TSEK03: Radio Frequency Integrated Circuits (RFIC)

Lecture 3a: Background

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Background: Overview

- Razavi:
  - Chapter 2.2 Effects of nonlinearity (mostly repetition from TSEK02)
  - Chapter 2.5 Matching
  - Chapter 2.6 Scattering parameters

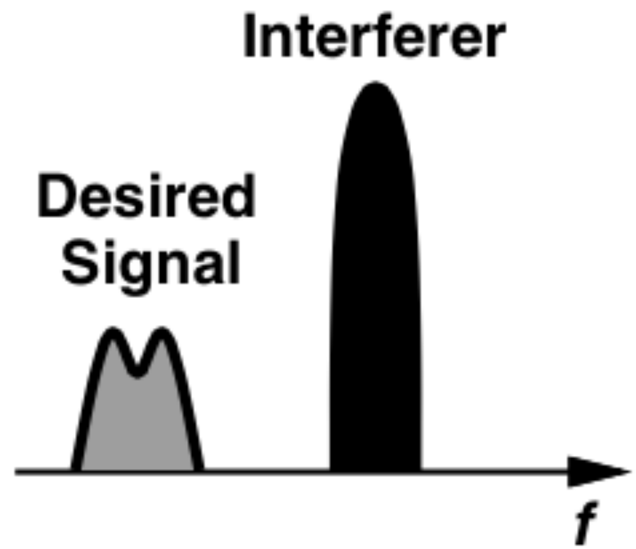
- Lee:
  - Chapter 7 Smith chart and s-parameters

2.2 Linearity

- For a nonlinear device:

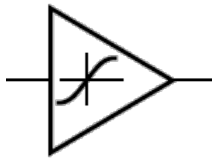
$$i(V_{DC} + v) \approx a_0 + a_1 v + a_2 v^2 + a_3 v^3 + \dots$$

- When strong signals are received, the LNA should remain linear
- Typically, weak signals are received in the presence of a strong interference. Linearity is important to suppress intermodulation distortion.



2.2.1 Harmonic Distortion

- Consider a nonlinear system

x(t)  $y(t) = \alpha_1 V_{in} + \alpha_2 V^2_{in} + \alpha_3 V^3_{in} + \ldots$

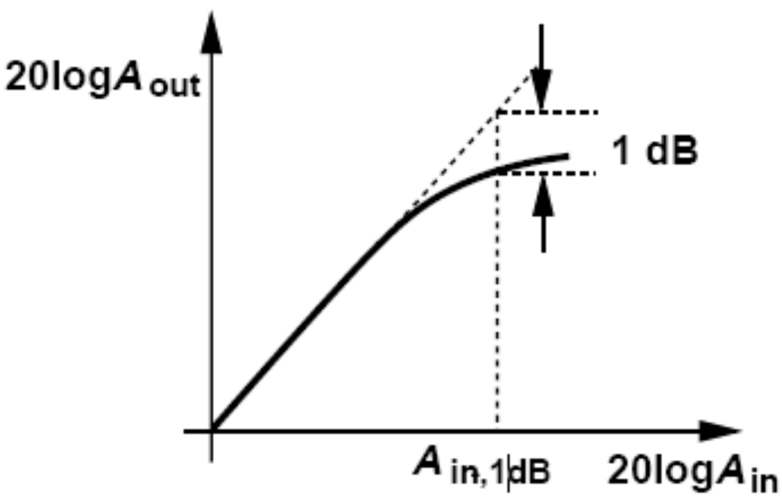
Let us apply a <u>single-tone</u> ($A\cos\omega t$) to the input and calculate the output:

$$
\begin{aligned}
y(t) &= \alpha_1 A \cos \omega t + \alpha_2 A^2 \cos^2 \omega t + \alpha_3 A^3 \cos^3 \omega t \\
&= \alpha_1 A \cos \omega t + \frac{\alpha_2 A^2}{2}(1 + \cos 2\omega t) + \frac{\alpha_3 A^3}{4}(3 \cos \omega t + \cos 3\omega t) \\
&= \frac{\alpha_2 A^2}{2} + \left(\alpha_1 A + \frac{3\alpha_3 A^3}{4}\right) \cos \omega t + \frac{\alpha_2 A^2}{2} \cos 2\omega t + \frac{\alpha_3 A^3}{4} \cos 3\omega t.
\end{aligned}
$$

DC | **Fundamental** | **Second Harmonic** | **Third Harmonic**

2.2.2 Gain compression

- If sign of α_1 and α_3 are opposite then the point in which the output falls below its ideal value by 1 dB is called 1-dB compression point or P-1dB:

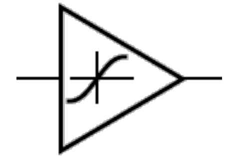


$$20 \log \left| \alpha_1 + \frac{3}{4} \alpha_3 A_{in,1dB}^2 \right| = 20 \log |\alpha_1| - 1 \text{ dB}.$$

$$A_{in,1dB} = \sqrt{0.145 \left| \frac{\alpha_1}{\alpha_3} \right|}.$$

The P-1dB point correlates well to loss of linear behavior, getting out-of-spec in standards (EVM, ACPR, etc.) so for linear applications, operation beyond this point is useless.

2.2.4 Intermodulation



- If a two-tone signal is applied to a non-linear device:

  $v = A[\cos(\omega_1 t) + \cos(\omega_2 t)]$

  $i(V_{DC} + v) \approx c_0 + c_1 v + c_2 v^2 + c_3 v^3 + \ldots$

- **By combining these equations we get several tones:**
  - **DC and fundamental tones**

$$(c_0 + c_2 A^2) + (c_1 A + \frac{9}{4} c_3 A^3)[\cos(\omega_1 t) + \cos(\omega_2 t)]$$

  - **Second and third harmonic terms**

$$(\frac{c_2 A^2}{2})[\cos(2\omega_1 t) + \cos(2\omega_2 t)] + (\frac{c_3 A^3}{4})[\cos(3\omega_1 t) + \cos(3\omega_2 t)]$$

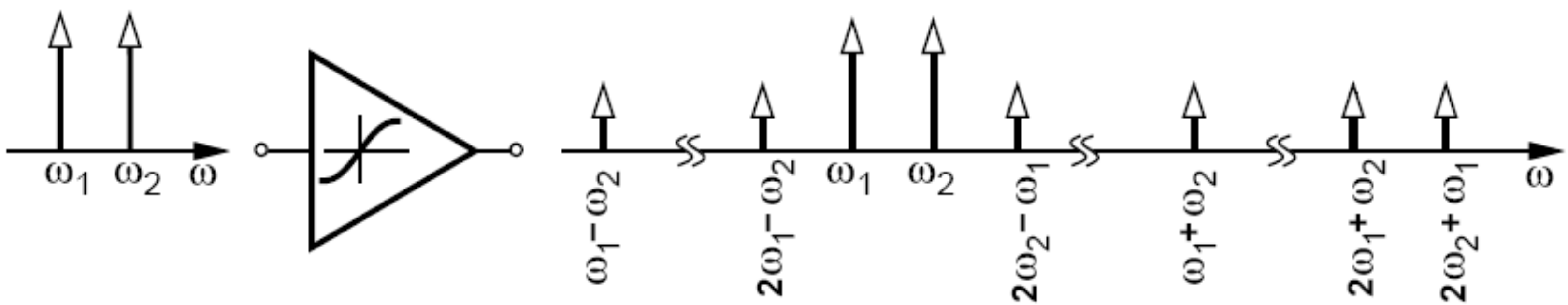
  - **Second order intermodulation (IM) products**

$$(\frac{c_2 A^2}{2})[\cos(\omega_1 + \omega_2)t + \cos(\omega_1 - \omega_2)t]$$

  - **Third order IM products**

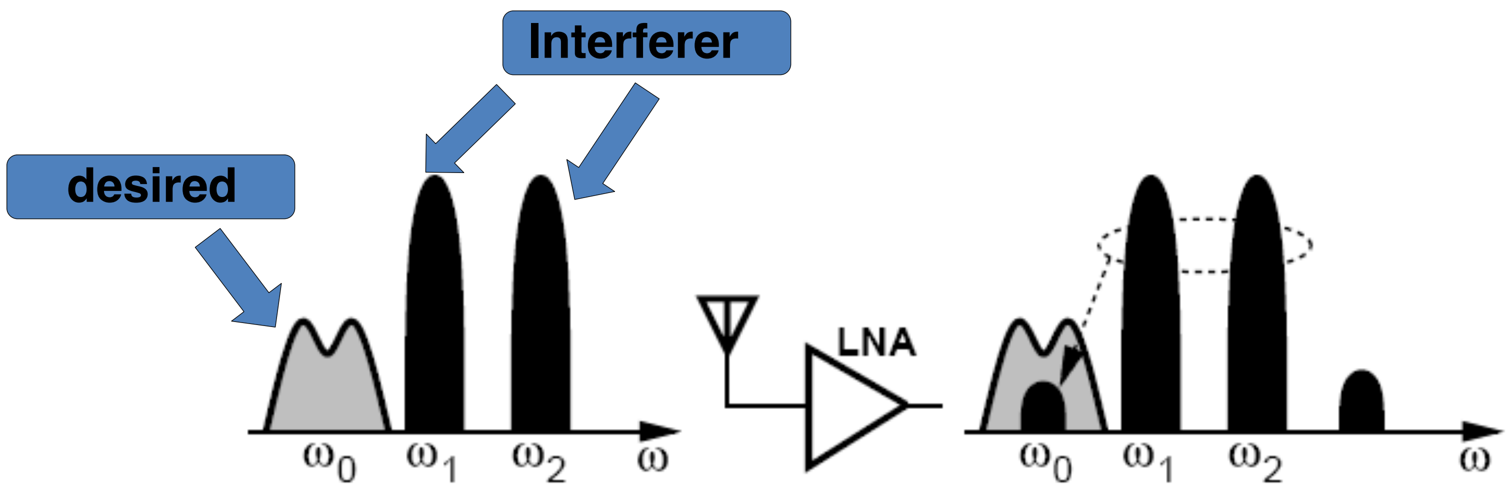
$$(\frac{3 c_3 A^3}{4})[\cos(2\omega_1 + \omega_2)t + \cos(2\omega_1 - \omega_2)t + \cos(\omega_1 - 2\omega_2)t + \cos(\omega_1 + 2\omega_2)t]$$

Intermodulation



Intermodulation

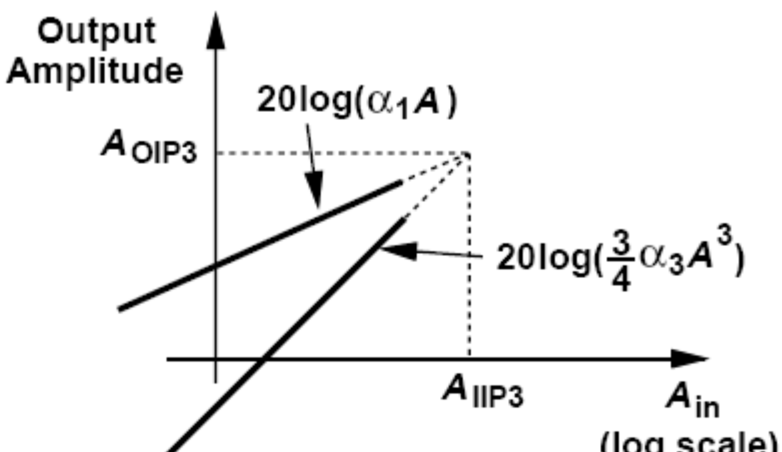
- Typically, weak signals are received in the presence of a strong interference. This might cause intermodulation distortion.



$$2\omega_1 - \omega_2 = \omega_0$$

Third-Order Intercept Point (IP3)

- Input-referred third-order intercept point (IIP3) is calculated by setting the IM3 products equal to the amplitude of the fundamental tone (A_1=A_2=A):



$$|\alpha_1 A_{IIP3}| = \left|\frac{3}{4}\alpha_3 A_{IIP3}^3\right|$$

$$A_{IIP3} = \sqrt{\frac{4}{3}\left|\frac{\alpha_1}{\alpha_3}\right|}.$$

$$\frac{A_{IIP3}}{A_{1dB}} = \sqrt{\frac{4}{0.435}} \approx 9.6 \text{ dB}.$$

IP3: measurement

- Circuits are non-linear at high power levels
- Measured at low power levels (check the slopes!), extrapolate!

