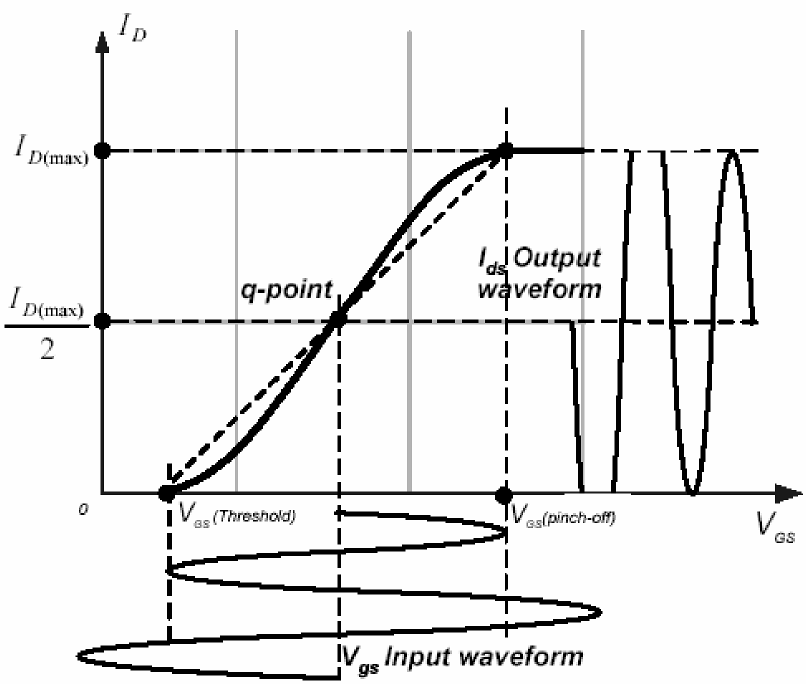
FUNDAMENTAL CONSIDERATIONS IN POWER AMPLIFIER DESIGN:

CLASS-A AMPLIFIERS

The circuit consists of a FET, a tuned circuit and a load, R_L.



Ideally, we assume that **the FET is an ideal transconductance** without resistive or reactive parasitics (the most important element is the controlled current source).



The bias current is I_d. The application of a RF excitation $v_s(t)$ to the gate generates an RF component of drain current, $\Delta I_d(t)$.

The RF component of the drain voltage, $\Delta V_d(t)$, is equal to the voltage drop across R_L, i.e.,

$$V_L(t) = \Delta V_d(t) = -\Delta I_d(t) R_L$$

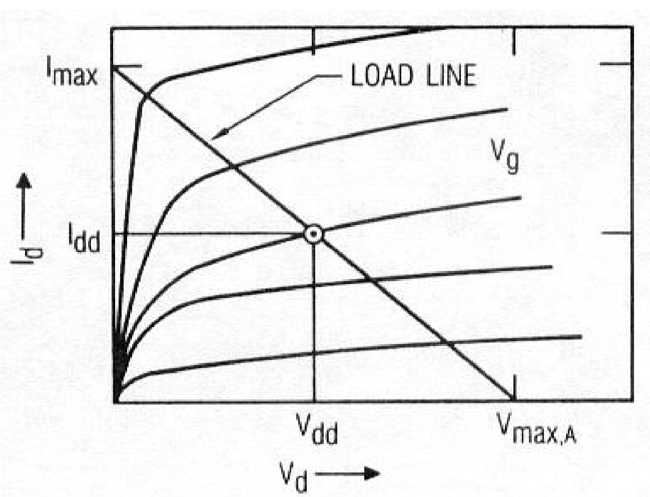
Since we wish to maximize the power delivered to the load, **we can maximize the excursion** of both $\Delta V_d(t)$ and $\Delta I_d(t)$, with

$$V_L(t) = V_{dd} \qquad I_L(t) = I_{dd}$$

The load line dictates that

$$R_L = \frac{V_{\max,A}}{I_{\max}} = \frac{V_{dd}}{I_{dd}}$$

Then, the drain voltage will vary **from 0 to 2*V_{dd}** (same for the current: from **0 to 2*I_{dd}**).



Under these conditions, the output power is equal to

$$P_L = 0.5\ V_L(t)\ I_L(t) = 0.5\ V_{dd}\, I_{dd}$$

$$\longrightarrow \quad P_{L\max} = \frac{1}{2}\left(\frac{1}{2}V_{\max,A}\right)\left(\frac{1}{2}I_{\max}\right) = \frac{1}{8}V_{\max,A}I_{\max}$$

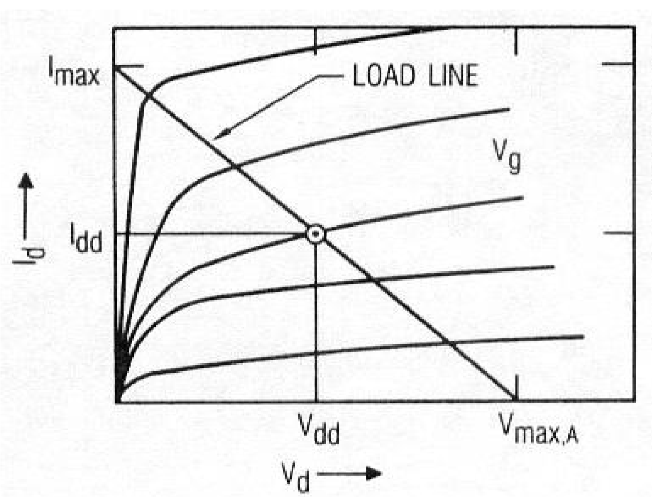
The dc current remains constant at I_{dd} at all excitation levels; therefore, by defining the dc power as

$$\{\ P_{dc} = V_{dd}\, I_{dd}\ \},$$

Thus, the dc-RF conversion efficiency is equal to

$$\eta_{dc} = \frac{P_L}{P_{dc}} = \frac{0.5 * V_{dd} * I_{dd}}{V_{dd} * I_{dd}} = 50\%$$

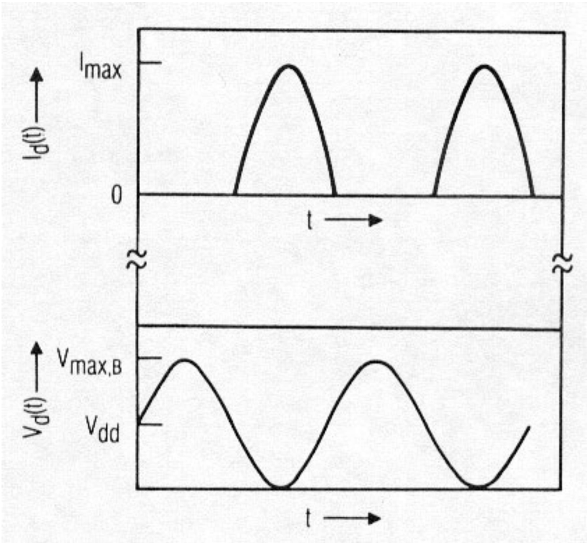
As expected, the maximum efficiency of ***Class-A amplifier*** is 50%.



FUNDAMENTAL CONSIDERATIONS IN POWER AMPLIFIER DESIGN:

CLASS-B AMPLIFIERS

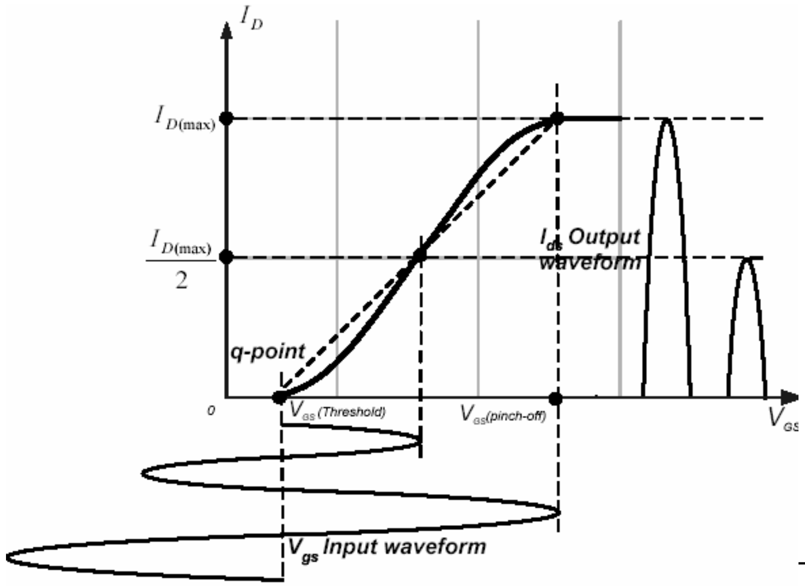
For such amplifiers, the transistor conducts during half of the input signal cycle. This is accomplished by biasing the transistor at cut-off.



$$I_{dc} = \frac{I_{\max}}{\pi} \qquad \rightarrow \qquad P_{dc} = V_{dd}\frac{I_{\max}}{\pi}$$

A tuned circuit allows only the fundamental component of $I_d(t)$ to pass. The power delivered to the load is

$$P_L = \frac{1}{2}\left|I_L(t)\right| * \left|V_L(t)\right| = \frac{1}{2} I_1 * \left|V_L(t)\right|$$



where I_1 is the magnitude of the fundamental component of the load current equal to $0.5{*}I_{max}$.

