Semester S1

Basics of Active and Non-Linear Electronics

PRACTICAL WORK

PW2:

VECTOR NETWORK ANALYSIS. APPLICATION TO THE EXPERIMENTALLY AND LINEARLY INPUT AND OUTPUT MATCHING OF A FET

OUTLINE

VECTOR NETWORK ANALYSIS. APPLICATION TO THE EXPERIMENTALLY AND LINEARLY INPUT AND OUTPUT MATCHING OF A FET

I OBJECTIVE

The objective of this Practical Work is to learn how to use a Vector Network Analyzer (VNA). This instrument is used to carry out measurements of Scattering parameters ([S]) at microwave frequencies. Because this instrument allows measuring complex quantities (modules and phases of the parameters [S]), it is called Vector Network analyzer. It is the basic tool for the characterization of microwave devices.

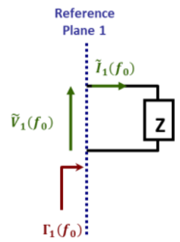
In this PW, theoretical basics associated with one-port and two-port are reminded in the microwave domain (reflection coefficients, TOS, parameters [S], Smith chart ...).
The experimental part will then consist in familiarizing with the VNA, performing calibration procedures and then characterizing and matching an active component (FET).

II THEORETICAL REMINDERS.

II.1 ONE-PORT CHARCATERISTICS

II.1.1 Impedance and Reflection Coefficient.

A microwave one-port is essentially characterized at a given f_0 frequency by its own impedance or its reflection coefficient:



:

$\tilde{V}_1(f_0)$, $\tilde{I}_1(f_0)$ are defined as voltage and current phasors at reference plane 1 or port 1. These phasors are associated with the physical instantaneous voltage and current defined as follows.

$$v_1(t) = V_1 \cos(\omega_0 t + \varphi_1) \Leftrightarrow v_1(t) = \mathcal{R}\left(\tilde{V}_1(f_0) e^{j\omega_0 t}\right) \text{ with: } \tilde{V}_1(f_0) = V_1 e^{j\varphi_1}$$

$$i_1(t) = I_1 \cos(\omega_0 t + \theta_1) \Leftrightarrow i(t) = \mathcal{R}\left(\tilde{I}_1(f_0) e^{j\omega_0 t}\right) \text{ with: } \tilde{I}_1(f_0) = I_1 e^{j\theta_1}$$

The complex impedance of the one-port is then defined as the following ratio:

$$Z(f_0) = \frac{\tilde{V}_1(f_0)}{\tilde{I}_1(f_0)} \Leftrightarrow Z(f_0) = \frac{V_1}{I_1} e^{j(\varphi_1 - \theta_1)}$$

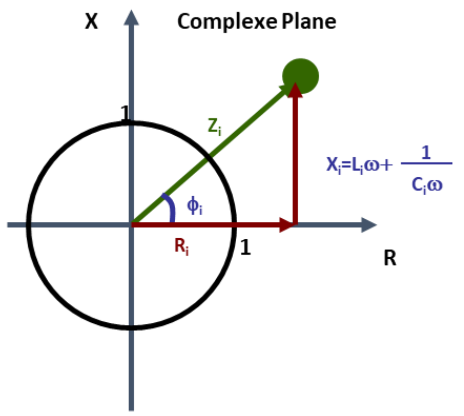
With these relationships:

$$Z_i(f_0) = \frac{\tilde{V}_1(f_0)}{\tilde{I}_1(f_0)} = R_i + jX_i$$

Impedance Resistance Reactance

$$Y_i(f_0) = \frac{\tilde{I}_1(f_0)}{\tilde{V}_1(f_0)} = G_i + jB_i$$

Admittance Conductance Susceptance



$$Y_i(f_0) = \frac{1}{Z_i(f_0)} = G_i + jB_i = \frac{1}{R_i + jX_i}$$

$$Y_i(f_0)_i = \frac{R_i - jX_i}{R_i^2 + jX_i^2} \Rightarrow \begin{cases} G_i = \dfrac{R_i}{R_i^2 + jX_i^2} \\ B_i = \dfrac{-X_i}{R_i^2 + jX_i^2} \end{cases}$$

In the microwave domain, the concept of scattering power wave is used: the incident power wave noted $\tilde{a}_1(f_0)$ and the reflective power wave noted $\tilde{b}_1(f_0)$:

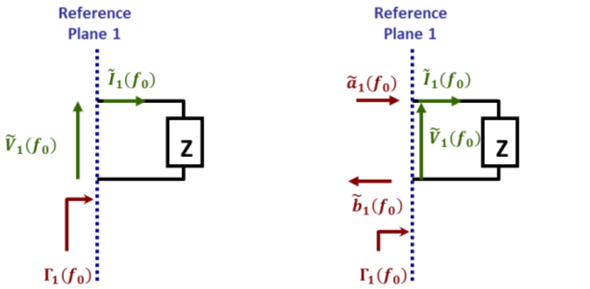
The relationships between these power waves and the voltage and current phasors are the following ones:

$$\begin{cases} \tilde{V}_i(f_0) = [\tilde{a}_i(f_0) + \tilde{b}_i(f_0)]\sqrt{Z_0} \\ \tilde{I}_i(f_0) = \dfrac{[\tilde{a}_i(f_0) - \tilde{b}_i(f_0)]}{\sqrt{Z_0}} \end{cases}$$

Zo is the normalization impedance or reference impedance. Virtually all microwave instrumentation devices have a purely real reference impedance equal to 50Ω.

The power waves $\tilde{a}_i(f_0)$ and $\tilde{b}_i(f_0)$ are homogeneous to $\sqrt{W}$.

The representation of the one-port in the power wave formalism is as follows:



$$\begin{cases} \tilde{V}_i(f_0) = [\tilde{a}_i(f_0) + \tilde{b}_i(f_0)]\sqrt{Z_0} \\ \tilde{I}_i(f_0) = \dfrac{[\tilde{a}_i(f_0) - \tilde{b}_i(f_0)]}{\sqrt{Z_0}} \end{cases}$$

The reflection definition at port1 and at the given f_o working frequency is defined by:

$$\Gamma(f_0) = \frac{\tilde{b}_1(f_0)}{\tilde{a}_1(f_0)}$$

The reflection coefficient is a complex number without any unit.