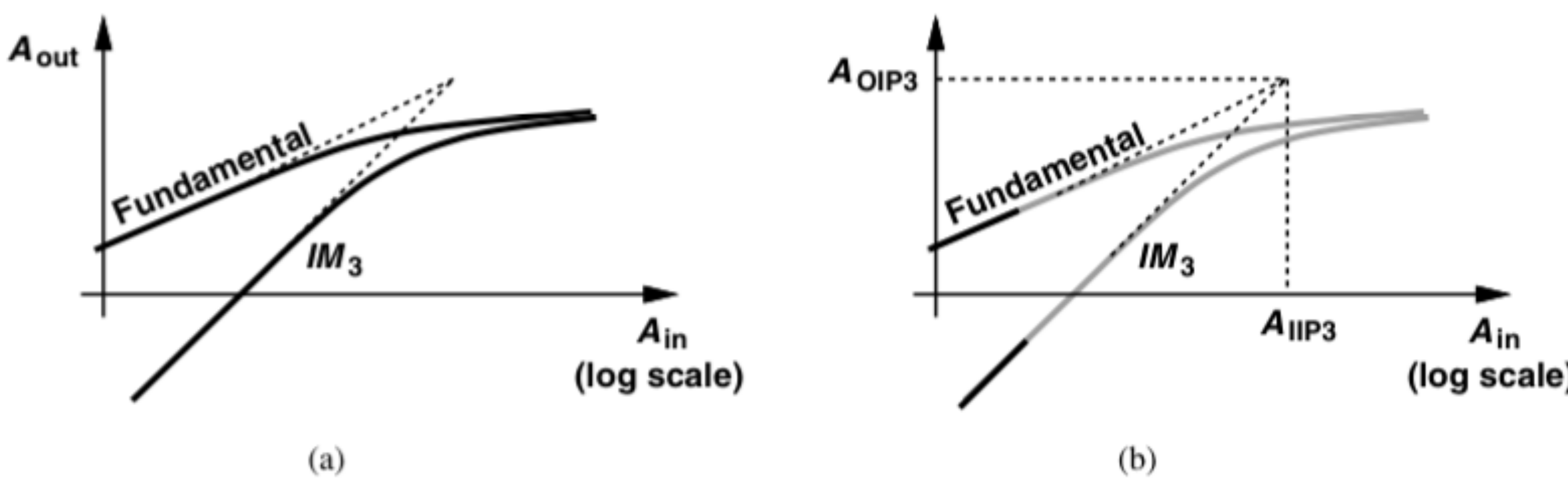
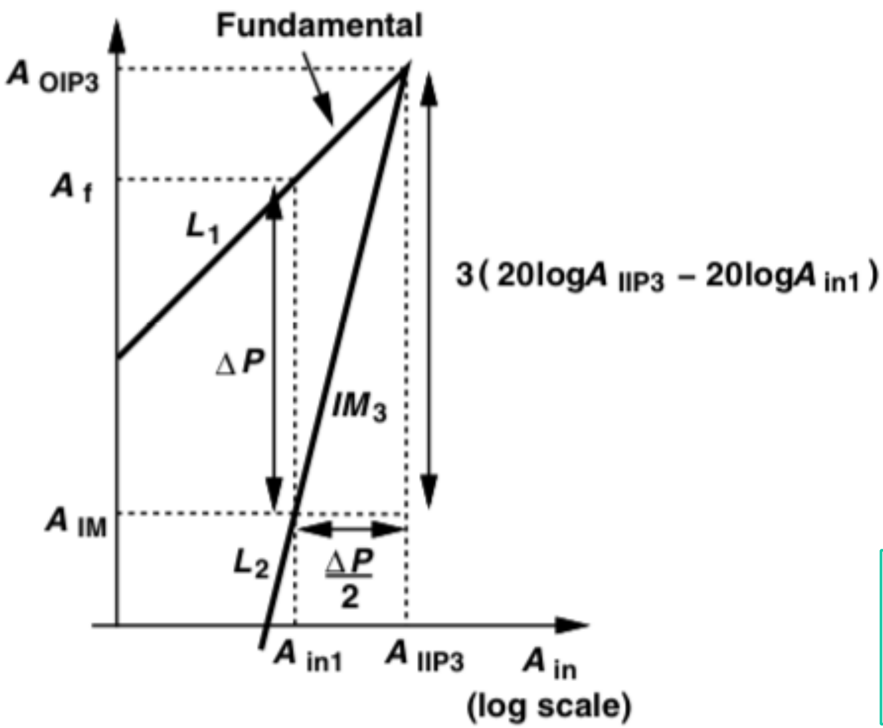
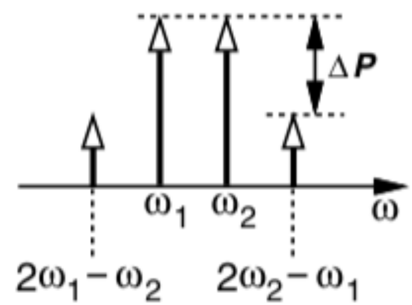


Figure 2.21 *(a) Actual behavior of nonlinear circuits, (b) definition of IP_3 based on extrapolation.*

IP3: measurement

- Shortcut technique if slopes are OK.

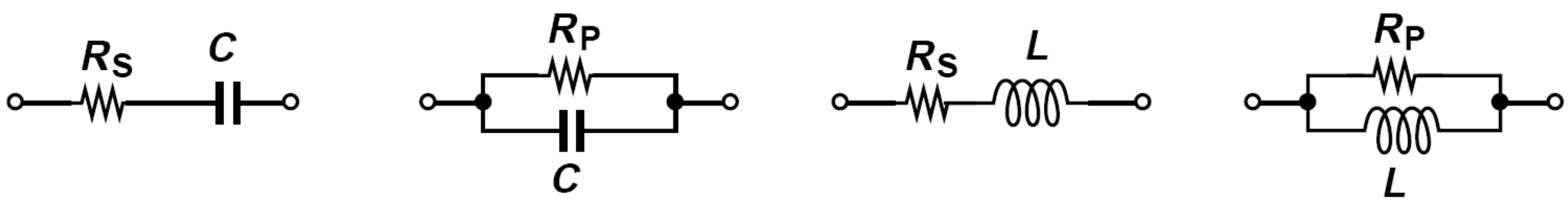




$$IIP_3\big|_{dBm} = \frac{\Delta P\big|_{dB}}{2} + P_{in}\big|_{dBm}$$

2.5 Passive impedance transformation

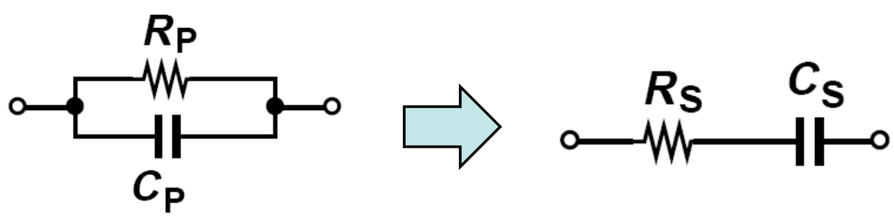
- a.k.a. *Matching networks*
- 2.5.1 Quality Factor, Q, indicates how close to ideal an energy-storing device is.



$$Q_S = \frac{\frac{1}{C\omega}}{R_S} \qquad Q_P = \frac{R_P}{\frac{1}{C\omega}} \qquad Q_S = \frac{L\omega}{R_S} \qquad Q_P = \frac{R_P}{L\omega}$$

Parallel-to-series conversion

- Series-to-Parallel Conversion: will retain the value of the capacitor but raises the resistance by a factor of Q_S^2
- Parallel-to-Series Conversion: will reduce the resistance by a factor of Q_P^2



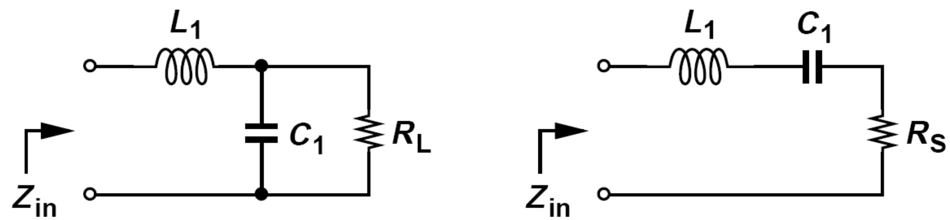
$$R_S = \frac{R_P}{Q_P^2}$$

$$C_s = C_P$$

valid if $Q_S^2 >> 1$
(”relative accurate”)

Basic matching networks

- Load resistance transformed to a lower value ($Z_{in} < R_L$):



(a) (b)

$$Z_{in}(j\omega) = \frac{R_L(1 - L_1C_1\omega^2) + jL_1\omega}{1 + jR_LC_1\omega}$$

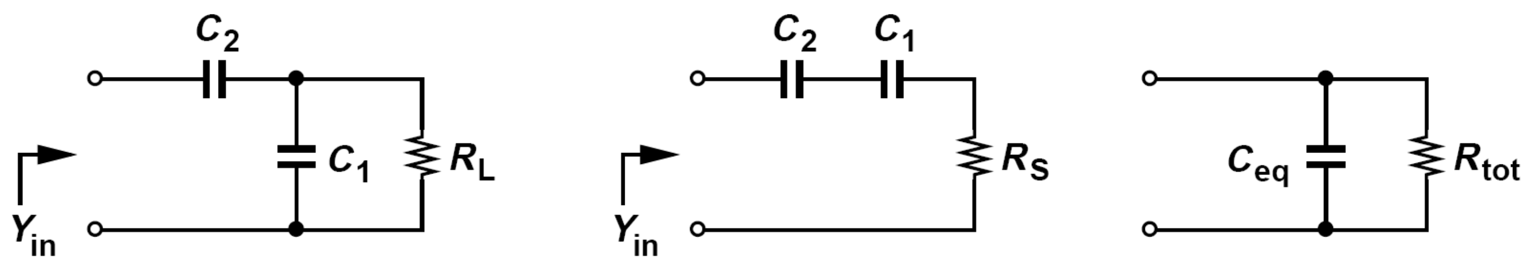
$$Re\{Z_{in}\} = \frac{R_L}{1 + R_L^2C_1^2\omega^2} = \frac{R_L}{1 + Q_P^2},$$

$$L_1 = \frac{R_L^2C_1}{1 + R_L^2C_1^2\omega^2} = \frac{R_L^2C_1}{1 + Q_P^2}.$$

$$Q_P^2 \gg 1$$

$$Re\{Z_{in}\} \approx \frac{1}{R_LC_1^2\omega^2}$$
$$L_1 = \frac{1}{C_1\omega^2}.$$

Transfer a resistance to a higher value



$$Q^2 \gg 1 \qquad R_S \approx [R_L(C_1\omega)^2]^{-1} \qquad C_S \approx C_1$$

Note that any imaginary component (often capacitance) must first be cancelled by an inductor at the input.