Thus, we have

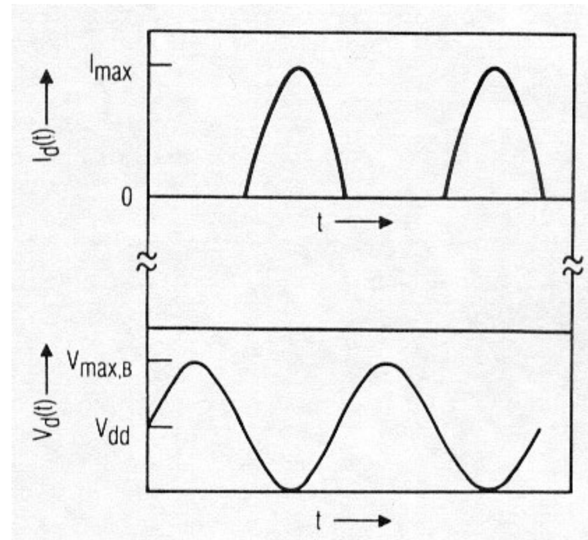
$$\left|V_L(t)\right| = \left|\Delta V_d(t)\right| = V_{dd}$$

$$\rightarrow \qquad P_L = \frac{1}{2} I_1 * \left|V_L(t)\right| = \frac{1}{2}\left(\frac{1}{2} I_{\max}\right) V_{dd} = \frac{1}{4} I_{\max} V_{dd}$$

FUNDAMENTAL CONSIDERATIONS IN POWER AMPLIFIER DESIGN:

CLASS-B AMPLIFIERS

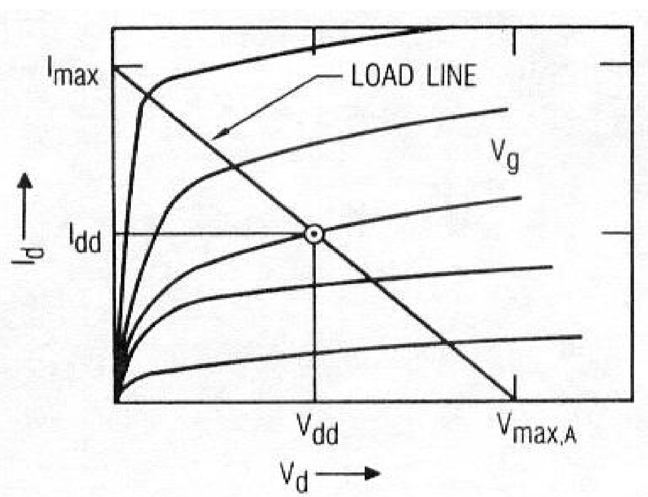
In order to achieve the maximum output power, the load resistance must be



$$R_L = \frac{V_{dd}}{I_1} = \frac{V_{dd}}{0.5 * I_{max}} = \frac{2\,V_{dd}}{I_{max}} = \frac{V_{max,B}}{I_{max}}$$

The dc-RF efficiency is

$$\eta_{dc} = \frac{P_L}{P_{dc}} = \frac{\pi}{4} = 0.78 = 78\%$$

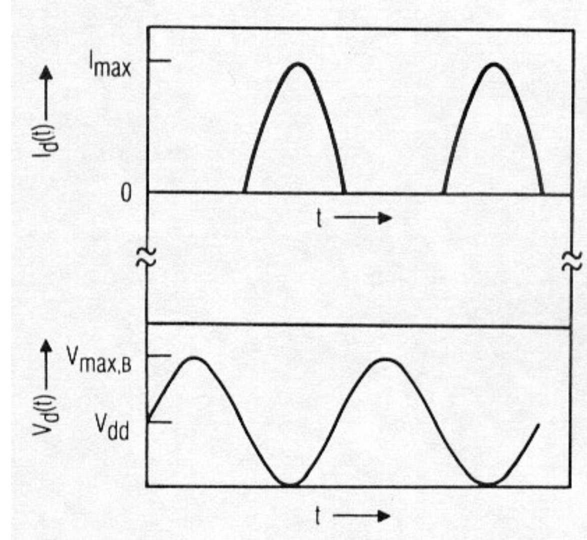


Since $V_{max,B} = 2V_{dd}$, the maximum output power for Class-B is

$$P_{L\,max} = \frac{1}{2}\left(\frac{1}{2}V_{max,B}\right)\left(\frac{1}{2}I_{max}\right) = \frac{1}{8}V_{max,B}I_{max}$$

which is the same as that of the Class-A amplifier if $V_{max,A} = V_{max,B}$.

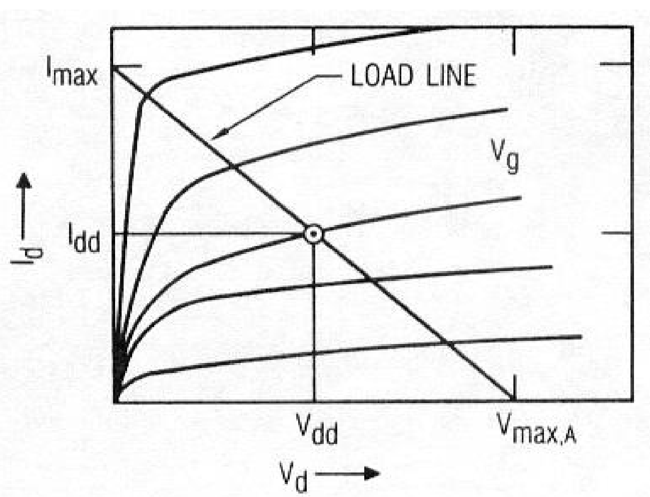
***Disadvantages** of the Class-B amplifier*

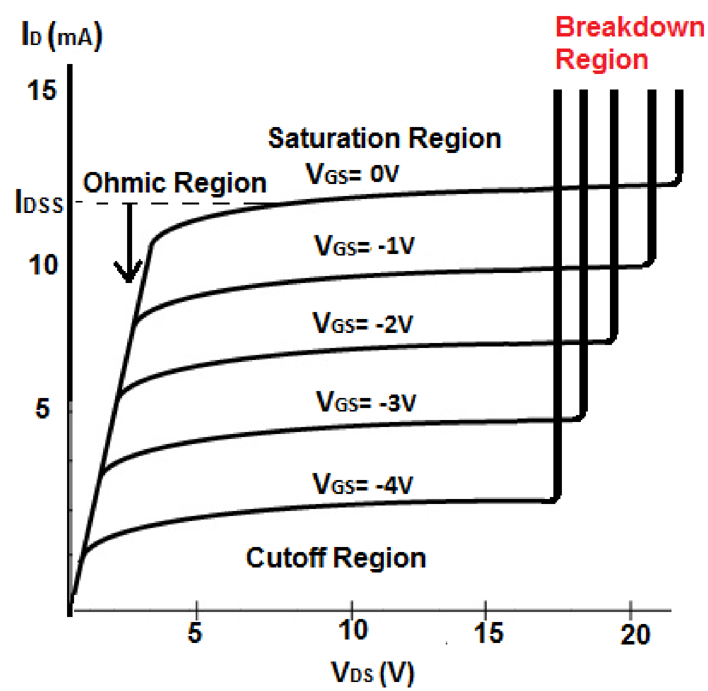


In practice, because the maximum drain voltage is limited by the gate-drain avalanche breakdown ($|V_{gd}| = V_{ds} + |V_{gs}|$), $V_{max,A}$ is **always greater** than $V_{max,B}$.

Accordingly, the maximum output power of Class-B amplifier is slightly lower that of a Class-A amplifier using the same device.

Class-B amplifiers have inherently lower gain than Class-A amplifiers.