It can be demonstrated that:

$$\Gamma(f_0) = \frac{\tilde{b}_1(f_0)}{\tilde{a}_1(f_0)} \Leftrightarrow \Gamma(f_0) = \frac{Z(f_0) - Z_0}{Z(f_0) + Z_0}$$

Or $Z(f_0) = \frac{1+\Gamma(f_0)}{1-\Gamma(f_0)}$

Special cases:

- Short-circuit : $Z_{short}(f_0) = 0e^{j\alpha}$ and $\Gamma_{short}(f_0) = -1$ i.e. $|\Gamma_{short}(f_0)| = 1$ and $phase[\Gamma_{short}(f_0)] = 180$
- Open-circuit : $Z_{open}(f_0) = \infty e^{j\alpha'}$ and $\Gamma_{open}(f_0) = 1$ i.e. $|\Gamma_{open}(f_0)| = 1$ and $phase[\Gamma_{open}(f_0)] = 0$
- Match : $Z_{short}(f_0) = 50e^{j0°}$ and $\Gamma_{Match}(f_0) = 0e^{j\varepsilon}$ i.e. $|\Gamma_{match}(f_0)| = 0$ and $phase[\Gamma_{match}(f_0)] = \varepsilon$

The Standing Wave Ratio (SWR) is defined as :

$$\mathrm{SWR}(f_0) = \frac{1 + |\Gamma(f_0)|}{1 - |\Gamma(f_0)|}$$

For a passive impedance, the following relationship is always valid:

$$0 \leq |\Gamma(f_0)| \leq 1 \Leftrightarrow \mathrm{SWR}(f_0) \geq 1$$

II.1.2 Rectangular/Polar Frequency Representation of Reflection Coefficient

Two types of representation of the reflection coefficient as a function of the frequency exist:

- Magnitude and phase:

  In this case, we represent the magnitude of $\Gamma(f_0)$ (either in linear or in decibels), and the phase of $\Gamma(f_0)$ as a function of frequency.

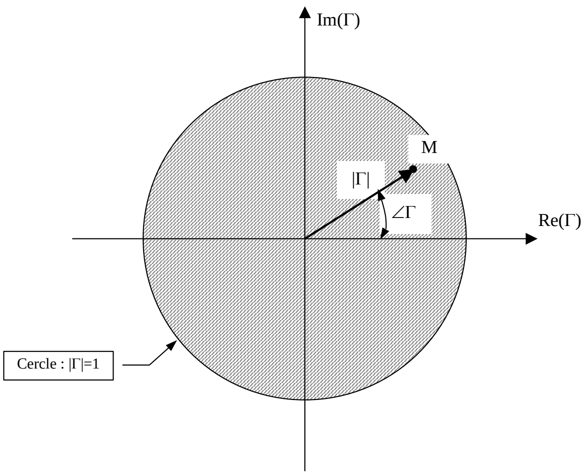
- Real and imaginary:

  In this case, we represent the real part and the imaginary part of $\Gamma(f_0)$ as a function of the frequency.

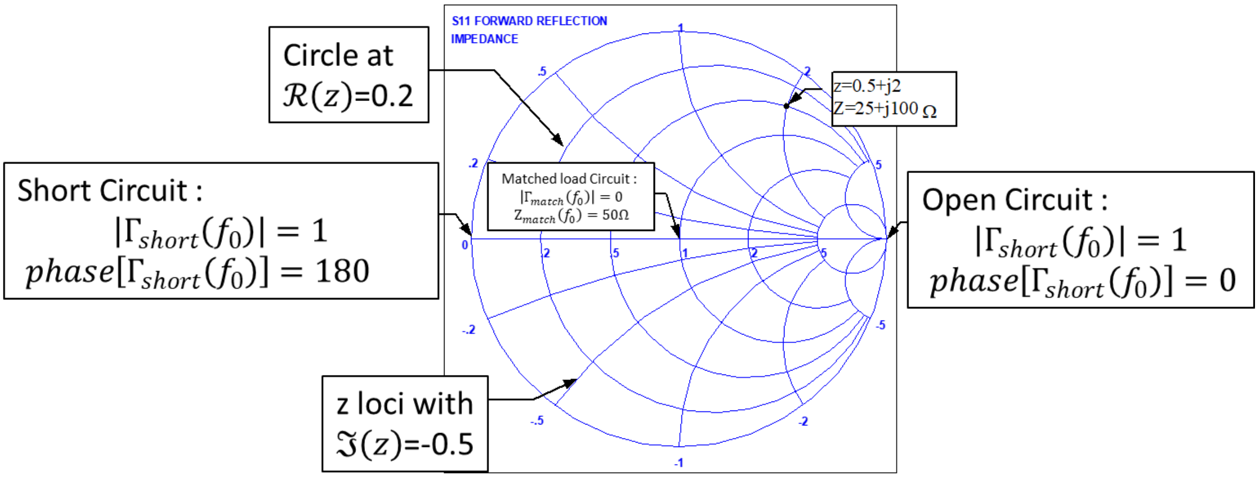
Reminder: $|\Gamma(f_0)|_{(dB)} = 20log_{10}(|\Gamma(f_0)|)$

II.1.3 Representation of Reflection Coefficient using Smith Chart

The Smith chart is constructed by considering a Cartesian landmark representing the imaginary part of a reflection coefficient according to its real part:



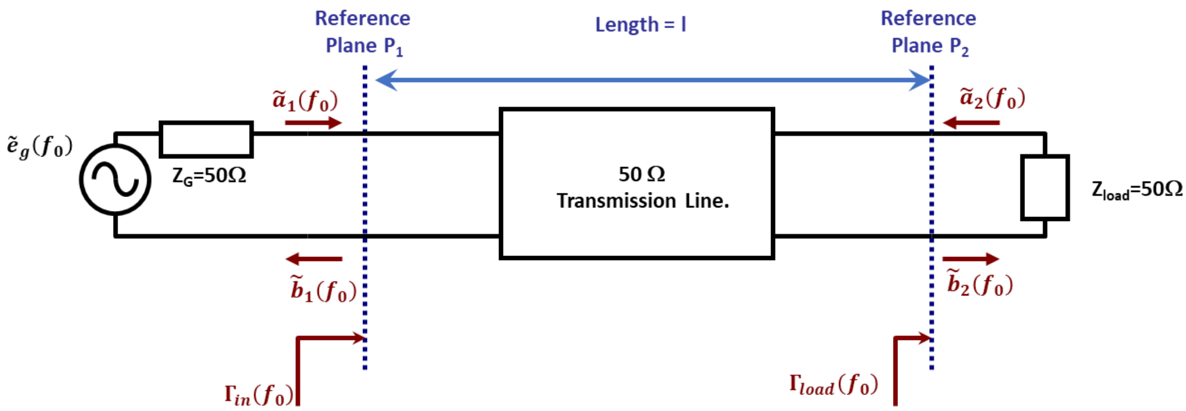
In this reference, an impedance (M point) is represented by its reflection coefficient. For a passive impedance, the M point is obligatorily located inside the circle $|\Gamma(f_0)| \leq 11$ (hatched area). This circle represents the perimeter of Smith Chart. This representation allows plotting or reading only reflection coefficients. To directly plot or to directly read the values of the impedances, the loci with constant real part of $Z(f_0)$ as well as loci with constant imaginary part of $Z(f_0)$ have been represented. The standardized Smith chart shown below is thus obtained. Normalization consists in representing an impedance $Z(f_0)$ by its normalized value $z(f_0)$ with respect to the reference impedance Z_0:



$$z(f_0) = \frac{Z(f_0)}{Z_0}$$

II.1.4 Change of reference plane

Consider a lossless transmission line excited by an internal impedance generator 50Ω and loaded by a reflection coefficient charge $\Gamma_{Load}(f_0)$



Let P_2 be the plane of the load and P_1 be the plane at a distance l of P_2.

