

# LEARNING HOW TO USE A VECTOR NETWORK ANALYZER (VNA)

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IW2FND Attolini Lucio Via XXV Aprile, 52/b 26037 San Giovanni in Croce (CR) [www.iw2fnd.it](http://www.iw2fnd.it)

## 1. Premise

Following the great availability on the market of the ultra-fast direct digital synthesizer (DDS) and the logarithmic amplifiers, the production of the vector analyzers (VNA) has gained considerable momentum, to the point that they even became accessible to the radio-amateurs for no more than some tens of euro (see the NanoVNA or the vector antenna analyzers). Nowadays, on the market, people can find VNAs of any kind: single or double port, vectorial, PC based, portable and more... During the 80s, when I started my activity, they were rather bulky, with a prohibitive price and only accessible for industrial or university research purposes.

Given that the device in question is not exactly user friendly, my objective with this article is to explain its functioning and potential. To do so, I necessarily have to introduce certain basic notions, as they are a pre-requisite to the comprehension and use of the device. Particularly, I am referring to the S-scatter parameters and the Smith chart; the topics that are discussed hereafter won't be exposed in a systematic way, in fact my explanations will only be limited to what's strictly necessary for the scope of this essay. Furthermore, I will also adopt many "simplifications" (although always maintaining mathematical rigor), simplifications which will serve the purpose of making the trickiest concepts more accessible to a wider audience. Lastly, I need to specify that the following article is not applicable to the spectrum analyzers, since they are based on completely different principles compared to the VNA.

## 2. S-scatter parameters

When you analyze a network, the use of S-scatter parameters (hereafter, just S) comes extremely handy because they just consider the forward tension waves and the reflected ones. From the measures of those two types of waves one can obtain the fundamental parameters (reflection coefficient, impedance, SWR and so on...) of the device under test (DUT from now on). Of course, we will have to consider the DUT, or, more generally, a linear type of network - invariant throughout time and with sinusoidal waves - to be able to represent the sinusoids with vectors (phasors) and complex numbers. This is true because the S parameters are vector quantities (in this text we will express vectors with a **bold** letter). I want to remind you that a vectorial quantity is representable with a vector (which gives direction and intensity, in addition to the value) on the Cartesian plane, the length (module) and the angle (positive counterclockwise) of which must be specified with respect to the x axis (from 0 to 360 degrees). If we switch from the Cartesian plane to the complex numbers plane, the horizontal axis will represent real numbers, while the vertical axis will show the imaginary ones. From the graphical point of view nothing changes, but the advantage is that we can exploit complex number mathematics to carry out all the calculations.

A scalar quantity, on the other hand, is only represented by its value; for example, temperature is indicated only by its value in degrees.

The S-scatter parameters lend themselves very well to the network exam in the radiofrequency field and the microwaves because the forward and reflected waves can be easily measured inside the VNA, both in amplitude and in phase, without interfering with the device under test (everyone knows the functioning principle of the ROS-meter). From now on, by mentioning wave, we will assume it is a tension wave.

The simplest device to measure is the one constituted by only one port, like an antenna or a simple load at the end of a line.

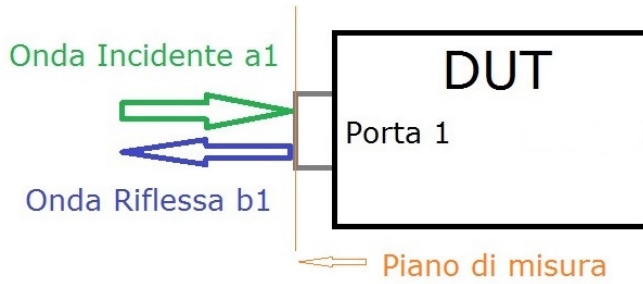


Figure 2.1: One port device

Figure 2.1 shows the DUT with its input connector, where the incident wave **a1** (onda incidente) comes in (took as a reference) and the wave **b1** gets reflected (onda riflessa): both of those are sinusoids represented by the complex numbers **a1** and **b1**.

Conventionally we consider the plane where the reflection phenomena happen as “plane of measure” (piano di misura); we will deepen this important aspect in a later moment.