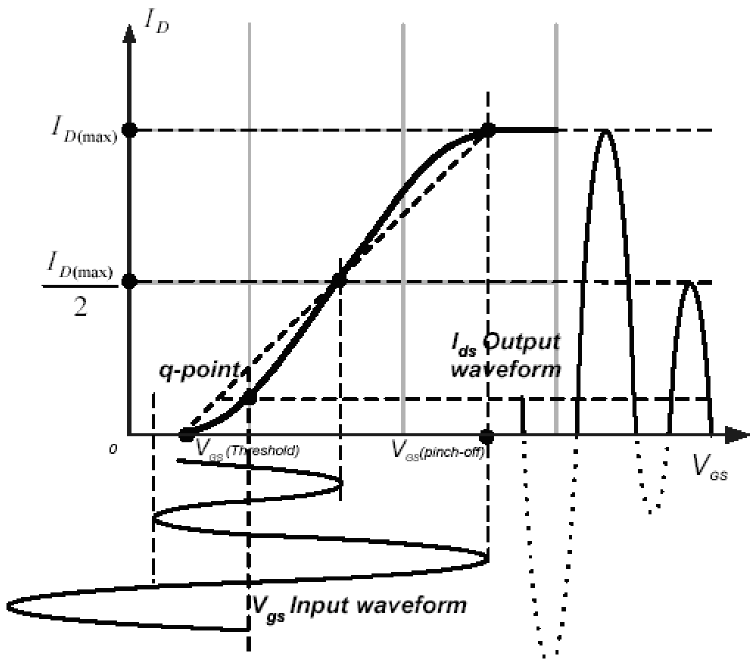


FUNDAMENTAL CONSIDERATIONS IN POWER AMPLIFIER DESIGN:

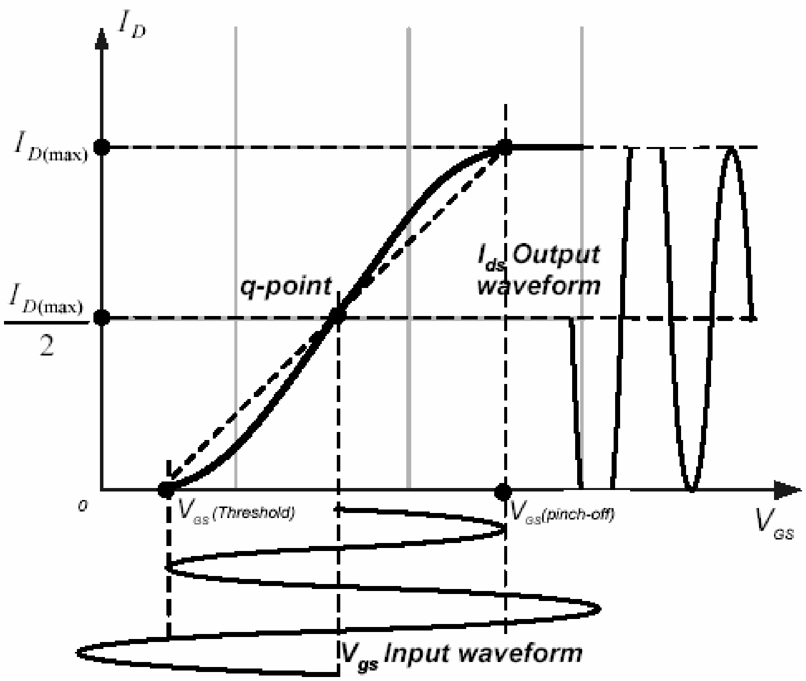
CLASS-AB AMPLIFIERS

For all the above reasons, we rarely operate power FETs in a true Class-B mode. So called Class-B amplifiers are usually biased near $0.1I_{dss}$ and actually operate in a mode between Class-B and Class-A.

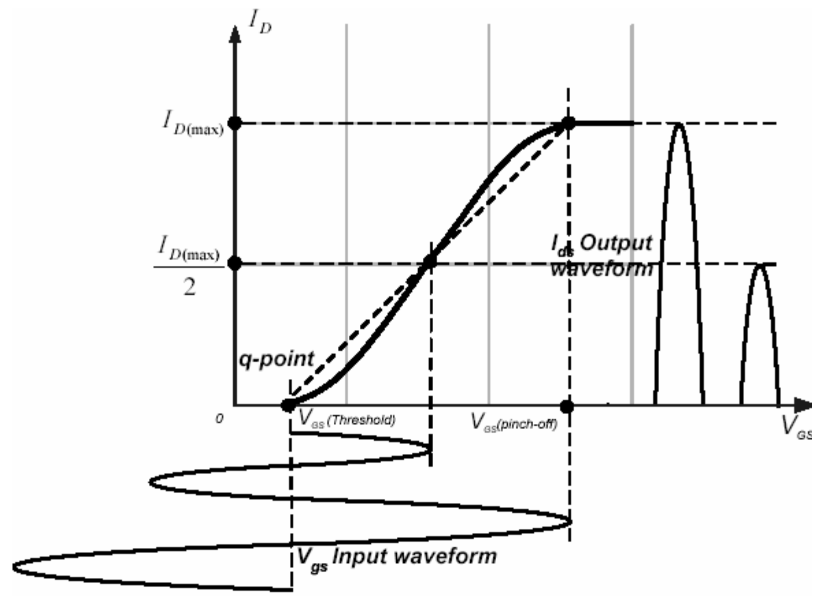
Conversely, Class-A amplifiers are often not operated in a classical Class-A mode. They are sometimes biased to a minimal current level and driven well into saturation. **Both types** of operation are called ***Class-AB***.



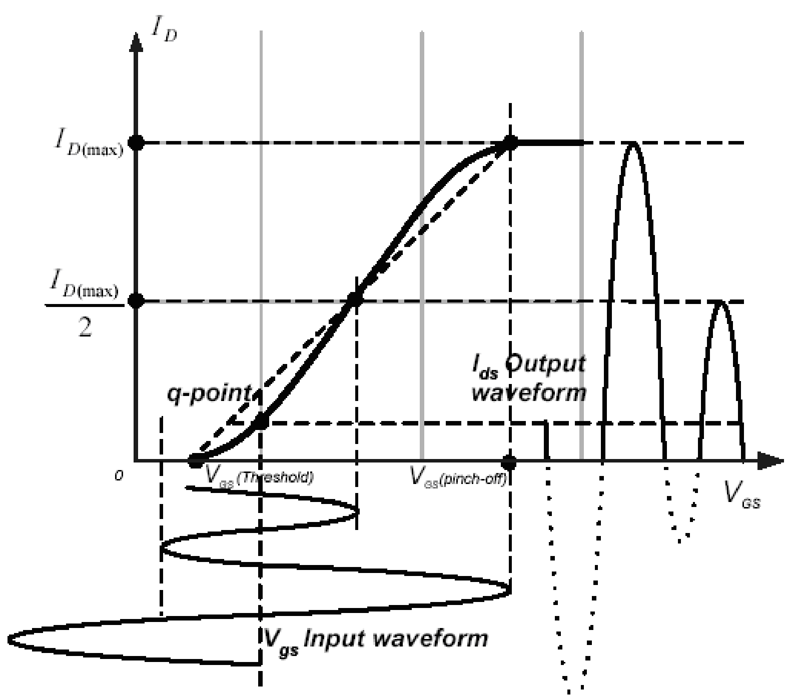
Class-AB amplifiers represent a **compromise** between the extremes of either class. Class-AB amplifiers usually have **better efficiency** than **Class-A amplifiers** and **better gain** than **Class-B amplifiers**.



Class-A



Class-B



Class-AB

FUNDAMENTAL CONSIDERATIONS IN POWER AMPLIFIER DESIGN:

CLASS-C AMPLIFIERS

In **Class-C**, the transistor is cut-off until the ac signal between the base/gate and emitter/source makes it conduct. The emitter (source) is grounded and the input sets the quiescent value of the base-to-emitter voltage (gate-to-source) at zero.

For the input ac signal, the input port is an open circuit and the input signal applied between the base (gate) and emitter (source) makes the transistor conduct.

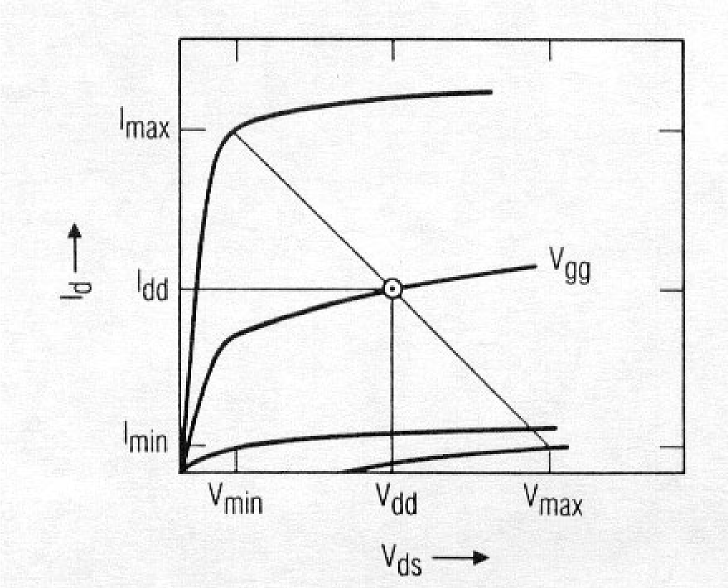
The conduction angle is less than 180^o, **about 140^o** for good efficiency and low harmonic content at the output.