The reflection coefficients of the impedances brought back in the planes P_1 and P_2 are:

$$\Gamma_{in}(f_0) = \frac{\tilde{b}_1(f_0)}{\tilde{a}_1(f_0)} \text{ and } \Gamma_{load}(f_0) = \frac{\tilde{a}_2(f_0)}{\tilde{b}_2(f_0)}$$

And we have the following relationships:

$$\begin{cases} \tilde{b}_2(f_0) = \tilde{a}_1(f_0)e^{-j\beta l} & (1) \\ \tilde{a}_2(f_0) = -\tilde{b}_1(f_0)e^{+j\beta l} & (2) \end{cases} \text{ with } \beta = \frac{2\pi}{\lambda_g} \text{ where } \lambda_g \text{ is the guided wavelength.}$$

For a propagating wave on a coaxial cable (TEM propagation mode), we have:

$$\lambda_g = \frac{\lambda_0}{\sqrt{\varepsilon_r}} \text{ with } \lambda_0 = \frac{c}{f_0} \text{ where } \begin{cases} c = light\ speed\ in\ vacuum = 3.10^8\ m/s \\ f_0 = frequency\ of\ the\ microwave\ signal \end{cases} \text{ and}$$

$$\varepsilon_r = permittivity\ of\ the\ dielectric\ material\ separating\ the\ two\ conductors\ of\ the\ coaxial\ cable$$

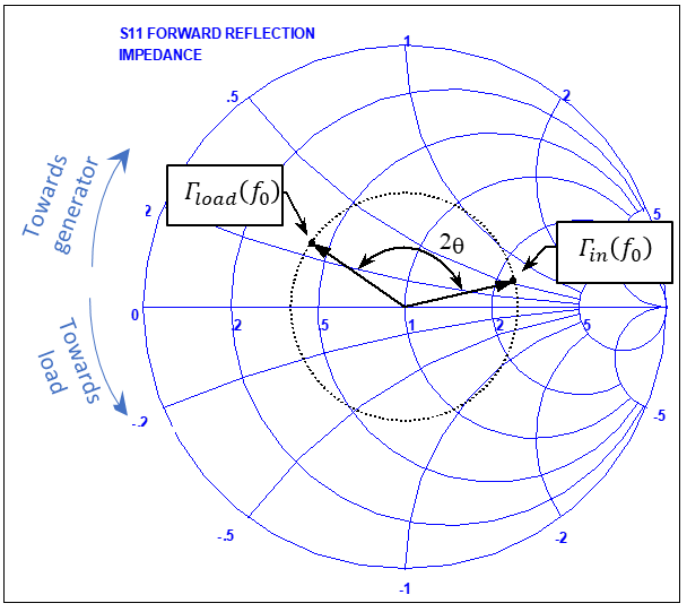
Material	air	wood	distilled water	Sea water	Alumina	Epoxy	Pyrex	plexiglass	Teflon	sapphire
ε_r	1	2	76.7	20	9.8	4.7	5	3.45	2.1	11.5

The product $\beta l = \frac{2\pi l}{\lambda_g} = \theta_{(rad)}$ is called the ***electric length.***

By dividing the equations (1) and (2) of previous power waves, we obtain:

$\Gamma_{in}(f_0) = \Gamma_{load}(f_0)e^{-j2\beta l} = \Gamma_{load}(f_0)e^{-j2\theta}$ or $\Gamma_{load}(f_0) = \Gamma_{in}(f_0)e^{+j2\theta}$

Finally, we have $\Gamma_{load}(f_0) = \Gamma_{in}(f_0)$ every $l = k\frac{\lambda_g}{2}$ with $k \in \mathbb{N}$

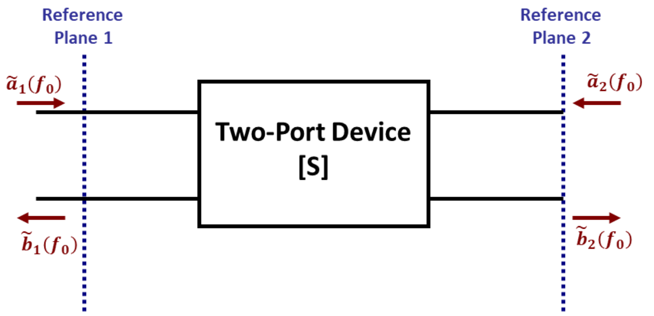


- A reference plane change along a lossless transmission line results in a phase rotation of the constant reflection coefficient magnitude:
  - ✓ anticlockwise if the change of plane is from the generator to the load,
  - ✓ clockwise if the change of plan is from the load to the generator.
- A complete turn of the smith Chart corresponds to a displacement of a length equal to: $\frac{\lambda_g}{2}$

II.2 TWO-PORT CHARACTERISTICS

II.2.1 Scattering Parameter : [S].

A linear microwave two-port can be described by its scattering matrix ([S] matrix) relative to reference planes. The [S] matrix is normalized with respect to the reference impedance Z_0. This matrix connects the incident power waves $\tilde{a}_1(f_0)$, $\tilde{a}_2(f_0)$ and reflected $\tilde{b}_1(f_0)$ and $\tilde{b}_2(f_0)$ present at the both ports of the two-port device:



The scattering matrix [S] is defined as:

$$[\tilde{b}(f_0)]=[S][\tilde{a}(f_0)] \text{ or } \begin{cases} \tilde{b}_1(f_0) = S_{11}(f_0)\tilde{a}_1(f_0) + S_{12}(f_0)\tilde{a}_2(f_0) \\ \tilde{b}_2(f_0) = S_{21}(f_0)\tilde{a}_1(f_0) + S_{22}(f_0)\tilde{a}_2(f_0) \end{cases}$$

The [S] matrix is then equal to: $[S] = \begin{bmatrix} S_{11}(f_0) & S_{12}(f_0) \\ S_{21}(f_0) & S_{22}(f_0) \end{bmatrix}$

Each element of the [S] matrix is a complex number:

The [S] parameters are intrinsic elements to the two-port and in no way depend on the two-port that can be connected either to the input or to the output of the latter.

II.2.1 S_{ij} Definitions and interpretation.

II.2.1.1 S_{11} and S_{21} definitions.

According to the previous equation system

$$S_{11}(f_0) = \left.\frac{\tilde{b}_1(f_0)}{\tilde{a}_1(f_0)}\right|_{\tilde{a}_2(f_0)=0} \text{ and } S_{21}(f_0) = \left.\frac{\tilde{b}_2(f_0)}{\tilde{a}_1(f_0)}\right|_{\tilde{a}_2(f_0)=0}$$

Physically, $\tilde{a}_2(f_0) = 0$ means that there is no reflected wave $\tilde{a}_2(f_0)$ by the load connected to the output of the two-port. This load therefore has a zero-reflection coefficient, it is a 50Ω matched load. The two-port is excited by its input port.