Having said so, let's define the scatter parameter **S<sub>11</sub>** in the following way:

$$\mathbf{b1} = \mathbf{S_{11}} \cdot \mathbf{a1} \quad 2.1$$

Therefore, the wave reflected on port 1 (first number as subscript of S) is equal to the incident wave on port 1 (second number as subscript of S) multiplied by the scatter parameter **S<sub>11</sub>**; it follows that:

$$\mathbf{S_{11}} = \frac{\mathbf{b1}}{\mathbf{a1}} \quad 2.2$$

**S<sub>11</sub>** can be obtained by the ratio of the two complex numbers **b1** and **a1** which represent respectively the reflected and incident wave on the measurement plane of port 1 of our DUT; it basically is the reflection coefficient **Γ**.

$$\mathbf{S_{11}} = \frac{\mathbf{b1}}{\mathbf{a1}} = \mathbf{\Gamma} \quad 2.3$$

The reflection coefficient **Γ** is a complex number that roughly represents how well the load is matched to the line that powers it. Its module can take different values which range from 1 (total reflection) to infinitely small values (when the load is perfectly matched), to the point where it is preferable to express them in logarithmic units of base 10 (decibel).

$$|\mathbf{\Gamma}| = 20 \log |\mathbf{S_{11}}| \text{dB} \quad 2.4$$

The module of the reflection coefficient **Γ** can also be expressed with the lowercase Greek letter ρ (rho).

Moreover, from the parameter **S<sub>11</sub>** one can also obtain some more data: the SWR, the RL, the impedance of the DUT **Z** (if the impedance specific to the supply line **Z<sub>0</sub>** is known) and many others. Basically, all the most important parameters for the device characterization are under examination [1].

$$\mathbf{SWR} = \frac{1+|\mathbf{S_{11}}|}{1-|\mathbf{S_{11}}|} = \frac{1+\rho}{1-\rho} \quad 2.5$$

The SWR is the stationary wave ratio and it is a scalar quantity (not a vector) whose value never goes below the unit, but can, on the other hand, be quite big in value.

$$\mathbf{RL} = 10 \log \frac{1}{|\mathbf{\Gamma}|^2} = 10 \log \frac{1}{|\mathbf{S_{11}}|^2} = -20 \log |\mathbf{S_{11}}| = -20 \log \rho \text{ dB} \quad 2.6$$

The return loss (RL) expresses how much the reflected wave gets reduced compared to the incident one on the measurement plane: it is the inverse of the **S<sub>11</sub>** module. The return loss is a simple indicator of the level of impedance match. Having a return loss RL=0dB means a total mismatch, while RL= infinite equals a perfect match. As a matter of fact, to obtain a perfect match, an impedance equal or greater than 50dB is enough.

$$\mathbf{Z} = \mathbf{Z}_0 \frac{1+\mathbf{S}_{11}}{1-\mathbf{S}_{11}} \Omega \quad 2.7$$

Being the impedance  $\mathbf{Z}$ , shown on the measurement plane, a complex number, it can be decomposed in its resistive and reactive components, R and X respectively, or with its inverse, the admittance  $\mathbf{Y}$ .

$$\mathbf{Z} = R + jX \Omega \quad 2.8$$

Because the values of R and X are known, the unloaded factor of merit  $Q_u$  can be calculated as well:

$$Q_U = \left| \frac{X}{R} \right| = \frac{2\pi f L}{R} = \frac{1}{2\pi f C R} = \frac{\sqrt{L/C}}{R} \quad 2.9$$

Where the parameters L, C and R are all referring to the series models of the respective components: inductor, condenser, and resonant circuit LC. Moreover, knowing the sign of the X and the frequency, one will also know the value of L and C:

$$L = \frac{X}{2\pi f} \text{ e } C = -\frac{1}{2\pi f X} \quad 2.10$$

It seems like the calculations are very convoluted, but they will get automatically executed by the VNA because it measures the module of the two waves: direct (**a1**) and reflected (**b1**).

A second typology of device to measure is the one constituted by two ports, be it a filter, an amplifier or more simply a transmission line.



Figure 2.2: Two ports device

Figure 2.2 shows a two port DUT, with its input and output connectors. The incident wave **a1** (taken as a reference) gets to port 1 where the reflected wave **b1** is generated, just like we saw previously in the case of the 1 port DUT when the  $\mathbf{S}_{11}$  is measured; the wave **b2** goes out from the second port without any reflection (onda trasmessa).