The output-matching network must have **a high Q** value in order to suppress the harmonics and pass the amplified fundamental signal.

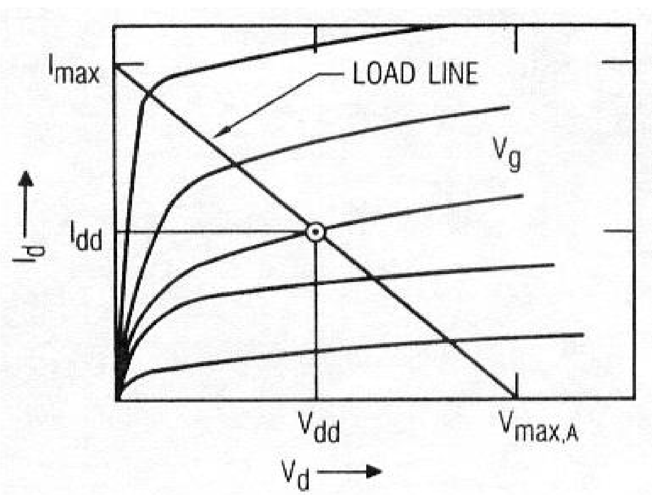
Efficiency is **better** in Class-C operation.

The “simplified” approach

APPROXIMATE APPROACH:

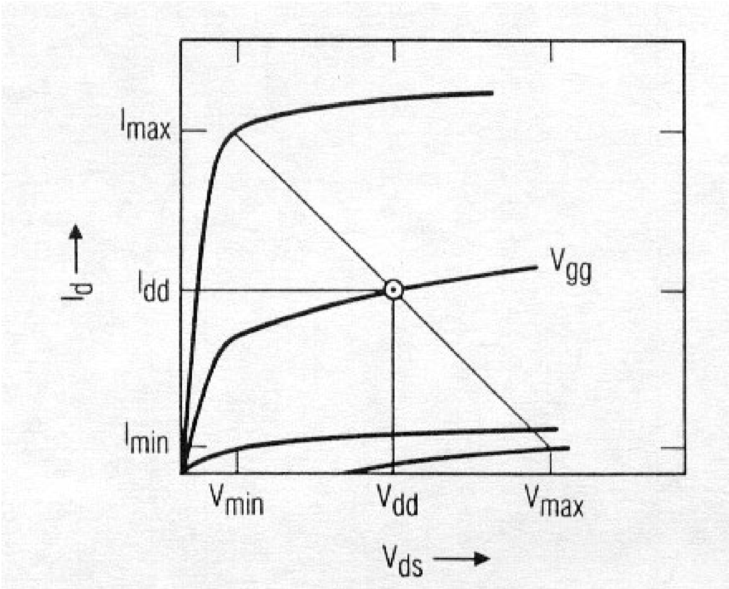


We base the design on the terminal I-V characteristics of a FET (i.e., using the terminal voltages V_{gs} and V_{ds} instead of the internal voltages V_g and V_d) as shown in the figure.



It is preferable to use the internal I-V curves, but they are not accessible through measurements, and recognizing that this initial design is **approximate**, we shall accept a plot of the FET's terminal as an **approximation** of the internal ones.

APPROXIMATE APPROACH:



To achieve so, we must take into account the limits on the drain voltage and current:

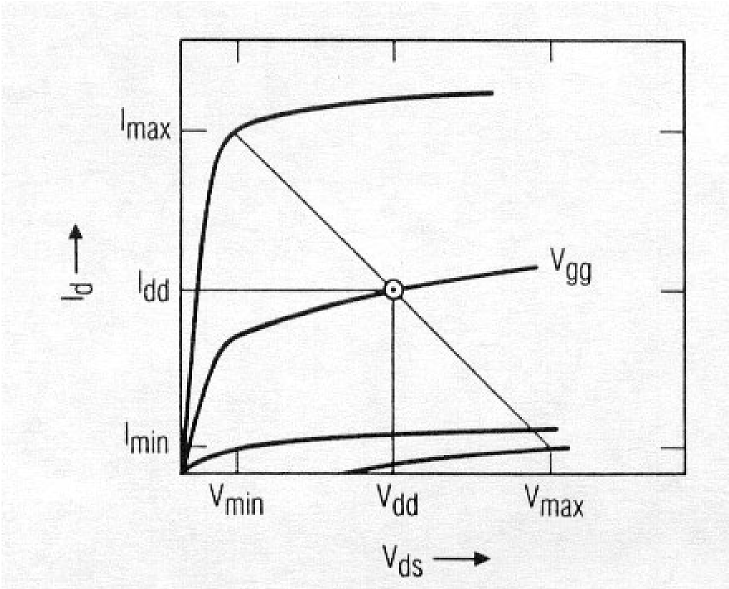
V_{min}: The minimum drain-source voltage, limited by the I-V curves knee at $V_g = 0.6$V ($V_{min} \approx 1.5$V).

V_{max}: The maximum drain-source voltage, fixed by the physical limitations of the transistor (in data sheets).

I_{min}: Because of the variation in V_t with V_d, and the gate-drain avalanche limitation, V_d cannot be driven to the point where $I_d = 0$. Thus there is a finite drain current I_{min} at V_{max}, the maximum drain-source voltage (in data sheets).

I_{max}: The maximum drain current, limited by the given transistor performance (in data sheets).

APPROXIMATE APPROACH:



Based on these values, the quiescent dc voltage V_{dd} is halfway between V_{max} and V_{min} and the quiescent dc current I_{dd} is halfway between I_{max} and I_{min}.

The load conductance is equal to the slope of the load line:

$$G_L = \frac{I_{\max} - I_{\min}}{V_{\max} - V_{\min}}$$

When an unpackaged FET is biased in its saturation region (for a packaged FET, the load admittance is more complex to determine due to the presence of parasitics, but the approach is still the same), the drain-source capacitance C_{ds} is the dominant component of the output admittance.

Because we wish to present a real load of conductance G_L to the terminals of the controlled source I_d, the susceptance of the load must resonate with C_{ds}.

Thus, **the initial estimate of the load admittance is**

$$Y_L = G_L - j\omega C_{ds}$$

APPROXIMATE APPROACH: DESIGN EXAMPLE

We will use an X–band power MESFET (gate width 2.4 mm and output power capability of 1 W) which equivalent circuit is shown below. The operating frequency is 10 GHz.

The I-V characteristics can be modeled using the drain-current Curtice cubic model

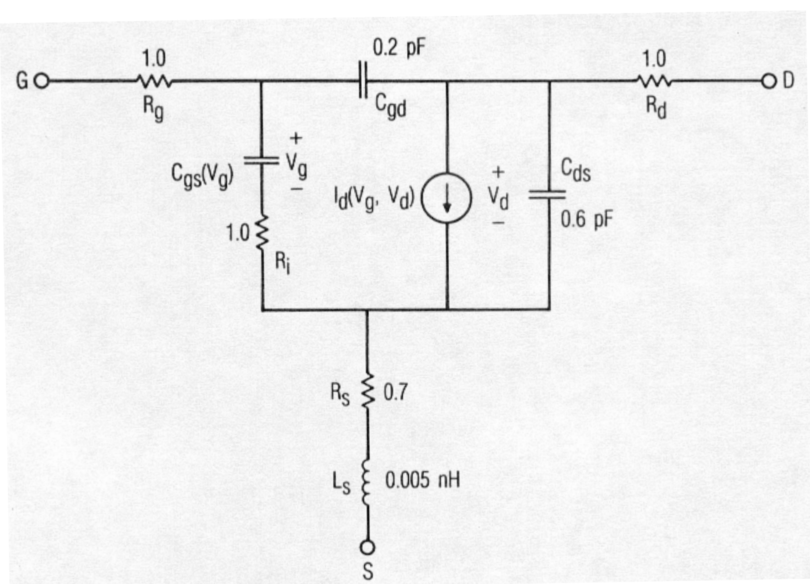
$$I_d = \left(a_o + a_1 V + a_2 V^2 + a_3 V^3\right)\tanh\left(\alpha V_{ds}\right)$$