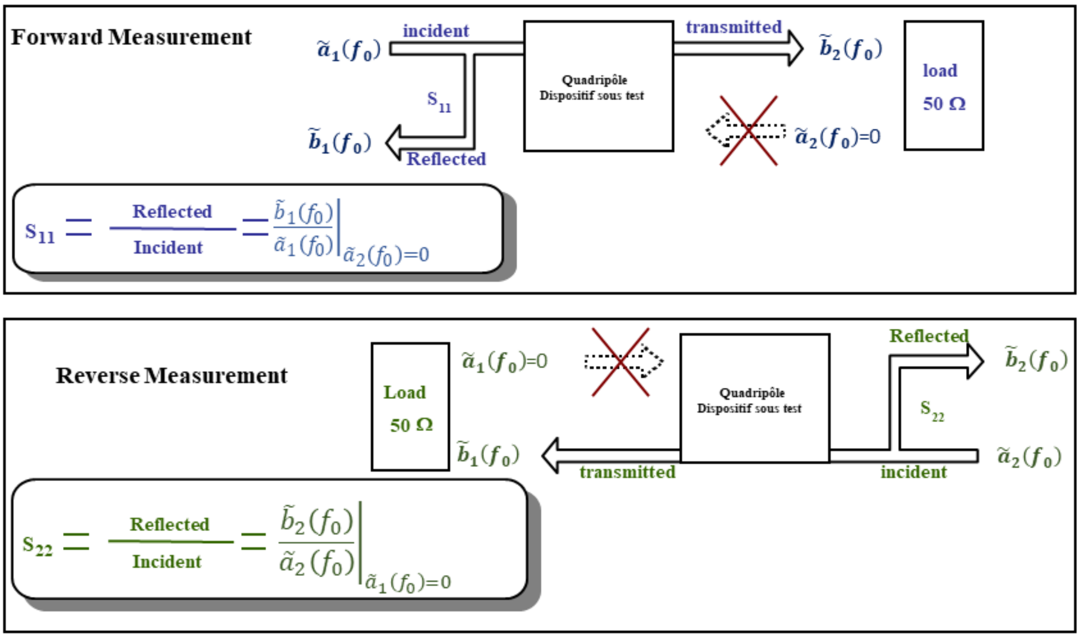
II.2.1.1 S_{22} and S_{12} definitions.

According to the previous equation system

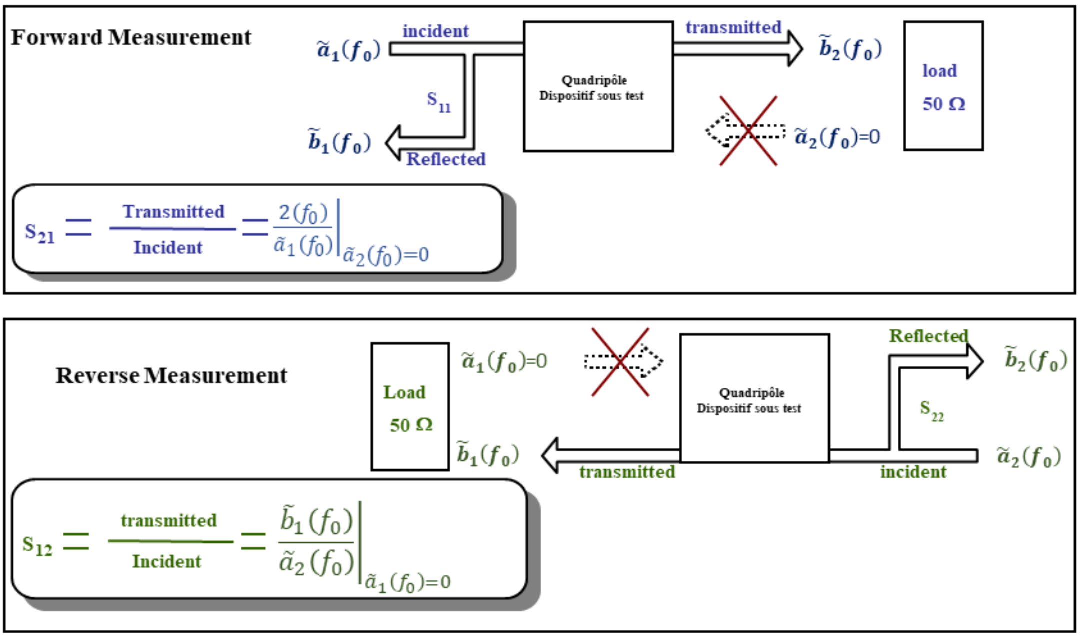
$$S_{12}(f_0) = \left.\frac{\tilde{b}_1(f_0)}{\tilde{a}_2(f_0)}\right|_{\tilde{a}_1(f_0)=0} \text{ and } S_{22}(f_0) = \left.\frac{\tilde{b}_2(f_0)}{\tilde{a}_2(f_0)}\right|_{\tilde{a}_1(f_0)=0}$$

Physically, $\tilde{a}_1(f_0) = 0$ means that there is no reflected wave $\tilde{a}_1(f_0)$ by the load connected to the input of the two-port. This load therefore has a zero-reflection coefficient, it is a 50Ω matched load. The two-port is excited by its output port.

Parameter $S_{11}(f_0)$ is the reflection coefficient at the input when the output is loaded by 50Ω. The parameter $S_{22}(f_0)$ is the reflection coefficient at the output when the input is loaded by 50 Ω.



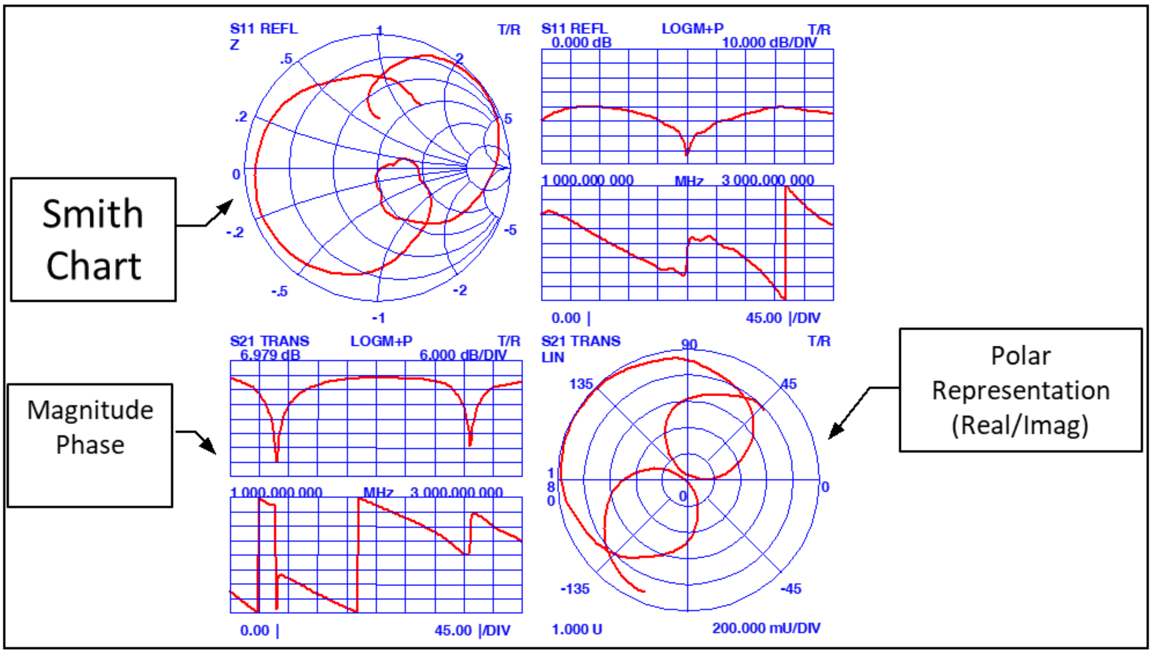
The parameter $S_{21}(f_0)$ is the input to output transmission coefficient (forward) when the output is loaded by 50 Ω. The parameter $S_{12}(f_0)$ is the transmission coefficient output to input (reverse) when the input is loaded by 50 Ω.