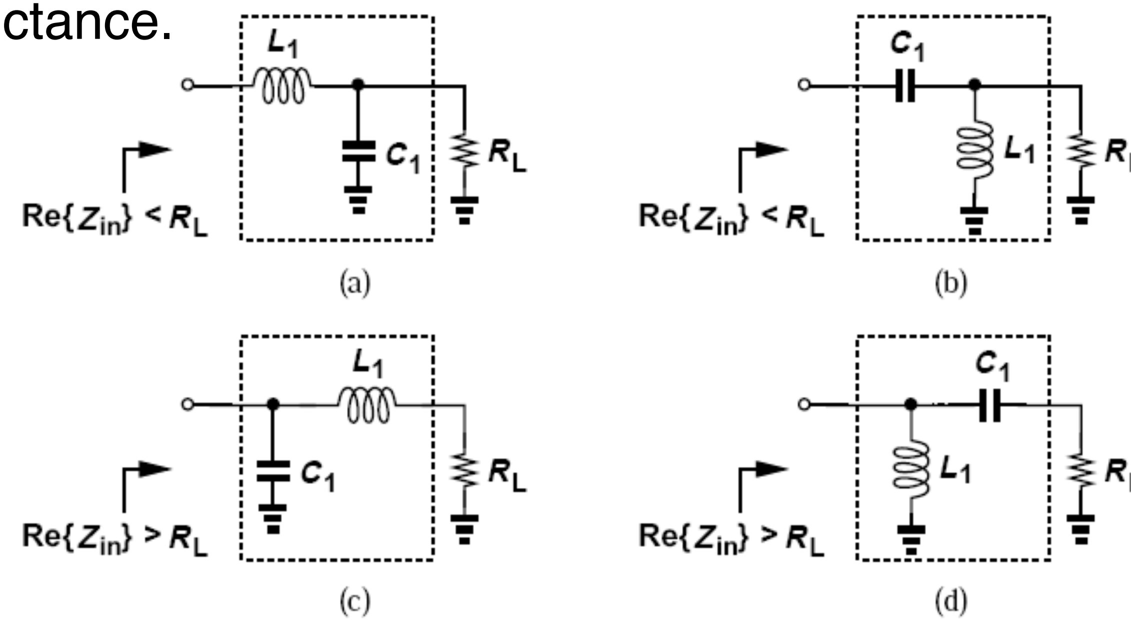
L sections

a) C_1 transforms R_L to a smaller series value and L_1 cancels C_1.

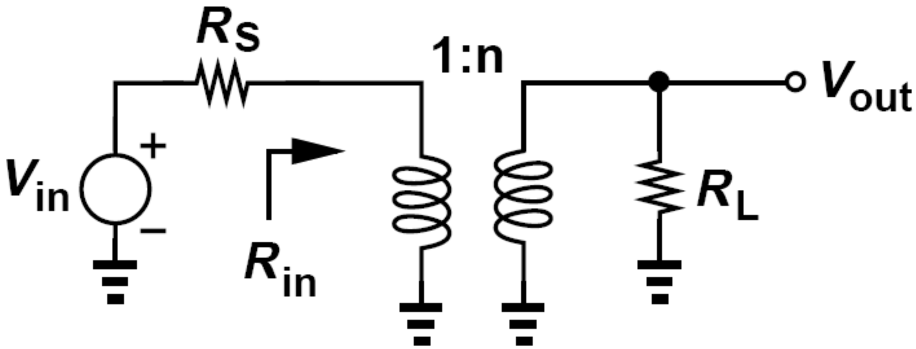
b) L_1 transforms R_L to a smaller series value while C_1 resonates with L_1.

c) L_1 transforms R_L to a larger parallel value and C_1 cancels the resulting parallel inductance.

d) same as c)

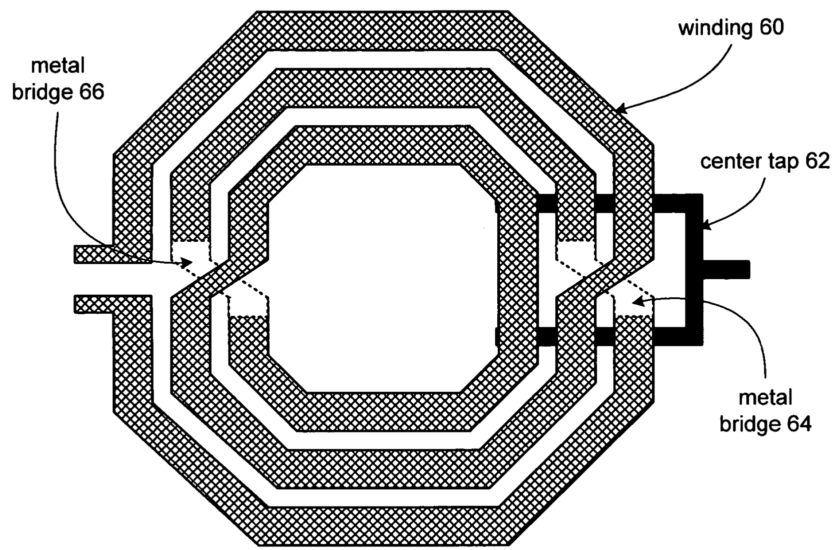


Impedance matching by transformers



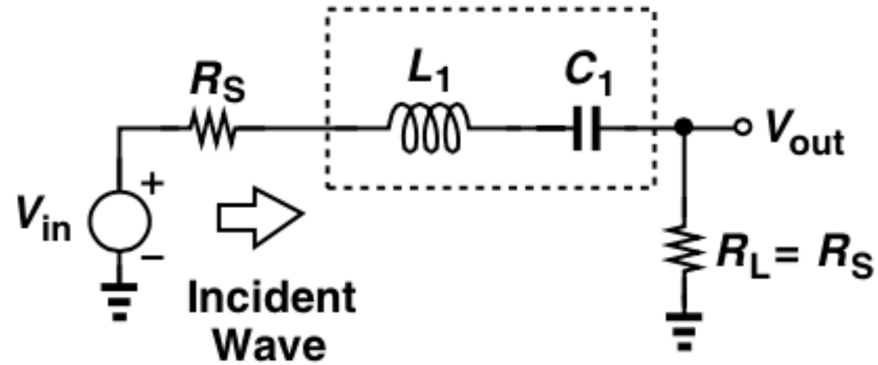
$$V_{in}^2/R_{in} = n^2 V_{in}^2/R_L$$

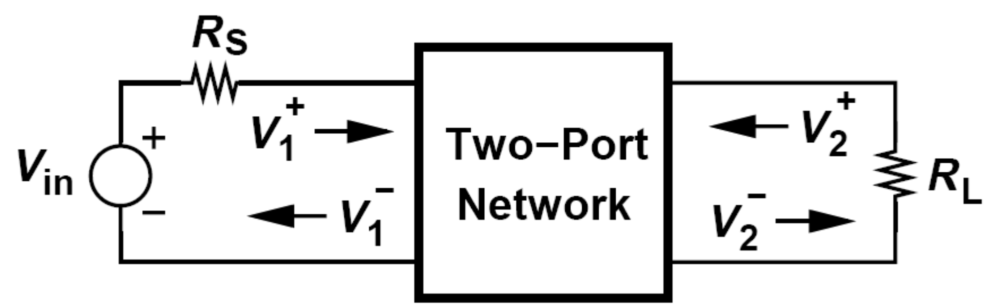
$$R_{in} = R_L/n^2$$



2.6 Scattering (S-) parameters

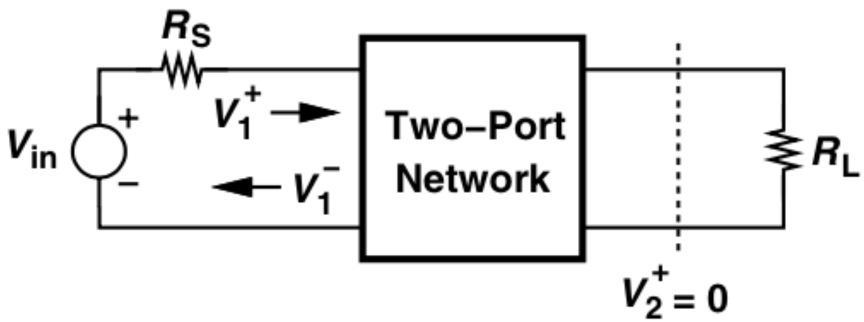
- Microwave circuits (transfer of power) vs. analog design (voltage). Affects tools, methods, models, etc.
- z-parameters require open/short termination for measurements which are hard to obtain for high frequencies (couple of 100 MHz).
- With S-parameters (scattering eller power wave) terminations (50 Ohms) are instead used for measurements





S-parameters

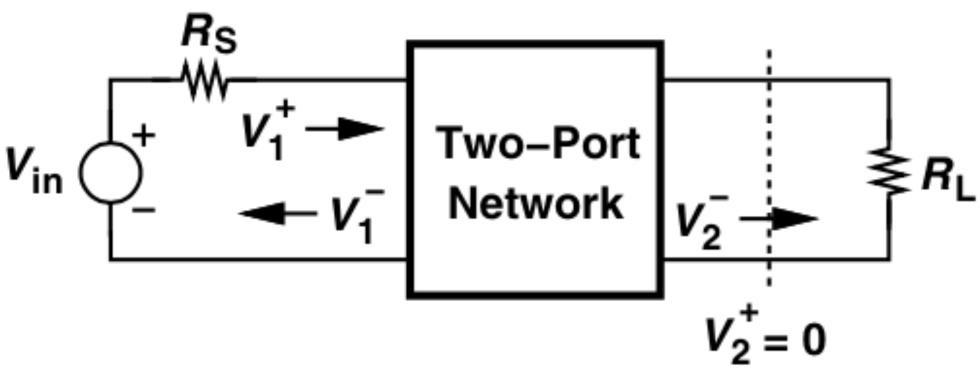
- S_{11} is the ratio of the reflected and incident waves at the input port when the reflection from R_L (i.e., V_2^+) is zero. Represents the input matching (but reflection, not impedance!).