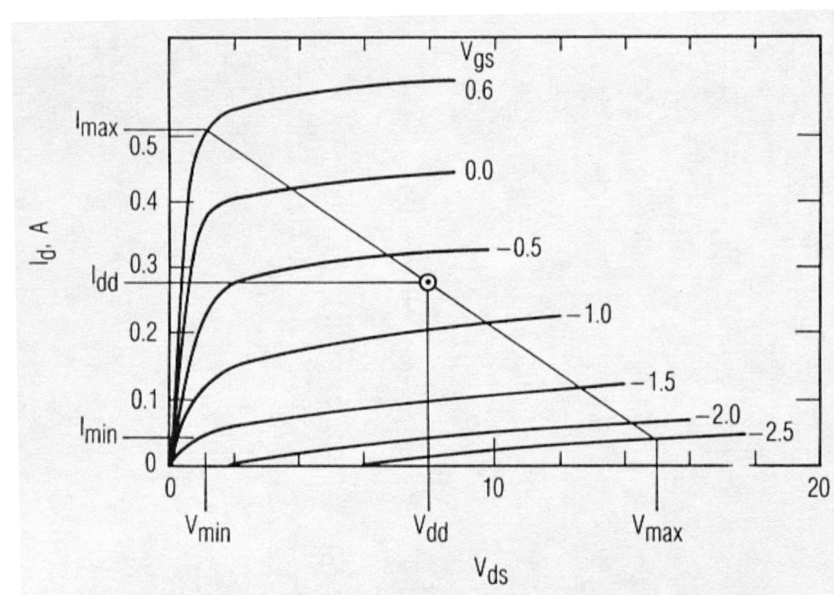
$$V = V_{gs}\left(1 + \beta\left(V_{dso} - V_{ds}\right)\right)$$

$a_0 =$ 0.5304 $a_1 =$ 0.2595

$a_2 =$ - 0.0542 $a_3 =$ - 0.0305

$a =$ 1.0 $V_t =$ - 2.5V

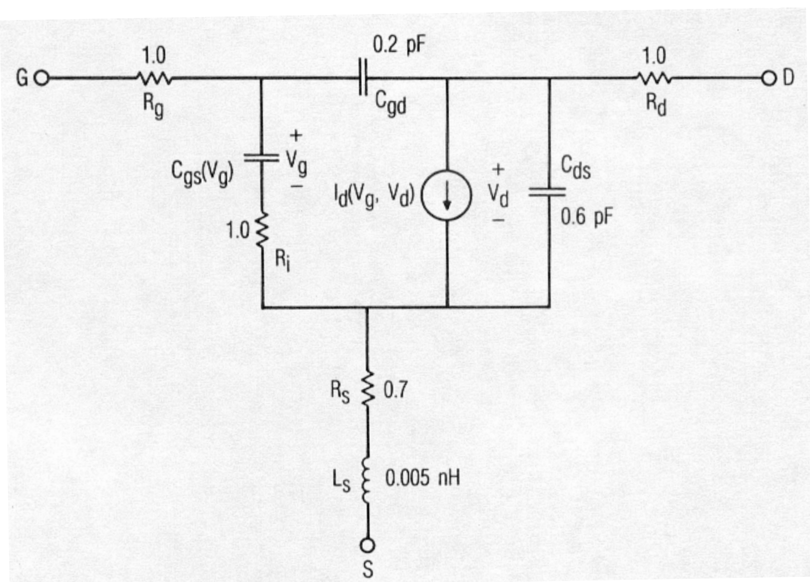


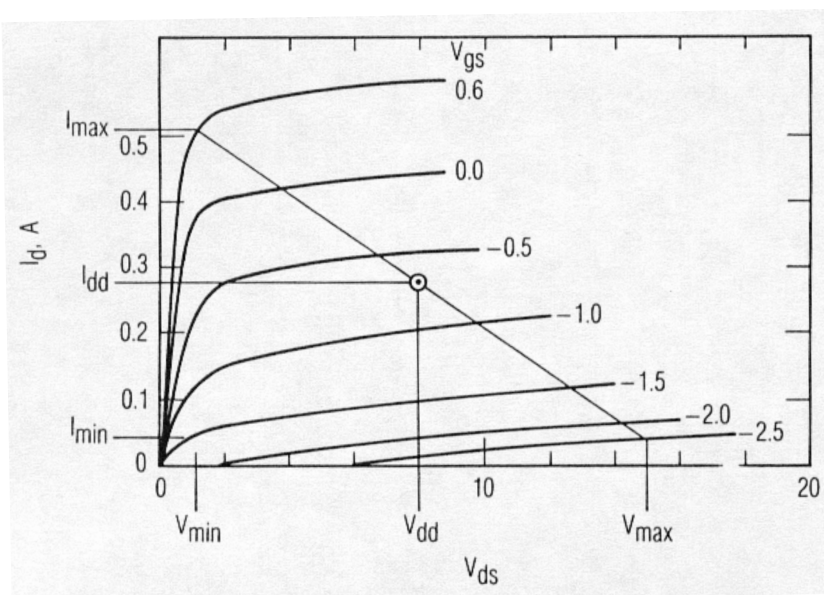


APPROXIMATE APPROACH: DESIGN EXAMPLE

The gate-source capacitance C_{gs} is modeled as a uniform doped Schottky barrier with $f = 0.7$V and $C_{gso} = 4.0$pF

$$C_{gs} = C_{gso}\left\{1 - \frac{V_j(t)}{\Phi}\right\}^{-1/2} = \frac{C_{gso}}{\sqrt{1 - \frac{V_j(t)}{\Phi}}}$$





APPROXIMATE APPROACH: Class-A amplifier

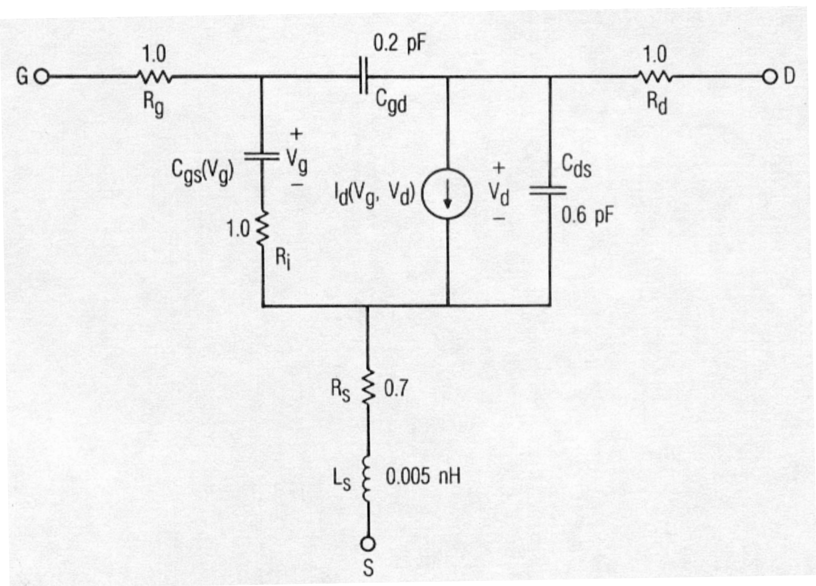
We determine first the load impedance. The load line is constructed under the constraint that $V_{dd} = 8.0\text{V}$ (and $I_{dd} = 270$ mA) and

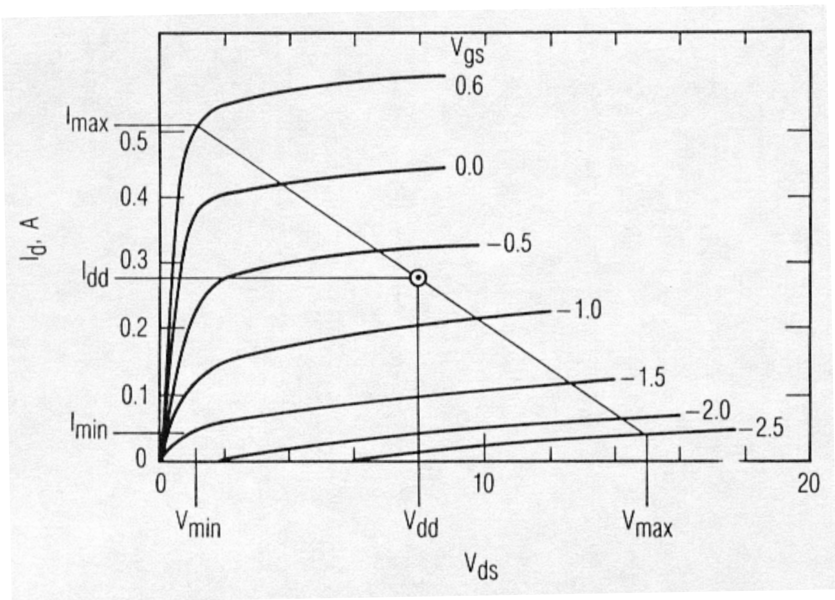
$$V_{max} = 14.7 \text{ V} \qquad V_{min} = 1.3 \text{ V} \qquad I_{max} = 500 \text{ mA} \qquad I_{min} = 40 \text{ mA}$$

which leads to

$$G_L = \frac{V_{\max} - V_{\min}}{I_{\max} - I_{\min}} = \frac{0.5 - 0.04}{14.7 - 1.3} = 0.034 \text{ S}$$

$$P_L = \frac{1}{2}\left[\frac{1}{2}\left(V_{\max} - V_{\min}\right)\right]\left[\frac{1}{2}\left(I_{\max} - I_{\min}\right)\right] = 0.771 \text{ W or } 28.9 \text{ dBm}$$





APPROXIMATE APPROACH: Class-A amplifier

$P_L = 28.9\,\text{dBm}$ (It is reasonable to anticipate a more realistic 1-dB compression point of 28 dBm).