II.2.2 Sij Representations

For parameters $S_{11}(f_0)$ and $S_{22}(f_0)$, the most common use is either the Smith chart or a magnitude / phase representation.

For parameters $S_{21}(f_0)$ and $S_{12}(f_0)$, either the magnitude / phase representation or a polar representation (Im / Re) is used.



III PRINCIPLE OF VECTOR NETWORK ANALYSIS.

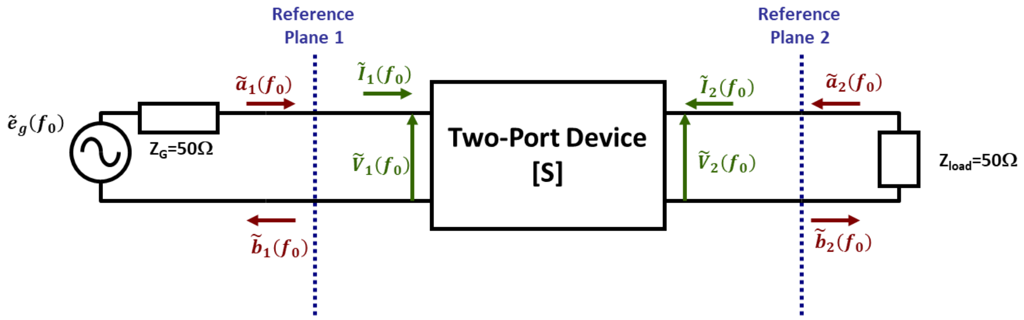
III.1 DESCRIPTION OF THE MEASUREMENT PRINCIPLE OF [S] PARAMETERS

To carry out the measurement of the *S* parameters, it is necessary to experimentally respect the conditions of the definitions of the *S*ij .

Reminder of $S_{11}(f_0)$ and $S_{21}(f_0)$ Definitions:

$$S_{11}(f_0) = \left.\frac{\tilde{b}_1(f_0)}{\tilde{a}_1(f_0)}\right|_{\tilde{a}_2(f_0)=0} \text{ and } S_{21}(f_0) = \left.\frac{\tilde{b}_2(f_0)}{\tilde{a}_1(f_0)}\right|_{\tilde{a}_2(f_0)=0}$$

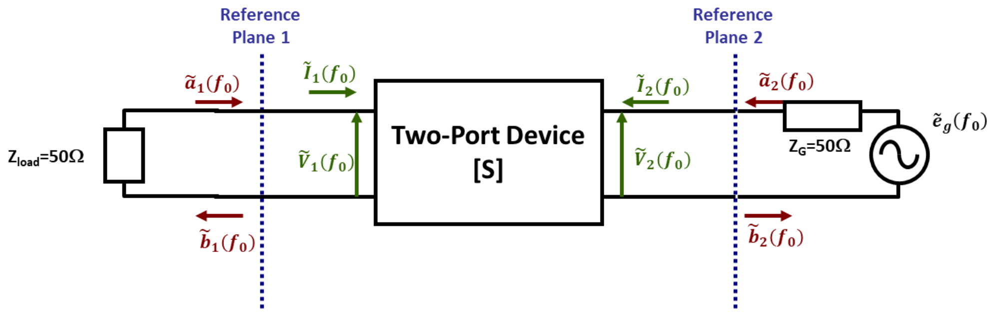
The schematic corresponding to the definitions of $S_{11}(f_0)$ and $S_{21}(f_0)$ is therefore the following:



Reminder of $S_{12}(f_0)$ and $S_{22}(f_0)$ Definitions:

$$S_{12}(f_0) = \left.\frac{\tilde{b}_1(f_0)}{\tilde{a}_2(f_0)}\right|_{\tilde{a}_1(f_0)=0} \text{ and } S_{22}(f_0) = \left.\frac{\tilde{b}_2(f_0)}{\tilde{a}_2(f_0)}\right|_{\tilde{a}_1(f_0)=0}$$

The schematic corresponding to the definitions of $S_{12}(f_0)$ and $S_{22}(f_0)$ is therefore the following:



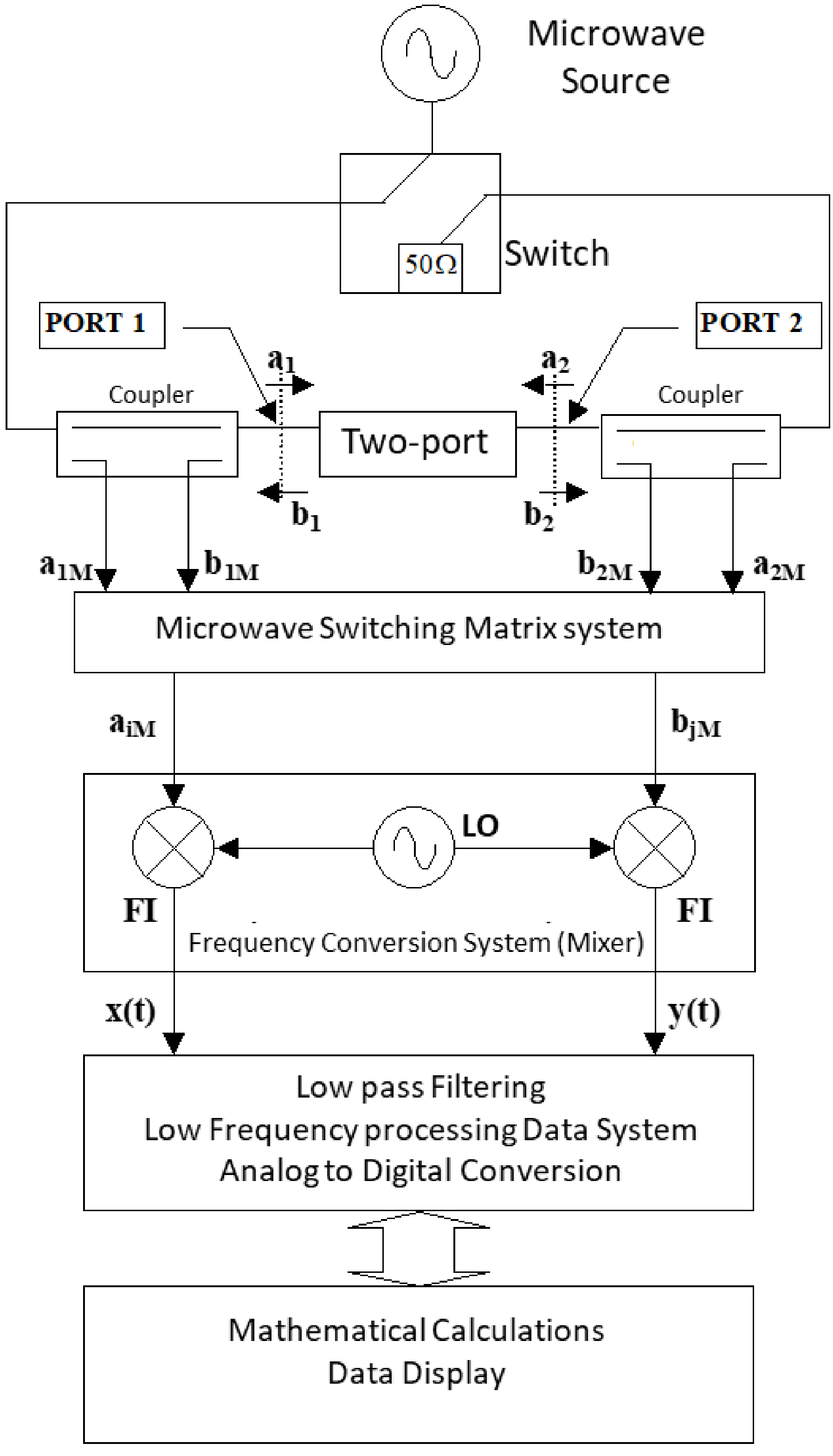
When the closing conditions are realized at the input and at the output, the characterization consists of measuring power wave ratios. These measurements must be vectorially performed (magnitude and phase) since each parameter S_{ij} is a complex quantity.

III.2 PRACTICAL IMPLEMENTATION

In practice, the measurement system must perform two essential functions:

1. To Realize the loading conditions at both ports of the Two-port device (definitions) and to scatter the different incident and reflected power waves in order to measure them separately,
2. To Perform frequency conversion for signal processing in the low frequency domain.

The general synoptic of a network vector measurement system is presented in the figure below.



VNA Fundamental Synoptic

A frequency synthesizer provides microwave signal generation.

A switch is used to send the microwave signal:

- ✓ either on the input of the two-port to be tested (port1) and to load its output with 50 Ω (port2). In this configuration, the loading conditions corresponding to the definitions of $S_{11}(f_0)$ and $S_{21}(f_0)$, that is to say $\tilde{a}_2(f_0) = 0$, are realized.
- ✓ either on the output of the quadrupole to be tested (port2) and to load its input on 50 Ω (port1). In this configuration, the closing conditions corresponding to the definitions of $S_{12}(f_0)$ and $S_{22}(f_0)$ are carried out, that is to say $\tilde{a}_1(f_0) = 0$.

Two bidirectional couplers allow to separate of the incident and reflected power waves. These power waves exist simultaneously on the same transmission medium (coaxial cable). The bidirectional coupler is a passive device using coupled lines. It allows to recover on the coupled channels a fraction of the incident and reflected waves present on its direct path.

a_{1M} is proportional to the wave a_1; b_{1M} is proportional to the wave b_1.

a_{2M} is proportional to the wave a_2; b_{2M} is proportional to the wave b_2.

A microwave switching system is then used to select two power waves from among the four outputs of the couplers. For example, to measure the wave ratio, the waves a_{1M} and b_{2M} are selected.

A frequency conversion system is used to translate the microwave signals to an IF intermediate frequency in the low frequency range (a few MHz). A low pass filtering is then required to eliminate the high frequency product at the output of the mixer and to only work with the low pass signal also created by the mixer. Current technology does not allow direct and simple processing of microwave signals. Frequency conversion is necessary to obtain LF signals to be easily processed and digitized without difficulty. The frequency conversion is actually carried out not by a single mixer by a chain of two and sometimes three mixers.

A LF signal processing system. This module provides the necessary filtering and amplification functions. In particular, it performs quantization and digitization of signals using analog / digital converters.

Finally, a computer system allows to manage the display, control and calculation functions.

- When the switch is positioned as shown in the previous figure, the system is configured for the measurement of $S_{11}(f_0)$ and $S_{21}(f_0)$. The switching system will select:
  - the power waves a_{1M} and b_{1M} to determine S_{11}. S_{11} is proportional to the ratio $\frac{b_{1M}}{a_{1M}}$.

$$S_{11} = \frac{b_1}{a_1} \propto \frac{b_{1M}}{a_{1M}}$$

  - the power waves a_{1M} and b_{2M} to determine S_{21}. S_{21} is proportional to the ratio $\frac{b_{2M}}{a_{1M}}$.

$$S_{21} = \frac{b_2}{a_1} \propto \frac{b_{2M}}{a_{1M}}$$

- When the switch is set to drive the Device Under Test (DUT) at port2 and to load port1 by 50Ω, the system will be configured to measure $S_{22}(f_0)$ and $S_{12}(f_0)$.. The switching system will select:
  - the power waves a_{2M} and b_{2M} to determine S_{22}. S_{22} is proportional to the ratio $\frac{b_{2M}}{a_{2M}}$.