$$S_{11} = \frac{V_1^-}{V_1^+}\bigg|_{V_2^+ = 0}.$$

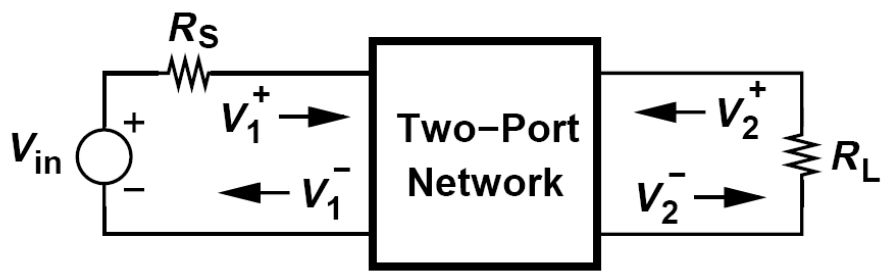
S-parameters

- S_{21} is the ratio of the wave incident on the load to that going to the input when the reflection from R_L is zero. Represents the gain.



$$S_{21} = \frac{V_2^-}{V_1^+}\Big|_{V_2^+ = 0}.$$

s-parameters



Input matching

Reverse isolation

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

Gain

Output matching

complex = Re+Im or A+Ph, and frequency dependent

S and Z parameters

$$Z_{11} = \frac{((1+S_{11})(1-S_{22})+S_{12}S_{21})}{\Delta_S} Z_0$$

$$Z_{12} = \frac{2S_{12}}{\Delta_S} Z_0$$

$$Z_{21} = \frac{2S_{21}}{\Delta_S} Z_0$$

$$Z_{22} = \frac{((1-S_{11})(1+S_{22})+S_{12}S_{21})}{\Delta_S} Z_0$$

Where

$$\Delta_S = (1-S_{11})(1-S_{22}) - S_{12}S_{21}$$

The input impedance of a two-port network is given by:

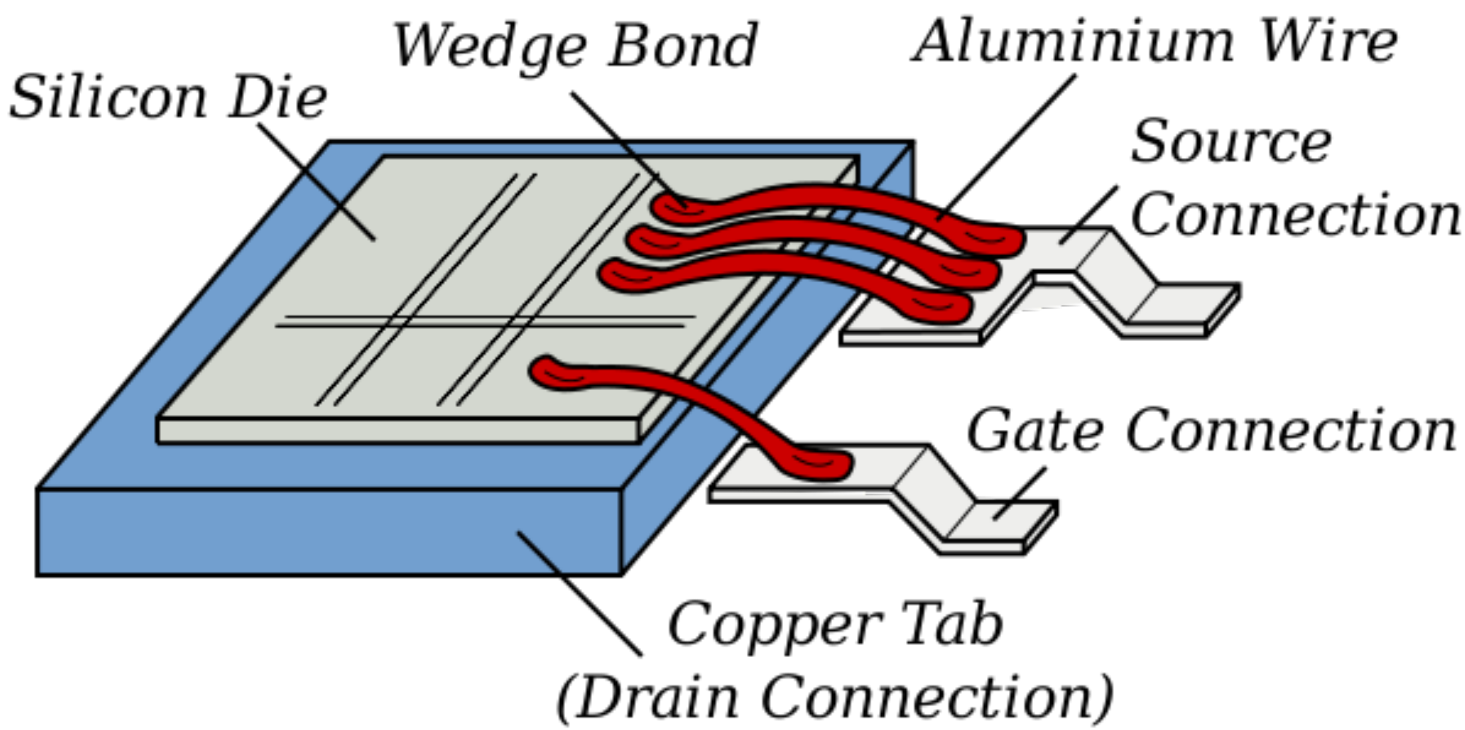
$$Z_{in} = Z_{11} - \frac{Z_{12}Z_{21}}{Z_{22}+Z_L}$$

where Z_L is the impedance of the load connected to port two.

S-parameters

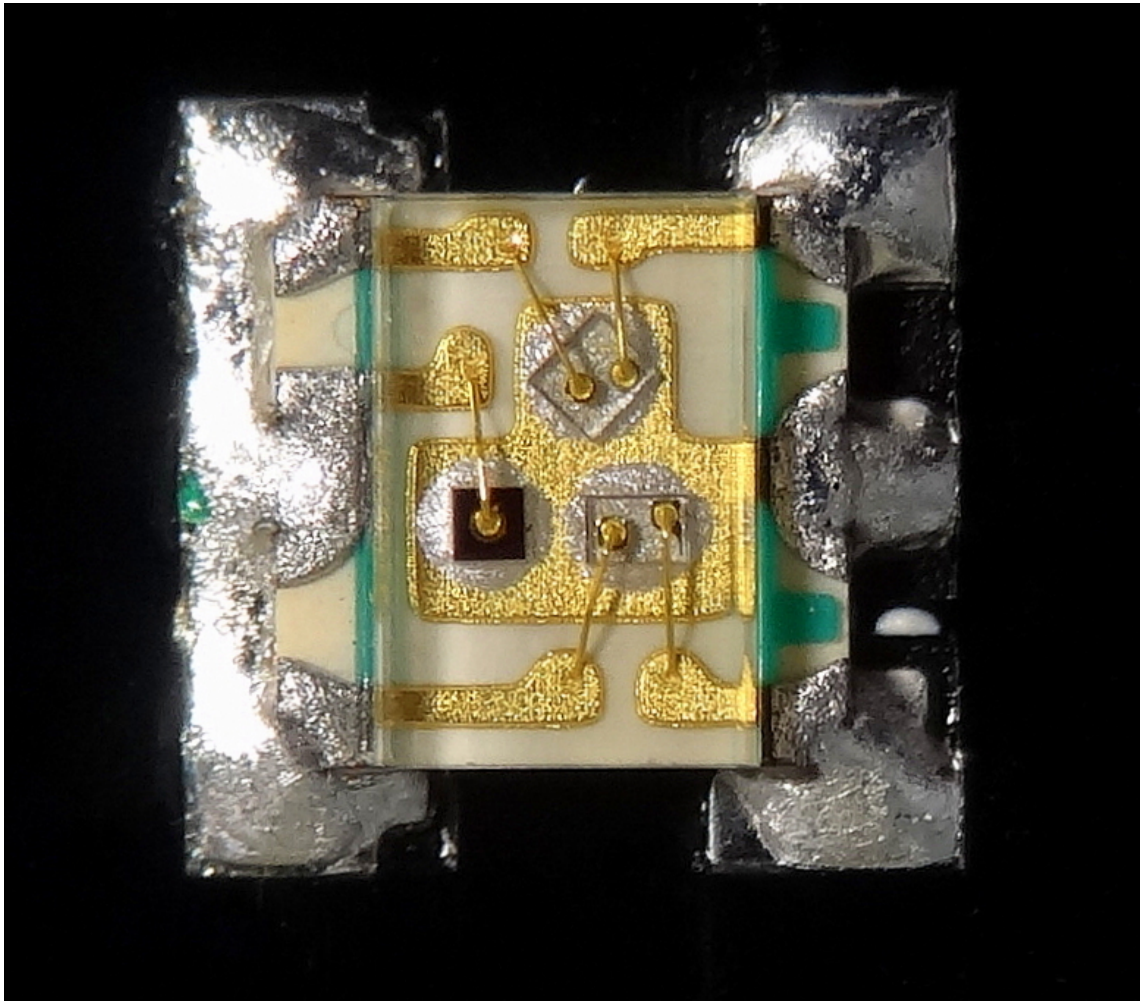
- S-parameters are typically measured using a network analyzer
- data is often displayed using Smith-charts