$G_L = 0.034\,\text{S}$

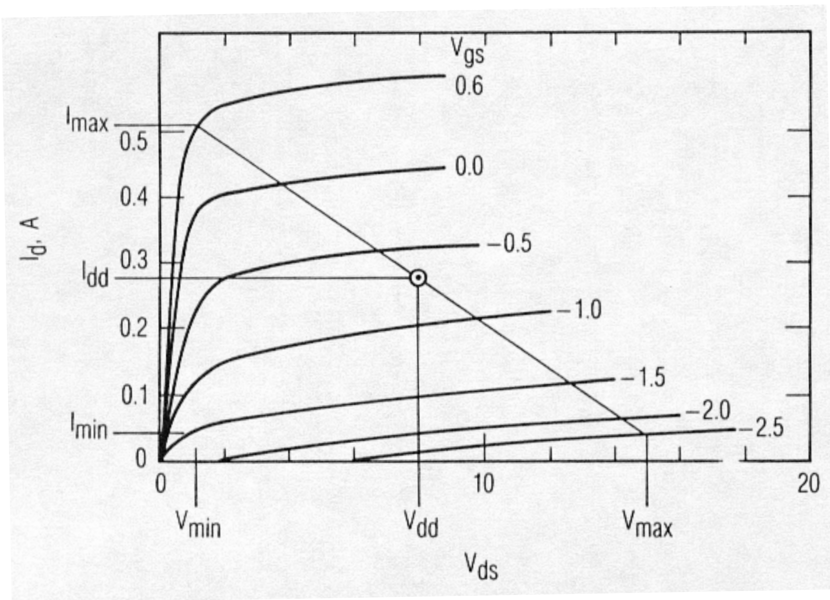
$$Y_L = G_L - j\omega C_{ds} = (0.034 - j0.05)\,\text{S}$$

$$Z_L = (9.3 + j13.5)\,\Omega$$

For the input impedance, we must find C_{gs} at the bias point (V_g = - 0.7 V):

$$C_{gs} = \frac{C_{gso}}{\sqrt{1 - \dfrac{V_j(t)}{\Phi}}} = \frac{4.0\,\text{pF}}{\sqrt{1 - \dfrac{-0.7}{0.7}}} = 2.8\,\text{pF}$$

The transconductance is found by differentiating I_d or from the I-V curves. Its value is 315 mS.



APPROXIMATE APPROACH: Class-A amplifier

$P_L = 28.9\ \text{dBm}$ $\quad Z_L = (9.3 + j13.5)\,\Omega$ $\quad G_L = 0.034\ \text{S}$

An adequate initial estimate of the input impedance is

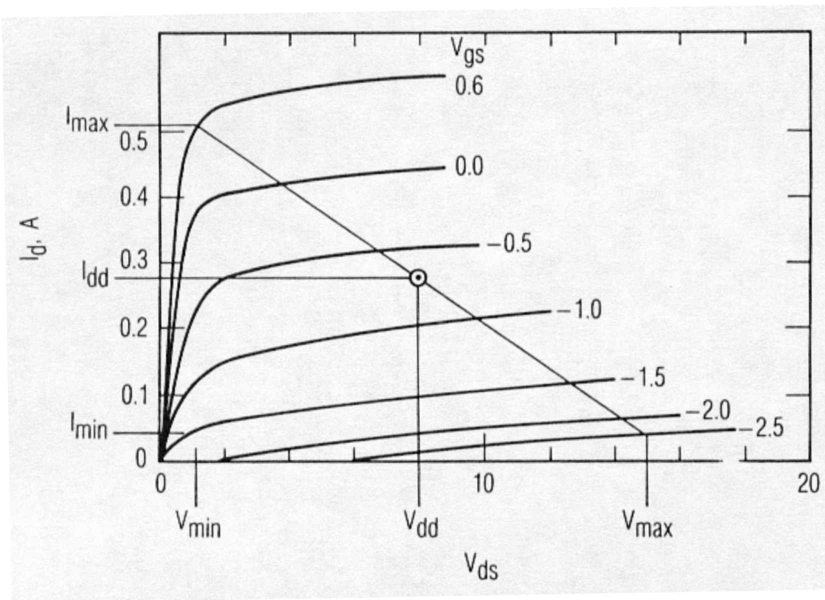
$$Z_{in} = R_g + R_i + R_s + \frac{<g_m> L_s}{C_{gs}(V_t)} + j\left[\omega L_s - \frac{1}{\omega C_{gs}(V_t)}\right]$$

The term $<g_m>$ is the transconductance averaged over the excitation cycle (*Note*: this value is around 20 % of the peak transconductance for a Class-B amplifier).

With these values, we get:

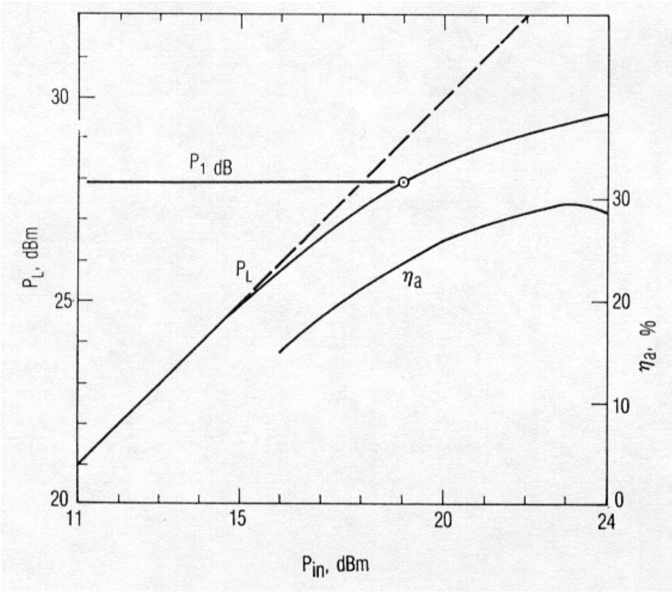
$$Z_{in} = Z_s^* = (2.3 - j\,4.3)\,\Omega$$

and a gain $G_t = 10$ dB.



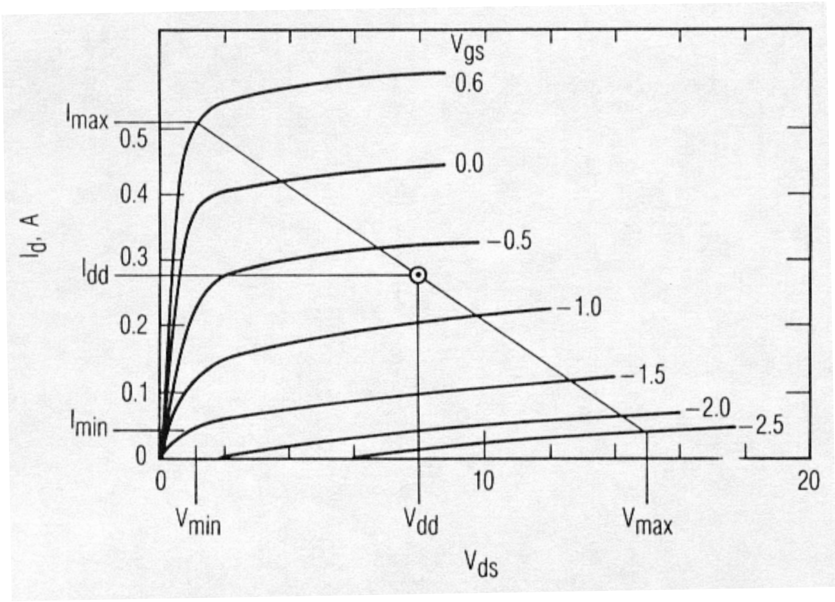
APPROXIMATE APPROACH: Class-A amplifier

$P_L = 28.9\ \text{dBm}$ $Z_L = (9.3 + j13.5)\,\Omega$ $Z_{in} = (2.3 - j4.3)\,\Omega$