$$S_{22} = \frac{b_2}{a_2} \propto \frac{b_{2M}}{a_{2M}}$$

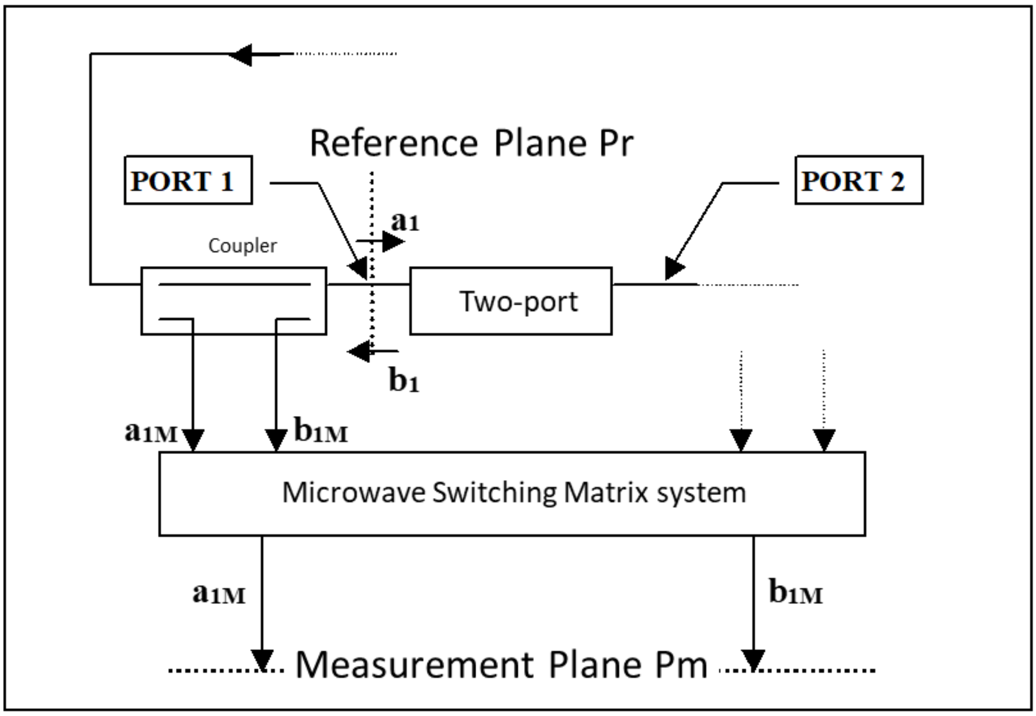
  - the power waves a_{2M} and b_{1M} to determine S_{12}. S_{12} is proportional to the ratio. $\frac{b_{1M}}{a_{2M}}$.

$$S_{12} = \frac{b_1}{a_2} \propto \frac{b_{1M}}{a_{2M}}$$

III.3 VNA CALIBRATION.

III.3.1 Why a calibration ?

Consider the microwave part of the measurement system for port 1:



Let assume that we want to measure the reflection coefficient $S_{11}(f_0) = \left.\frac{\tilde{b}_1(f_0)}{\tilde{a}_1(f_0)}\right|_{\tilde{a}_2(f_0)=0}$.

This reflection coefficient is defined in the reference plane Pr. The analyser, through the bidirectional coupler and the switching system, measures the reflection coefficient $S_{11m}(f_0) = \frac{\tilde{b}_{1m}(f_0)}{\tilde{a}_{1m}(f_0)}$defined in the measurement plane Pm.

Between the reference plane and the measurement plane, the power waves are attenuated, and phase shifted. The coefficient $S_{11m}(f_0)$ measured is therefore not equal to the desired coefficient $S_{11}(f_0)$. It is therefore necessary to correct the measured value to take into account the losses and phase shifts provided by the elements located between the reference plane and the measurement plane. The same situation is observed when measuring a transmission coefficient. In order to correct the measured value (or raw value), a calibration (or calibration) of the measuring system is carried out.

III.3.1 Calibration implementation.

The principle of calibration consists in characterizing the microwave equipment located between the Pr and Pm planes by determining complex coefficients called error terms. At a given frequency, these error terms relate the raw measurements in the Pm Plane to the corrected measures in the Pr plane by the following equation.

$$\begin{pmatrix} S_{11DUT}(f_0) \\ S_{12DUT}(f_0) \\ S_{21DUT}(f_0) \\ S_{22DUT}(f_0) \end{pmatrix}_{\mathrm{Pr}\, plane} = \begin{bmatrix} e_{00} & e_{01} & e_{02} & e_{03} \\ e_{10} & e_{11} & e_{12} & e_{13} \\ e_{20} & e_{21} & e_{22} & e_{23} \\ e_{30} & e_{31} & e_{32} & e_{33} \end{bmatrix} \begin{pmatrix} S_{11m}(f_0) \\ S_{12m}(f_0) \\ S_{21m}(f_0) \\ S_{22m}(f_0) \end{pmatrix}_{Pm\, plane}$$

Calibration therefore consists in determining the 16 unknown complex values e_{ij} of the error terms for each working f_0 frequency. To do this, we measure well-known microwave elements (one-port or two-port): these are calibration standards. Calibration standards are typically 50Ω loads, short circuits, open circuits, lossless lines. They are the different devices of the calibration Kit.

Due to a good isolation of the two ports of the VNA, the previous equation can be written as:

$$\begin{pmatrix} S_{11DUT}(f_0) \\ S_{12DUT}(f_0) \\ S_{21DUT}(f_0) \\ S_{22DUT}(f_0) \end{pmatrix}_{\mathrm{Pr}\, plane} = \begin{bmatrix} e_{00} & e_{01} & e_{02} & 0e^{j0°} \\ e_{10} & e_{11} & 0e^{j0°} & e_{13} \\ e_{20} & 0e^{j0°} & e_{22} & e_{23} \\ 0e^{j0°} & e_{31} & e_{32} & e_{33} \end{bmatrix} \begin{pmatrix} S_{11m}(f_0) \\ S_{12m}(f_0) \\ S_{21m}(f_0) \\ S_{22m}(f_0) \end{pmatrix}_{Pm\, plane}$$

Only 12 error terms can be determined thanks to a SOLT (Short Open Load Thru) or TOSM (Thru Open Short Match) calibration procedure).

From these standard measurements, the different error terms can be calculated. All procedures for calculating error terms and then correcting raw measurements are performed automatically by the network analyser. During the calibration process, the role of the user is limited to connecting the different standards and following the internal procedure of the analyser.

III.3.2 Calibration Kit.

The calibration kit used is for APC-7 type connectors. It consists of a short circuit, an open circuit and a suitable "broadband" load.

The open circuit is not perfect and behaves in fact like an capacitance. There is a capacitive effect between the two conductors at the end of an open coaxial cable. This imperfection is considered by four constant coefficients C_0, C_1, C_2 and C_3. The value of the equivalent capacitance C of the open circuit is modeled by a polynomial as a function of the frequency:

$$C = C_0 + C_1.f + C_2.f^2 + C_3.f^3.$$

These coefficients are well known for the calibration kit you are going to use. You will find them in the Microwave PW Room.

IV EXPERIMENTATIONS

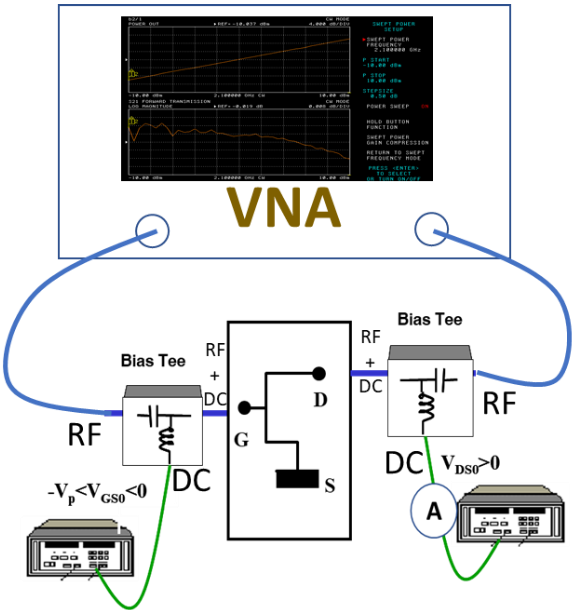
IV.1 S PARAMETER CHARACTERIZATION OF PASSIVE DEVICES

1. Reset the VNA.
2. Configure the analyser to display the 4 S parameters.
3. Perform a full (TOSM) calibration of the network analyser in the 1 to 3 GHz band.
4. Check this calibration from a direct connection between the 2 reference ports For this verification of the calibration, write the S matrix of a lossless 50Ω, 0mm length transmission line in different format: polar format (linear magnitude and phase, dB magnitude and phase), Smith Chart, Real and Imaginary Part.
5. Take many screenshots of the analyser with the 4 S parameters in different formats (polar format (linear magnitude and phase, dB magnitude and phase), Smith Chart, Real and Imaginary Part) with 3 coupled markers at 1GHz, 2GHz and 3 GHz.
6. Comment the results.
7. Perform the measurement of a passive device provided in the PW Room by the teacher.
8. Take many screenshots of the analyser with the 4 S parameters in different formats (polar format (linear magnitude and phase, dB magnitude and phase), Smith Chart, Real and Imaginary Part) with 3 coupled markers at 1GHz, 2GHz and 3 GHz.