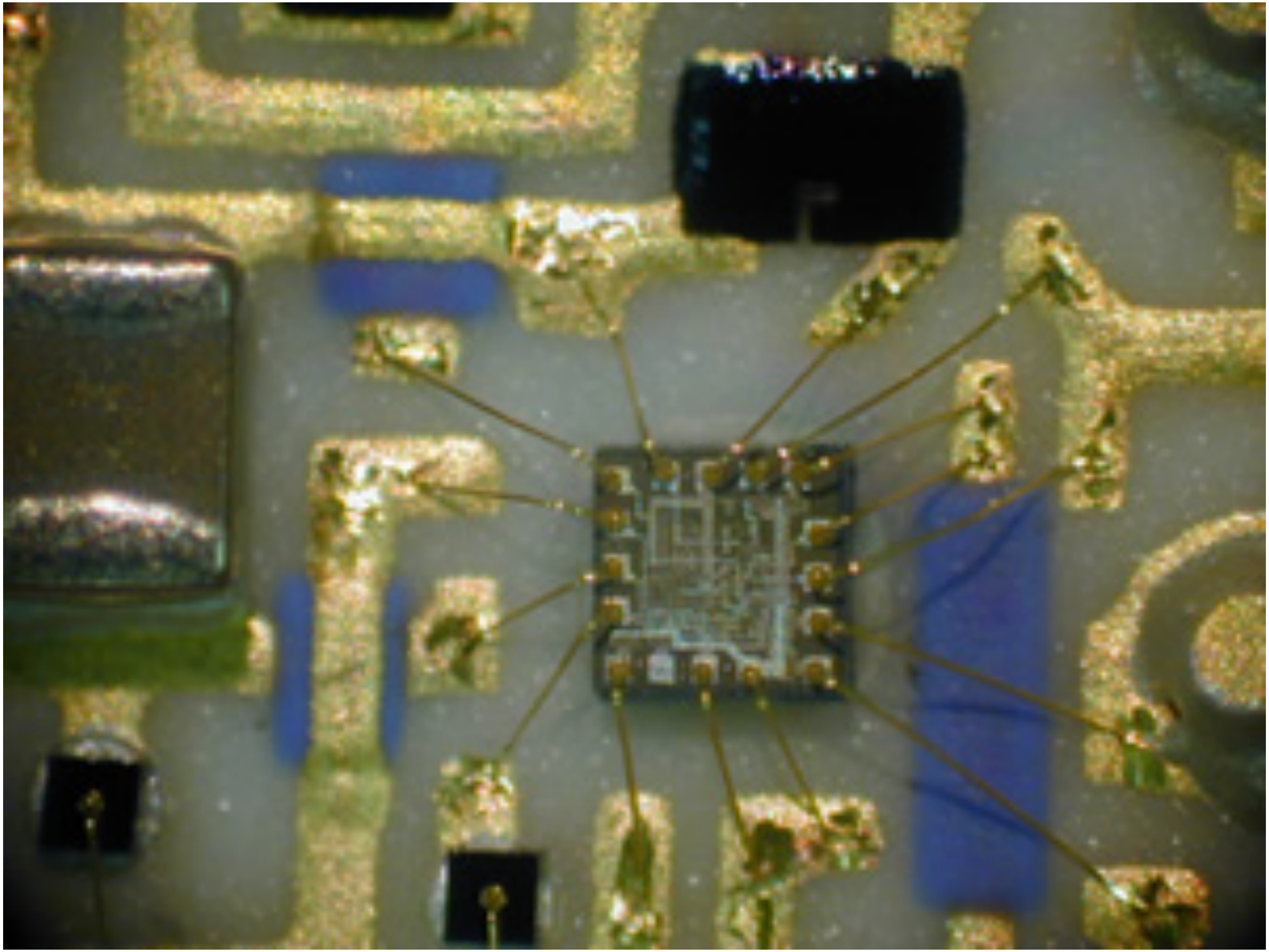
RF-IC packaging and measurements



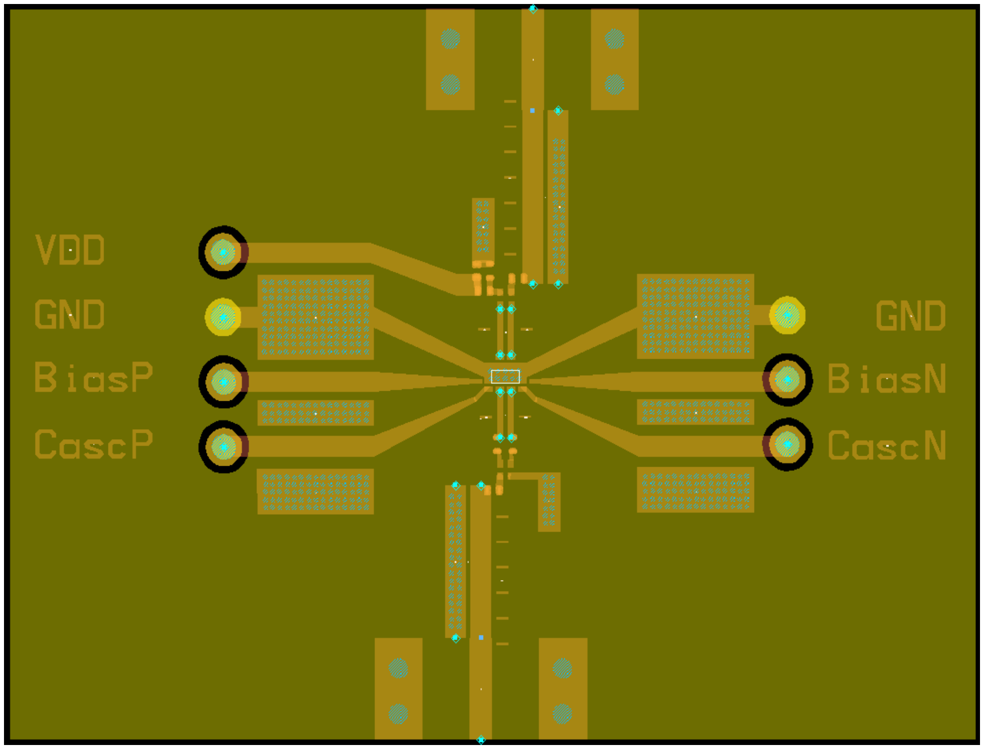
[Wikipedia]