Assuming that the matching networks present the optimum termination (short-circuit to the gate and drain at all harmonics), ***a harmonic-balance analysis*** shows that the approximate design is very good:

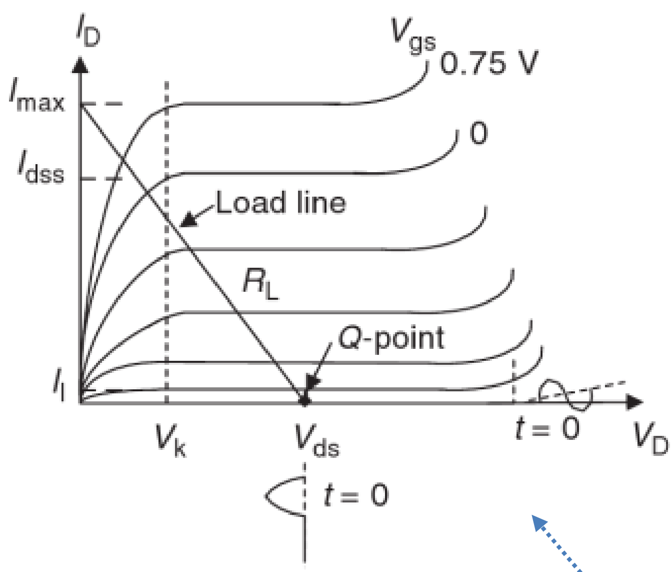
- output power at 1-dB compression point : 27.8 dBm
- output impedance : (9.0 + j12.0) Ω
- input impedance : (2.5 - j 4.5) Ω
- dc current at saturation : 290 mA
- dc power : 2.3 W.

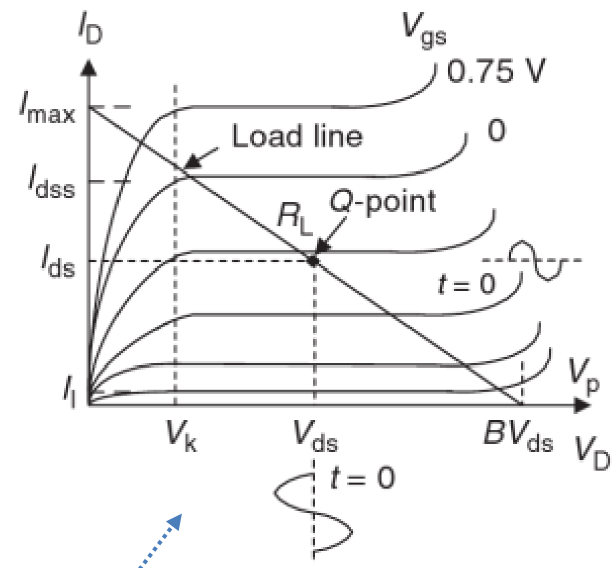


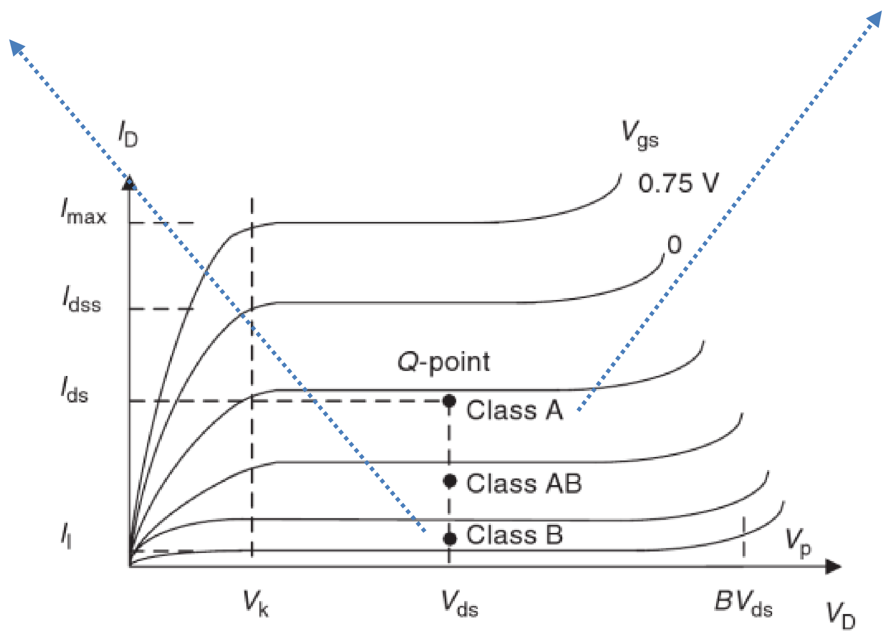
The output power and power added efficiency η_a are shown in the figure.

APPROXIMATE APPROACH: Class-B amplifier

We now design a Class-B amplifier by modifying the previous design and bias points.







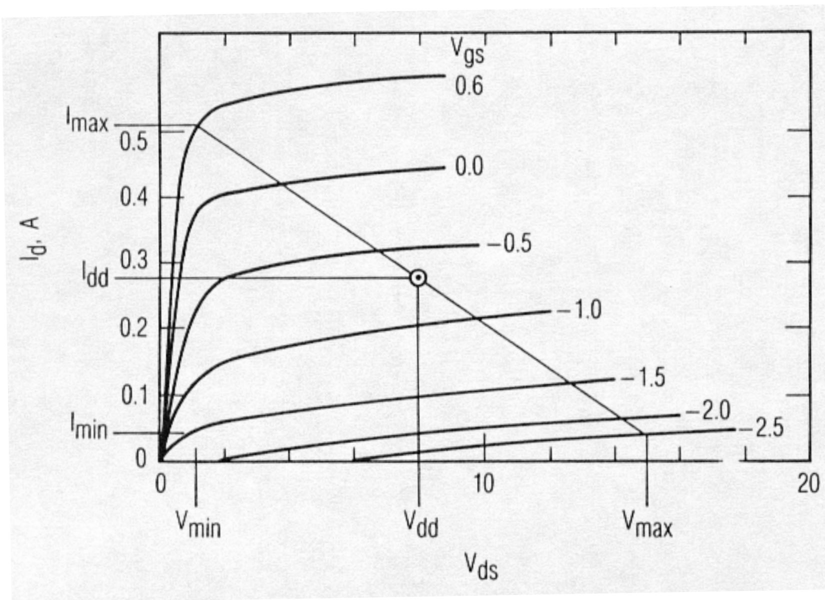
APPROXIMATE APPROACH: Class-B amplifier

The first modification is to reexamine the bias voltages: V_{dd} = 8 V and adjust the gate bias V_{gg} to –2.0 V, so that the quiescent current is 0.1*I_{dss} or 40 mA.

Assuming the harmonic source and load impedances are short-circuits, we can use the same load impedance for the Class-B as the Class-A. Note that the gate-source capacitance is lower because the voltage across C_{gs} is more negative.

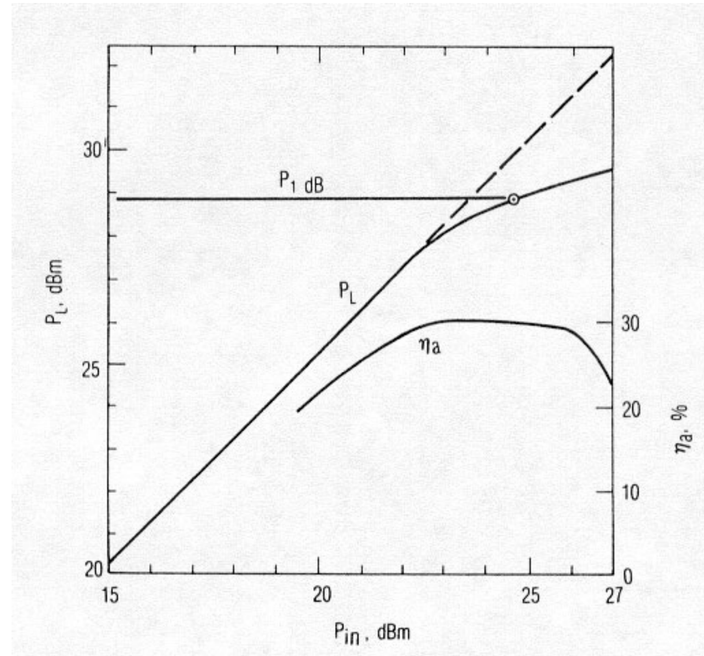
Consequently, the imaginary part of the input impedance should be greater. In fact, we obtain an initial estimate of