IV.2 S PARAMETER CHARACTERIZATION OF AN ACTIVE DEVICES

IV.2.1 FET Biasing.

A classical bias scheme of a field effect transistor is as follows:



The DC Block capacitances and the DC Feed Inductances are embedded into the bias tee.

It is FUNDAMENTAL to correctly bias the transistor according to a methodology that will be described to you during the session.

The following tasks should be performed in the process:

1. Check the polarity of the power supplies: $V_{gs0} < 0$ and $V_{ds0} > 0$

   Common Ground to both power supplies.
2. Adjust the short circuit currents for the gate and the drain
3. Adjust the power supply voltages before connecting them to the TEC at the values:

   $$V_{gs0} = 0V \text{ and } V_{ds0} = 0 \text{ V}$$
4. Connect the gate power supply to the FET.
5. Connect the drain supply to the FET.
6. Increase the gate voltage first to the selected value, depending on the transistor and the polarization class
7. - Increase the drain voltage secondly to the desired value.

MANDATORY !!!: Do not increase V_{DSO} when V_{GSO}=0V. The avalanche on the static characteristic at V_{GSO} =0V is unexpected and irreversible.

To unbias the transistor, perform the following tasks in order:

1. Decrease V_{ds0} to 0 V and then decrease V_{gs0} to 0V.
2. Disconnect the drain and gate ports respectively
3. Switch off the power supplies.

MANDATORY: Turn off the power supplies before disconnecting the TEC...

In all cases, ask first the teacher to define and justify the limits of the potential source power variations without risk of damage of the FET.

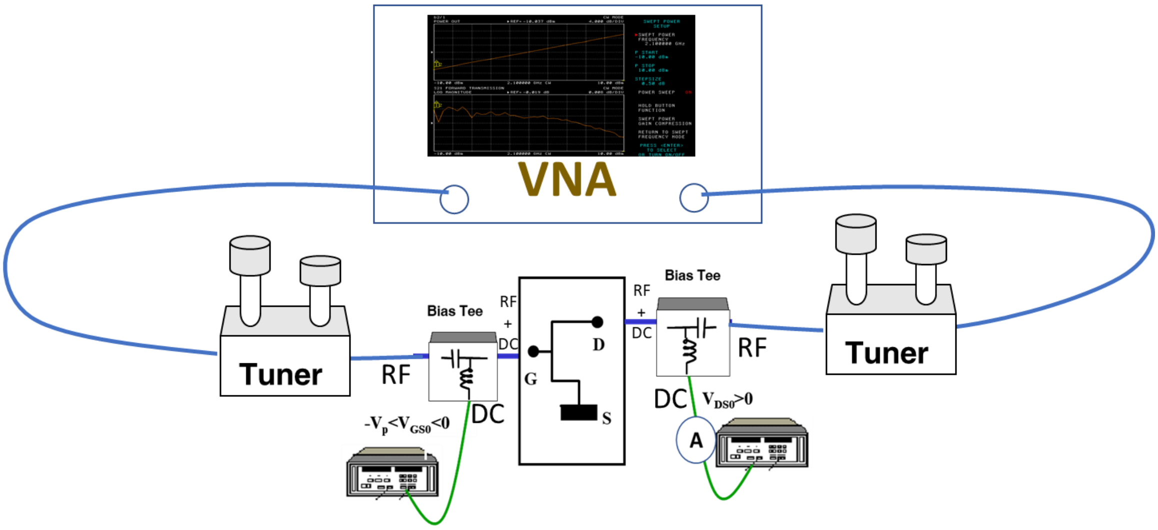
IV.2.2 Large Bandwidth FET S Parameter Measurements.

1. Perform the measurement of the FET.
2. Choose the biasing conditions to compare your results with the performances given in the datasheet. Use a lower V_{DS0} voltage (3 V) to avoid thermal issues.
3. Take many screenshots of the analyser with the 4 S parameters in different formats (polar format (linear magnitude and phase, dB magnitude and phase), Smith Chart, Real and Imaginary Part) with 3 coupled markers at 1GHz, 2GHz and 3 GHz.
4. Compare to the values given in the datasheet.
5. Comments.
6. Perform the depolarization of the transistor

IV.2.3 Input/output Matching of the FET @2GHz.

1. Remove the transistor.
2. Reset the VNA
3. Configure the analyser to display the 4 S parameters.

4. Perform a full (TOSM) calibration of the network analyser in the 1.95 to 2.05 GHz band.
5. Check this calibration from a direct connection between the 2 reference ports For this verification of the calibration, write the S matrix of a lossless 50Ω, 0mm length transmission line in different format: polar format (linear magnitude and phase, dB magnitude and phase), Smith Chart, Real and Imaginary Part.
6. Take many screenshots of the analyser with the 4 S parameters in different formats (polar format (linear magnitude and phase, dB magnitude and phase), Smith Chart, Real and Imaginary Part) with 3 coupled markers at 1GHz, 2GHz and 3 GHz.
7. Comment the results.
8. Use the following configuration of the test bench with the VNA:



The slug of the tuners must be raised (unscrewed)

9. Perform the [S] measurements of the unmatched transistor.
10. Take many screenshots of the analyser with the 4 S parameters in the following configuration dB magnitude and phase with 1 coupled markers at 2GHz.
11. Comments
12. Move the slugs of the tuner at the input of the transistor to match it at its input.
13. Perform the [S] measurements of the matched input/unmatched output transistor.
14. Take many screenshots of the analyser with the 4 S parameters in the following configuration dB magnitude and phase with 1 coupled marker at 2GHz.

15. Move the slugs of the tuner at the output of the transistor to match it at its output. Check the matching at the input and optimize the input and matching output finding the better trade-off between input and output.
16. Take many screenshots of the analyser with the 4 S parameters in the following configuration dB magnitude and phase with 1 coupled marker at 2GHz.
17. Comments.
18. Perform the depolarization of the transistor.
19. Perform the |S] measurements of the two tuners without moving the slugs. What are these measurements useful for ?