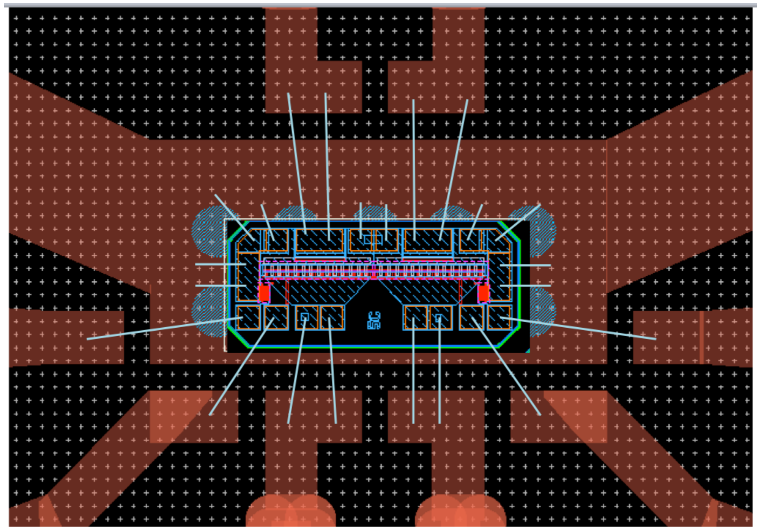
RF-IC packaging

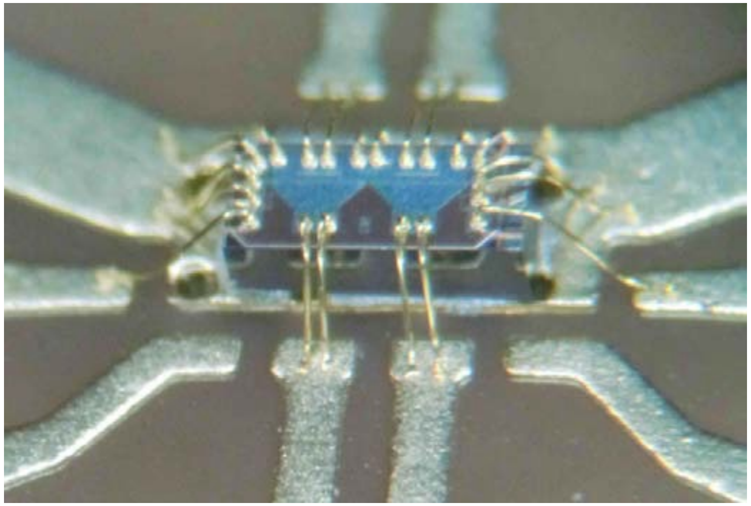




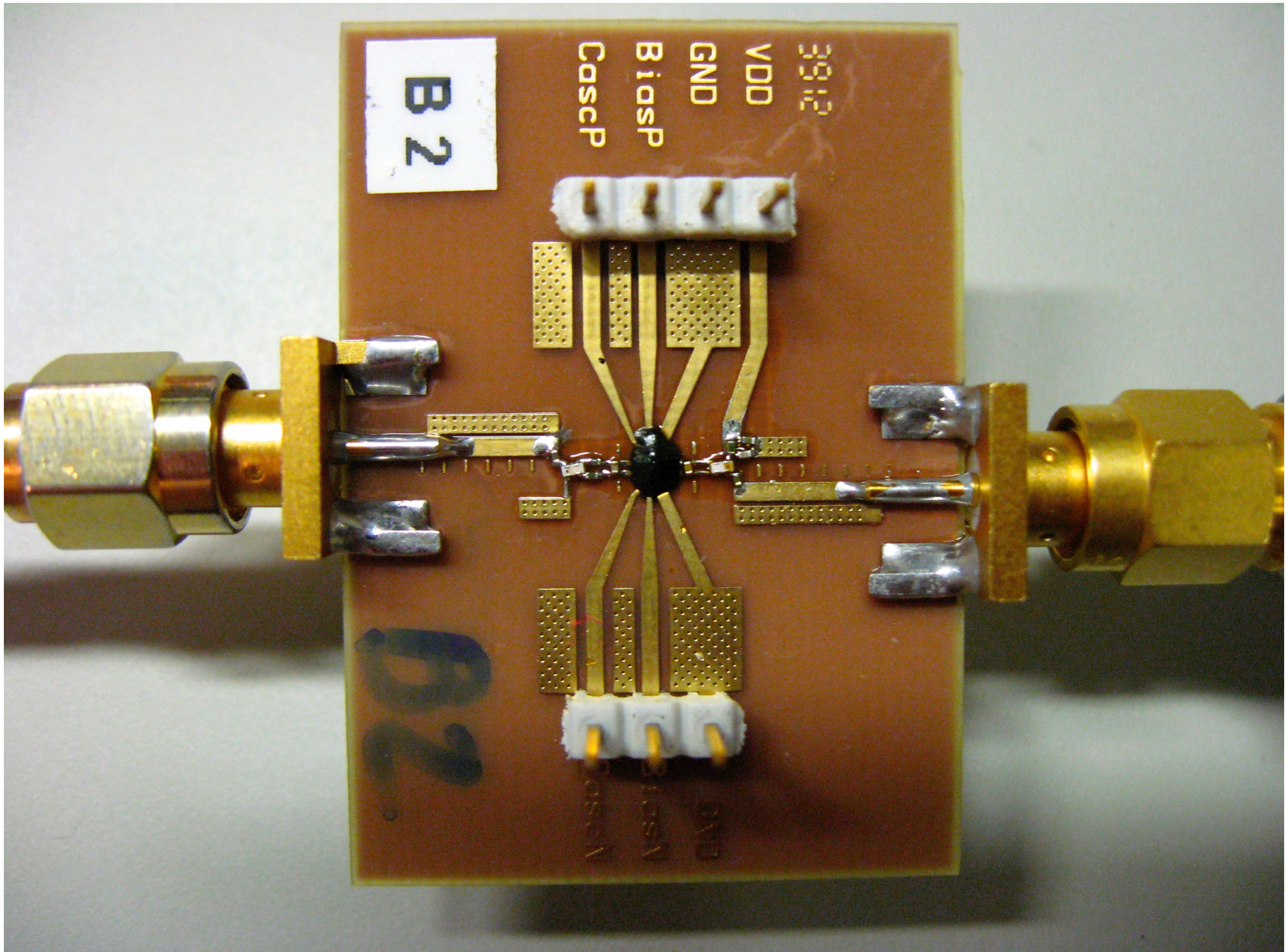
PCB and Bonding



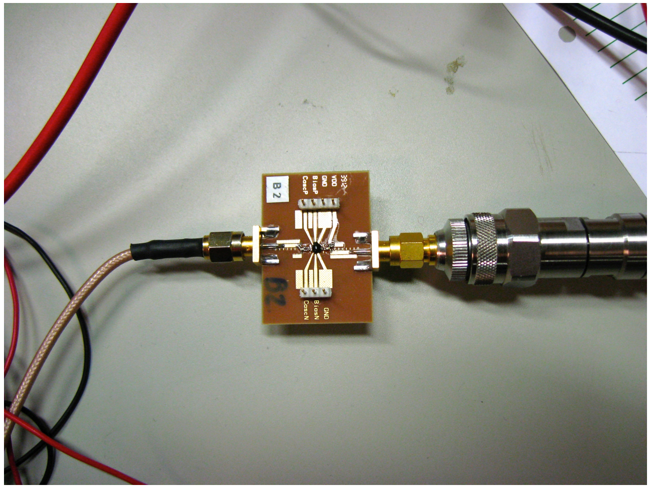


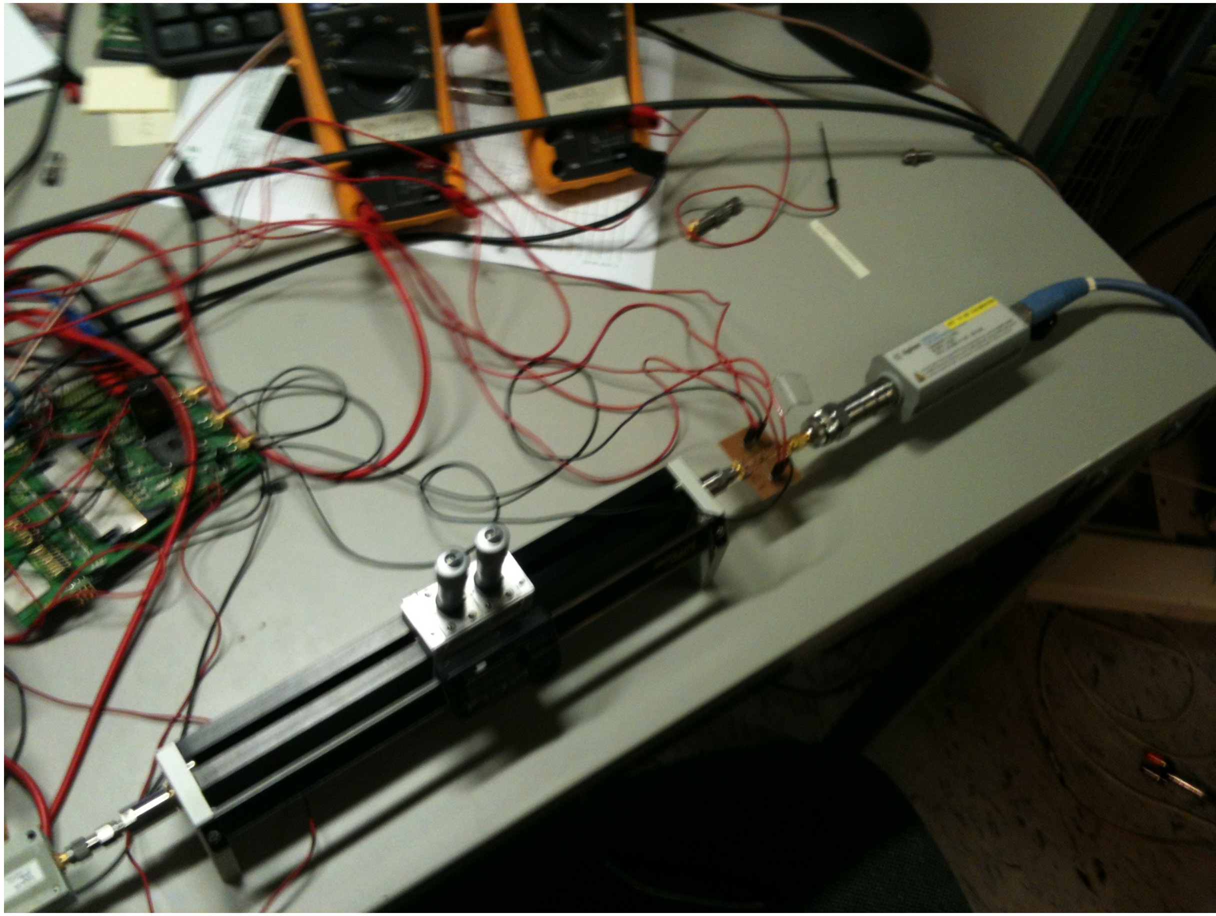


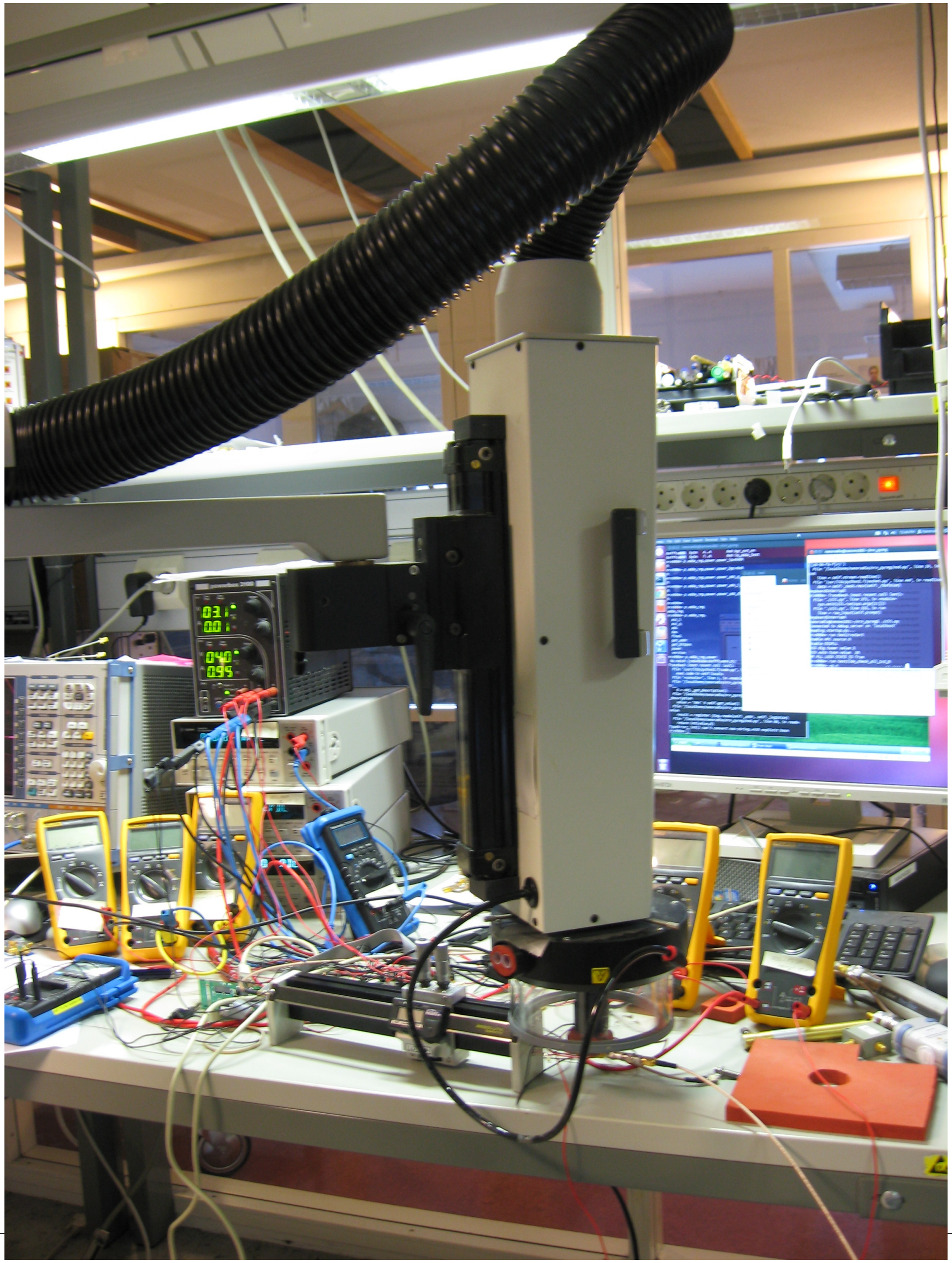
Mounted/soldered PCB



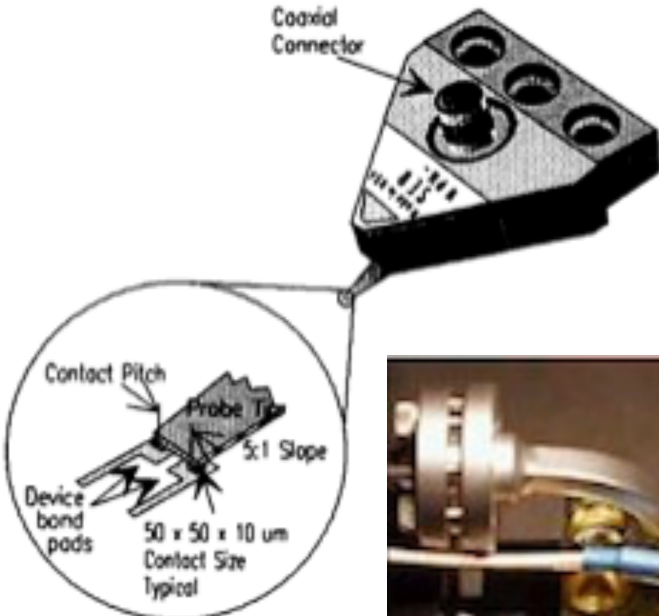
PCB measurements

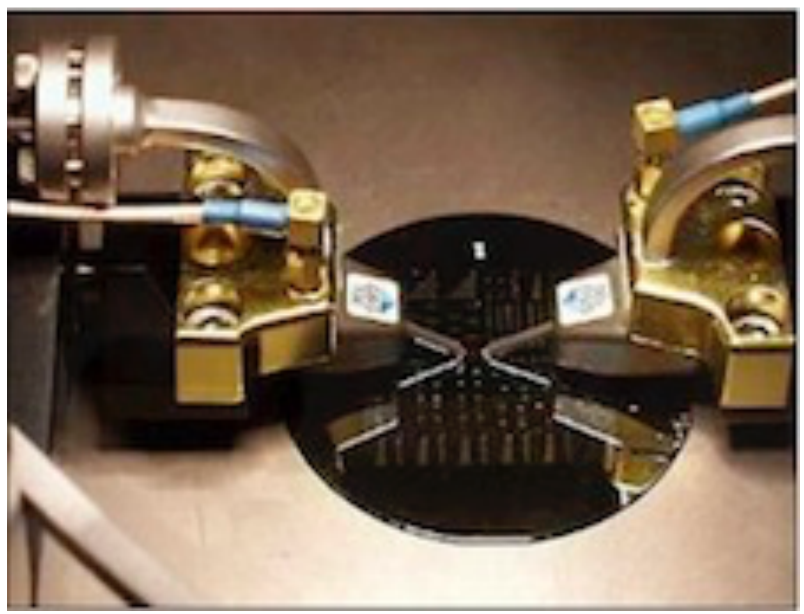


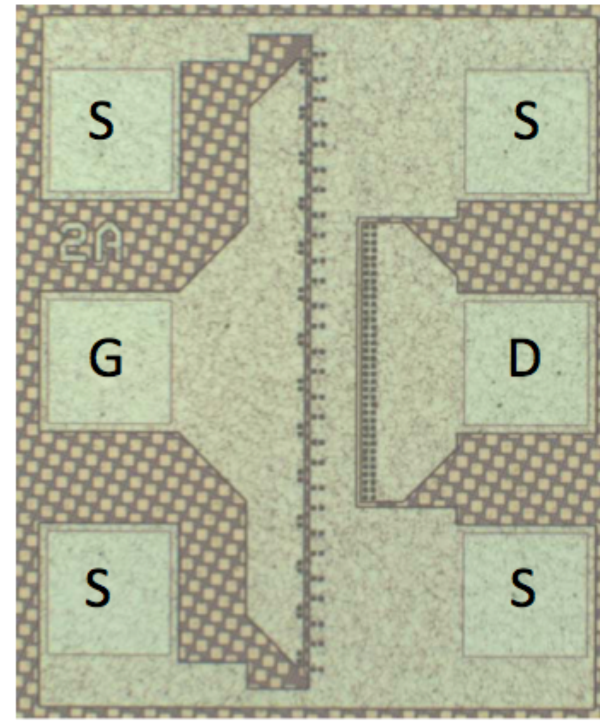




On-wafer measurements







Summary: S-parameters

- S-parameters are a powerful way to describe a linear electrical network at high frequency
- S-parameters change with frequency, load impedance, source impedance, network
- S_{11} is the reflection coefficient
- S_{21} describes the forward transmission coefficient (corresponds to gain)
- S-parameters have both magnitude and phase information
- S-parameters may describe large and complex networks

Stability (no self-oscillations)

- Stability of an RF circuit can be checked by Stern (Rollett) stability factor which is based on S-parameters:

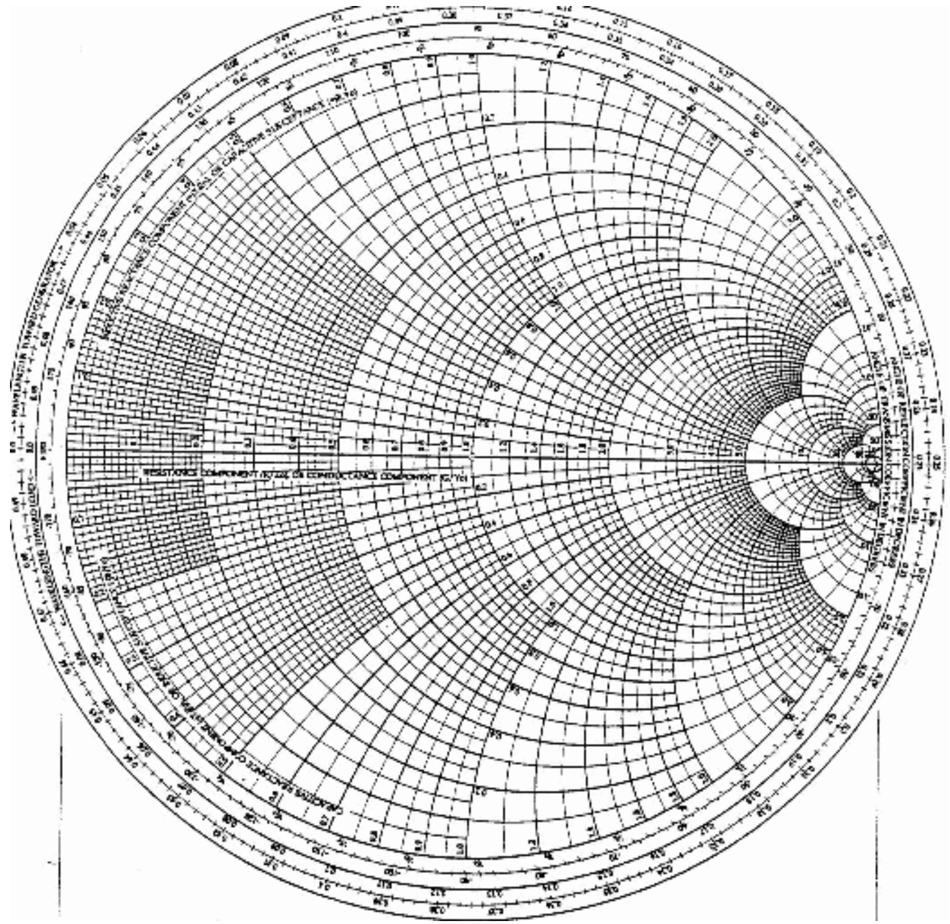
$$K = \frac{1 + |\Delta|^2 - |S_{11}|^2 - |S_{22}|^2}{2|S_{12}S_{21}|}$$

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

- If $K > 1$ and $|\Delta| < 1$, then the circuit is unconditionally stable for any combination of input and output impedances.

Smith chart

Philip H Smith
(1905 – 1987)



Smith chart

- The Smith chart is one of the most useful graphical tools for high frequency circuit applications.
- The goal of the Smith chart is to identify all possible impedances on the domain of existence of the reflection coefficient.

$$Z(d) = \frac{V(d)}{I(d)} = Z_0 \frac{1+\Gamma(d)}{1-\Gamma(d)}$$

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

- "Normalized reflection coefficient": $Z(d) = Z_0 * z(d)$

Smith chart

Z = R +/- j X

Impedance

The chart is normalized (Zn) so that any characteristic impedance (Zo) can be used.

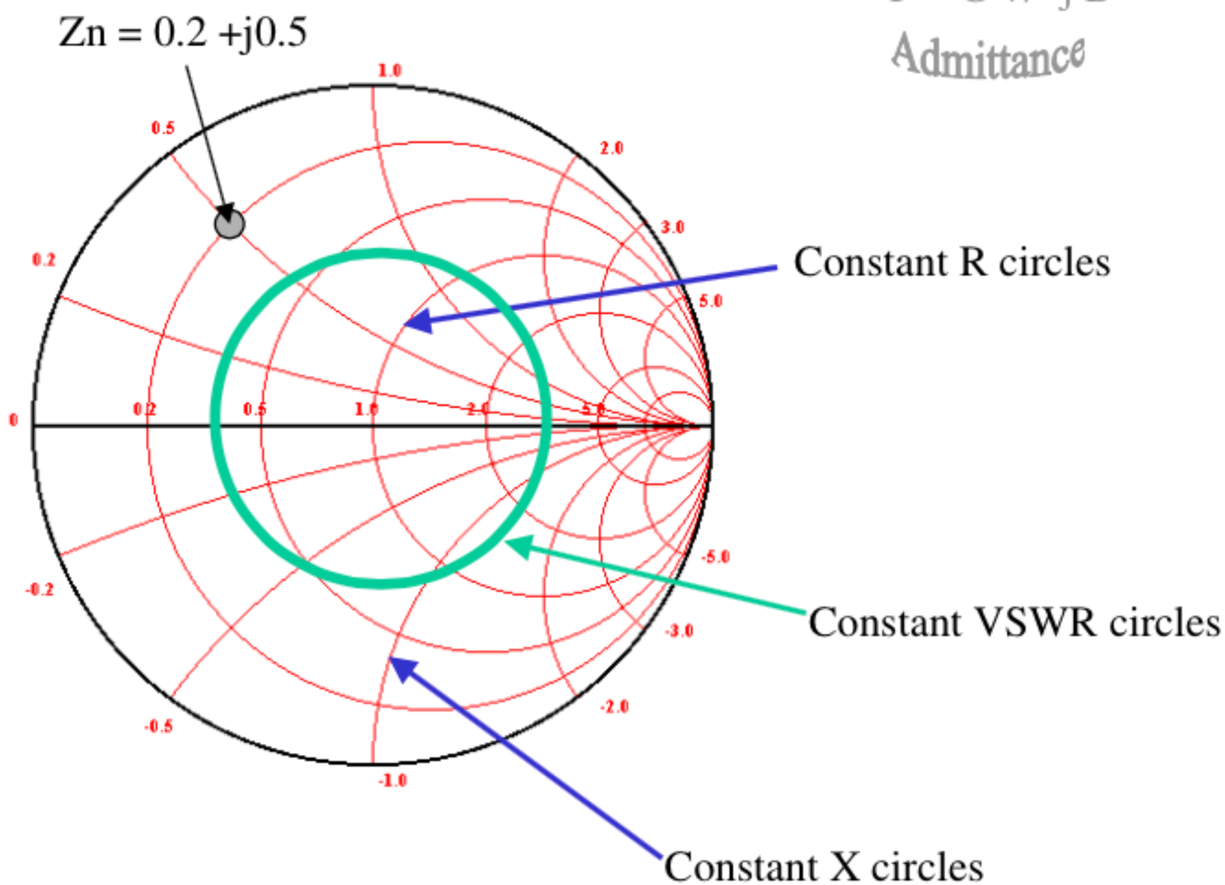
$$Zn = \frac{R}{Zo} +/- j \frac{X}{Zo}$$

$$VSWR = \frac{|V_{\max}|}{|V_{\min}|} = \frac{1+|\Gamma|}{1-|\Gamma|}.$$

"mismatch"