$$\{ Z_s^* = Z_{in} = (2.7 - j\ 7.6)\ \Omega \}.$$

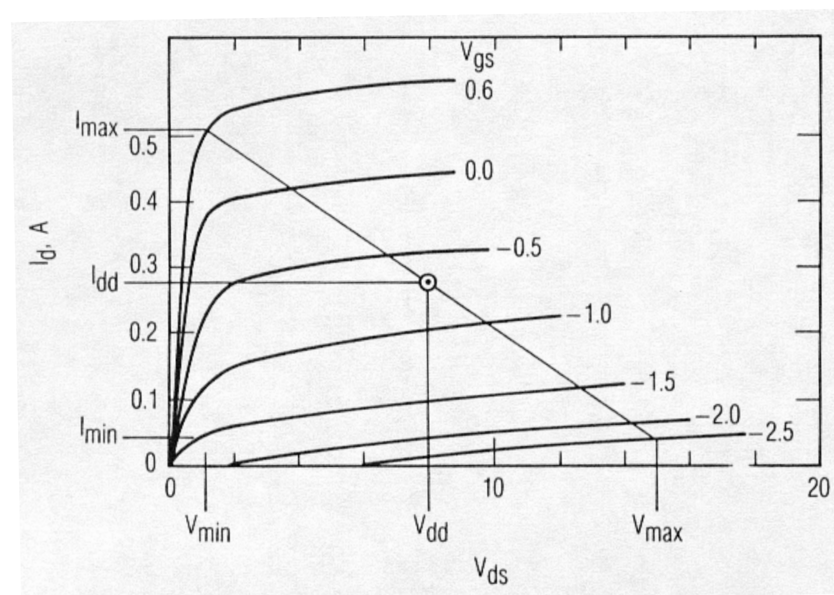


APPROXIMATE APPROACH: Class-B amplifier

Although the drain voltage and current ranges are slightly different (I_{min} = 0 in Class-B), we found, via the harmonic-balance analysis, that the Class-B load impedance is practically equal to the Class-A load and the input impedance is (2.7 – j 5.2) Ω.



P_L and η_a, as a function of P_{in}, are as in the figure.



Effect of nonzero harmonic terminations – Class-B design

We assumed that **<u>all</u>** harmonics are short-circuited. The effect of deviations from these ideal conditions is that the Class-B amplifier is more sensitive to harmonic terminations than the Class-A amplifier.

All the harmonic components of the drain current circulate in the terminating impedance, $Z_L(w)$. So, if $Z_L(w)$ has a nonzero real part at any harmonic frequency, there can be output power at that harmonic.

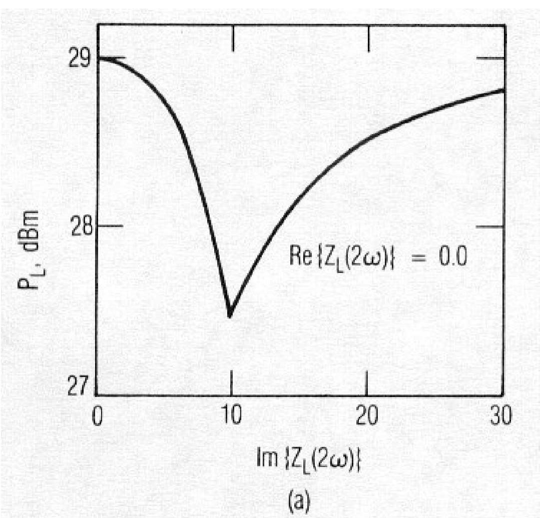
However, even if $Z_L(w)$ is purely reactive at some harmonic, V_d has a voltage component at that harmonic; this component can cause the fundamental output power to be reduced because output power is limited by the need to keep $\{ 0 < V_d(t) < V_{max} \}$.

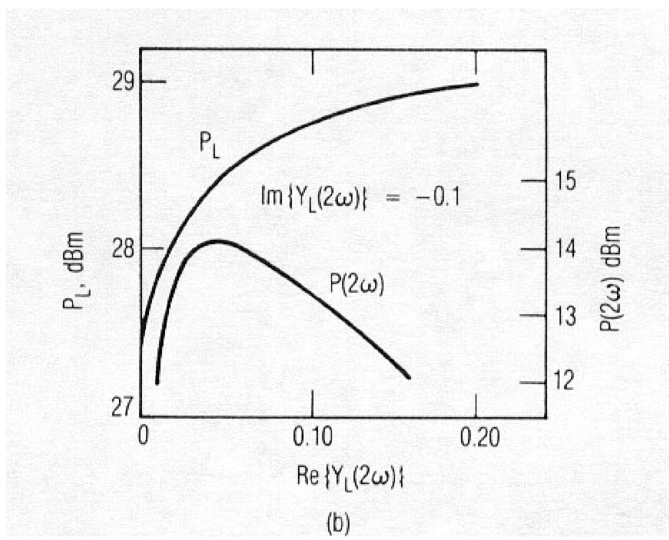
The addition of a harmonic component to $V_d(t)$ may force the fundamental component of $V_d(t)$ to be reduced, so that $V_d(t)$ remains within the prescribed limits.

Effect of nonzero harmonic terminations – Class-B design

However, we can understand that a relatively large value of C_{ds} (here 0.36 pF) will effectively short-circuit the channel at harmonics of the excitation frequency, making the amplifier relatively insensitive to harmonic termination effects.

Therefore, unless the second-harmonic termination resonates with C_{ds}, it has a minimal effect on fundamental output. However, if the load does resonate with C_{ds}, the results can have a serious effect on the performance of the amplifier. The figure shows the effect of a purely reactive second-harmonic termination on the output power of the Class-B amplifier.





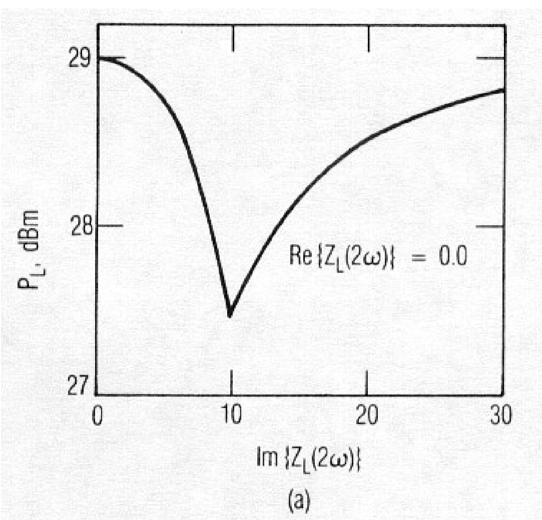
(a) effect of a purely reactive second-harmonic termination,
(b) fundamental and second-harmonic output powers as function of Re $(Y_L(2\omega))$.

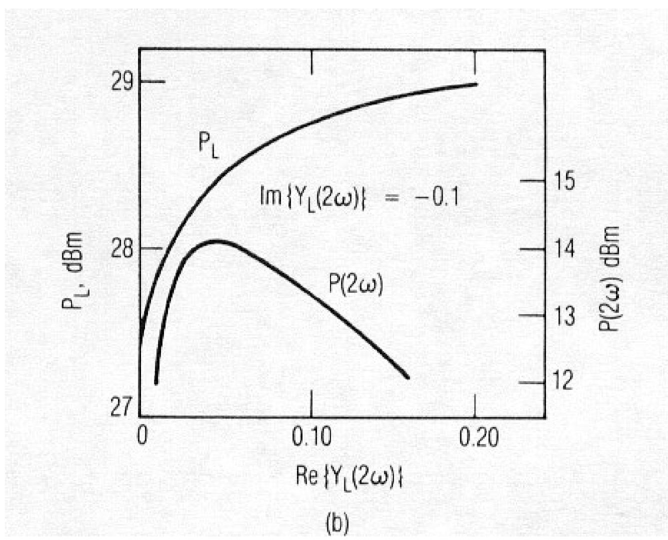
Effect of nonzero harmonic terminations – Class-B design

Here the output power decreases by about 1.5 dB. Figure (b) shows the effect of adding a conductance in parallel with the susceptance (-0.1 S) that resonates C_{ds}. Although the fundamental output power rises monotonically with Re $(Y_L(2\omega))$, the harmonic output peaks near Re $(Y_L(2\omega)) = 0.05$.

The worst case harmonic level is 14 dB below the fundamental output compared to 7 dB below the output in an ideal Class-B amplifier.

Note: This relative insensitivity to harmonic terminations resulted from the large value of C_{ds} in this particular device. Other MESFETs may have much lower values of C_{ds} and in these, the harmonic terminations may have much more significant effects.





(a) effect of a purely reactive second-harmonic termination,
(b) fundamental and second-harmonic output powers as function of Re $(Y_L(2\omega))$.

Thank you !

End of Chapter 4-3