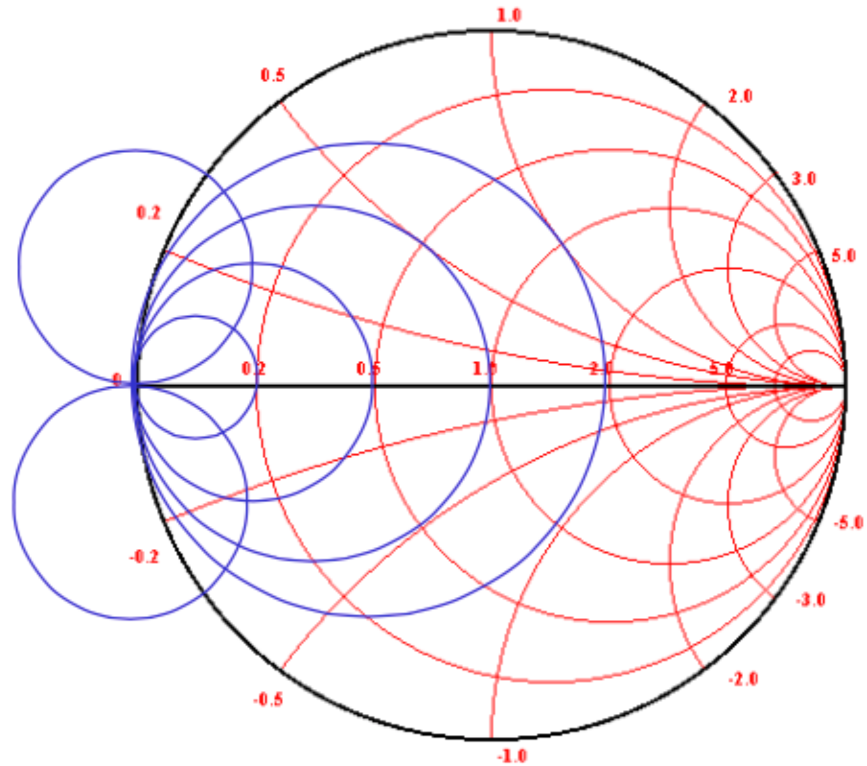
Y = G +/- j B

Admittance

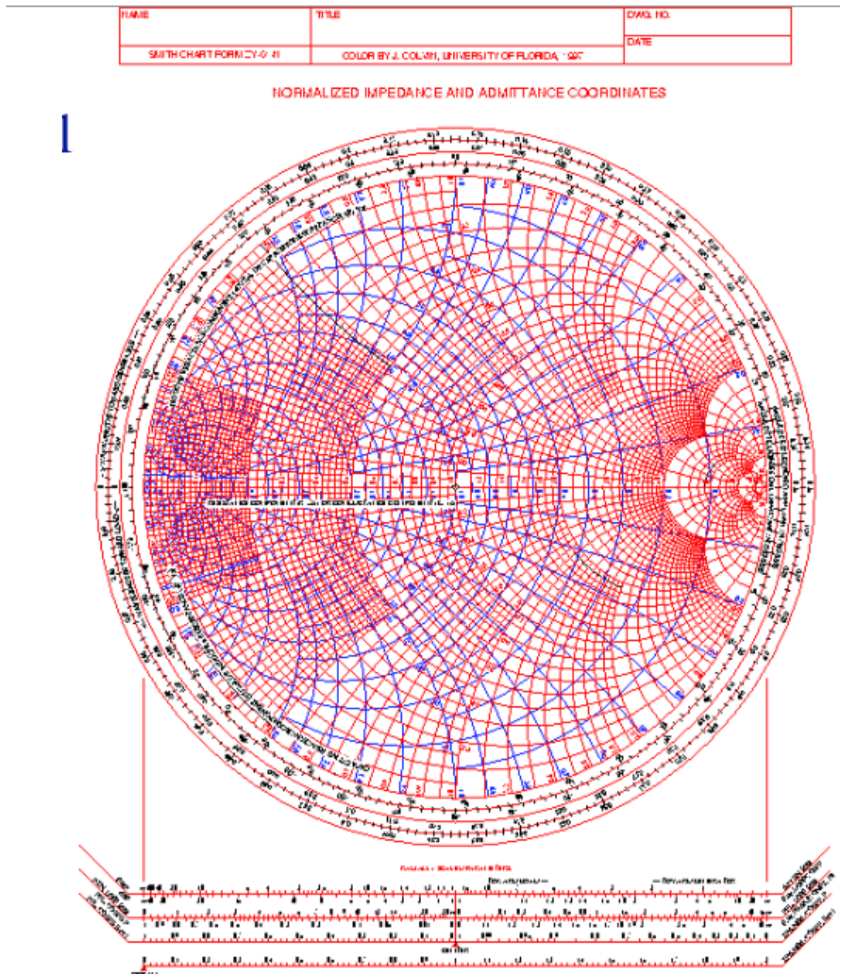


Smith chart

- There's also a mirror image of the chart that instead of having constant resistance circles, and constant reactance curves, has instead constant conductance circles and constant susceptance curves.



The full Smith chart



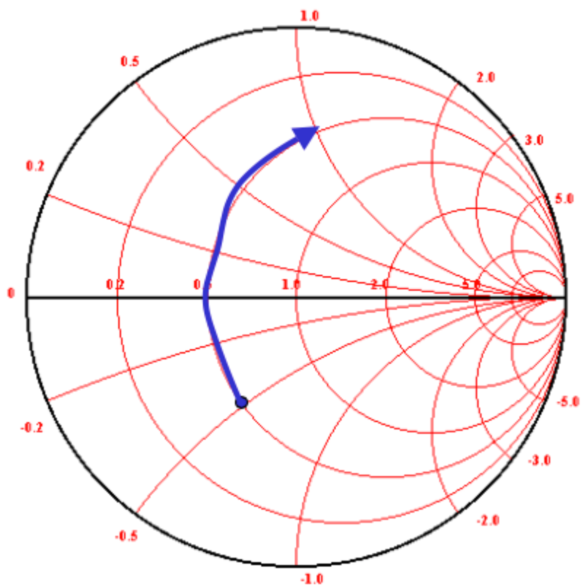
Smith Chart applications

- Plotting/displaying impedances, e.g. as a function of frequency.
- Matching (impedance transformation)
- Determine VSWR
- ....

L in the Smith chart

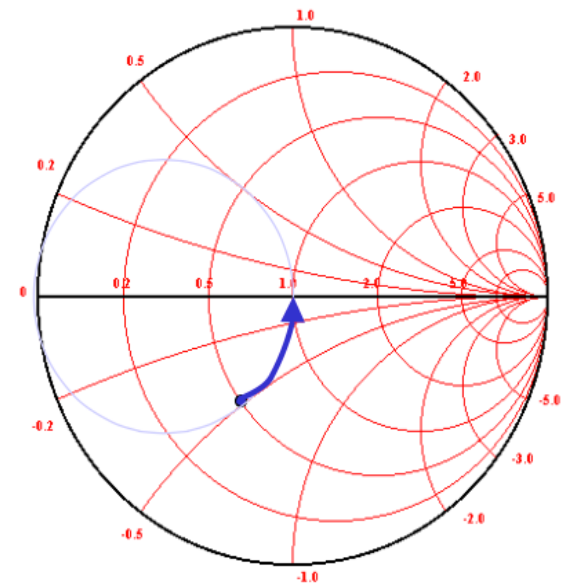
Series Inductors

Moves clockwise along circles of constant resistance



Shunt Inductors

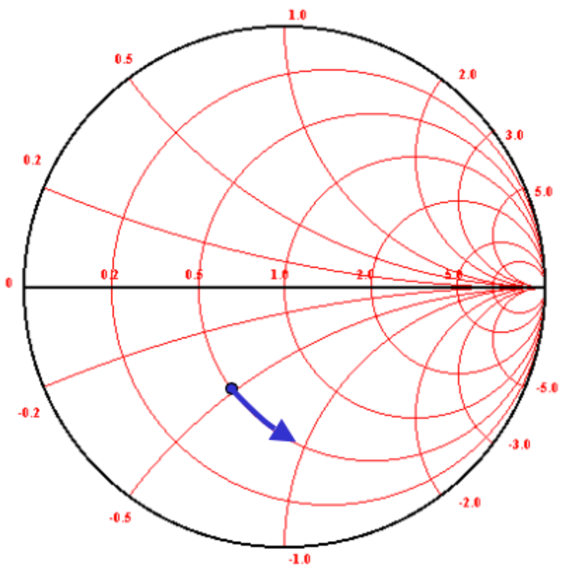
Moves counter-clockwise along circles of constant conductance



C in the Smith chart

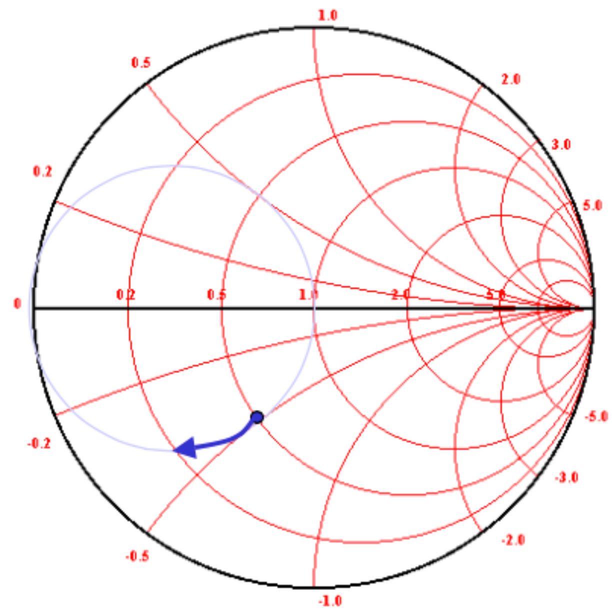
Series Capacitors

Moves counter-clockwise along circles of constant resistance



Shunt Capacitors

Moves clockwise along circles of constant conductance



Matching using Smith chart

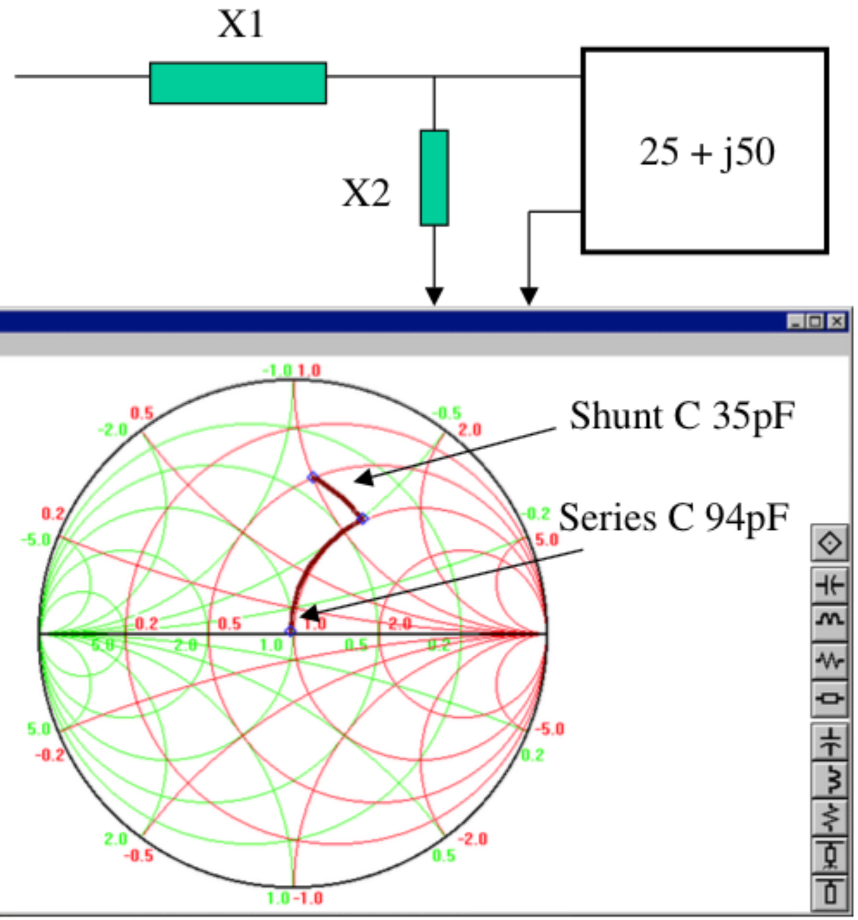
Let's do some matching with L's and C's

Pick Zo (50 Ohms sounds OK…)
Frequency is 28 MHz

The shunt Cap, transforms the R part of the Z to 1 (on the unit R circle..) from there a simple series C will take out the inductive reactance..

Easy as π

Are there other ways to match this same impedance??



Matching using Smith chart