Let's now define the scatter parameter  $\mathbf{S}_{21}$  in the following way:

$$\mathbf{b2} = \mathbf{S}_{21} \cdot \mathbf{a1} \quad 2.11$$

Meaning the wave that goes out from port 2 is given by the incident wave on port 1 multiplied by the scatter parameter  $\mathbf{S}_{21}$ ; it follows that:

$$\mathbf{S}_{21} = \frac{\mathbf{b2}}{\mathbf{a1}} \quad 2.12$$

In other words, it is a complex number that indicates how much the incident wave gets attenuated or amplified while traversing the DUT, in absence of reflections on port 2. Meanwhile, on port 1, the reflections can still be present and can be measured with the scatter parameter  $\mathbf{S}_{11}$ . I want to point out that in figure 2.2 there are 2 planes of measure and not just one.

The  $\mathbf{S}_{21}$  parameter is actually the transmission coefficient  $\mathbf{T}$  [2]:

$$\mathbf{T} = \mathbf{S}_{21} \quad 2.13$$

The transmission coefficient  $\mathbf{T}$ , whose module is expressed in dB, is known as a tension gain  $G_V$ :

$$G_V = -20 \log |\mathbf{T}| \text{ dB} \quad 2.14$$

The module of its reciprocal is the known Transmission Loss or Insertion Loss, expressed in dB. We will use the insertion Loss IL, defined in the following way [2]:

$$IL = TL = 20 \log \frac{1}{|S_{21}|} = -20 \log |S_{21}| \text{ dB} \quad 2.15$$

From the  $S_{21}$  we can obtain other parameters as well, like the phase delay and the group delay, but as for now, we are satisfied with the ones we have. If we sent the incident wave to port 2, instead of port 1, we would obtain other S parameters, similar to the previous ones, but with different subscripts; see figure 2.3.



Figure 2.3: Porte scambiate

$$S_{22} = \frac{b_2}{a_2} \quad 2.16$$

$$S_{12} = \frac{b_1}{a_2} \quad 2.17$$

Exchanging the ports is useful when examining asymmetrical DUTs, like amplifiers. When the two ports are not perfectly matched, the complete response, on the respective measurement planes, can be found by solving the following system (scatter matrix):

$$\begin{cases} b_1 = S_{11} \cdot a_1 + S_{12} \cdot a_2 \\ b_2 = S_{21} \cdot a_1 + S_{22} \cdot a_2 \end{cases} \quad 2.18$$

The system can be extended to other ports by applying the same criteria: the ones that are not involved in the measure do not reflect anything. If we consider an example with 3 ports, we will have the scatter parameters  $S_{33}$ ,  $S_{13}$ ,  $S_{23}$ ,  $S_{31}$  and  $S_{32}$ ; I am stressing this, because sometimes these parameters are used to characterise combines, splitters and BalUns. In this essay we won't go as far and we will deal only with 1 and 2 port DUT, even though the number of ports our device (or net) has is only of limited importance; this is because the ports will always be examined individually or in pairs to generate the respective  $S$  parameters. By doing so the ports of the DUT that are not under examination will have to be terminated with an impedance that does not generate reflections; this can be obtained with a termination impedance that is the complex and conjugated of the one seen beyond the measurement plane. In the majority of cases, it is  $Z_0=50+j0$  Ohm, meaning a good quality dummy load.

### 3. The Smith chart

The Smith chart (1905 - 1987) was invented by the homonymous author in 1939 and perfected in the following years: the form that we know today can be traced back to 1949 [3]. The author created it in an historical time in which there weren't any scientific calculators, nor computers; it was the era of slide rules and nomograms. As a matter of fact, at that time equations that regulate phenomena regarding lines were already known, but they were of difficult resolution. Therefore, Smith had the intuition to relate the lines impedance equation with the reflection coefficient  $\Gamma$ , which is a complex number. The outcome were three circumference bundle: one with the centres on the x axis, where there will be the circumferences that share the same real part, while the other two are symmetrical and have centres on the y axis. The only common point to all the circumferences is the axis origin. The circumferences above the x axis have the same positive reactance, while the ones below have negative reactance. All the complex impedances  $Z$  that a line can take with a positive real part will be represented inside the unit radius circumference (the outermost, black coloured one in figure 3.1). Outside the black circumference in figure 3.1 there are impedances with negative real part which are not useful to us.



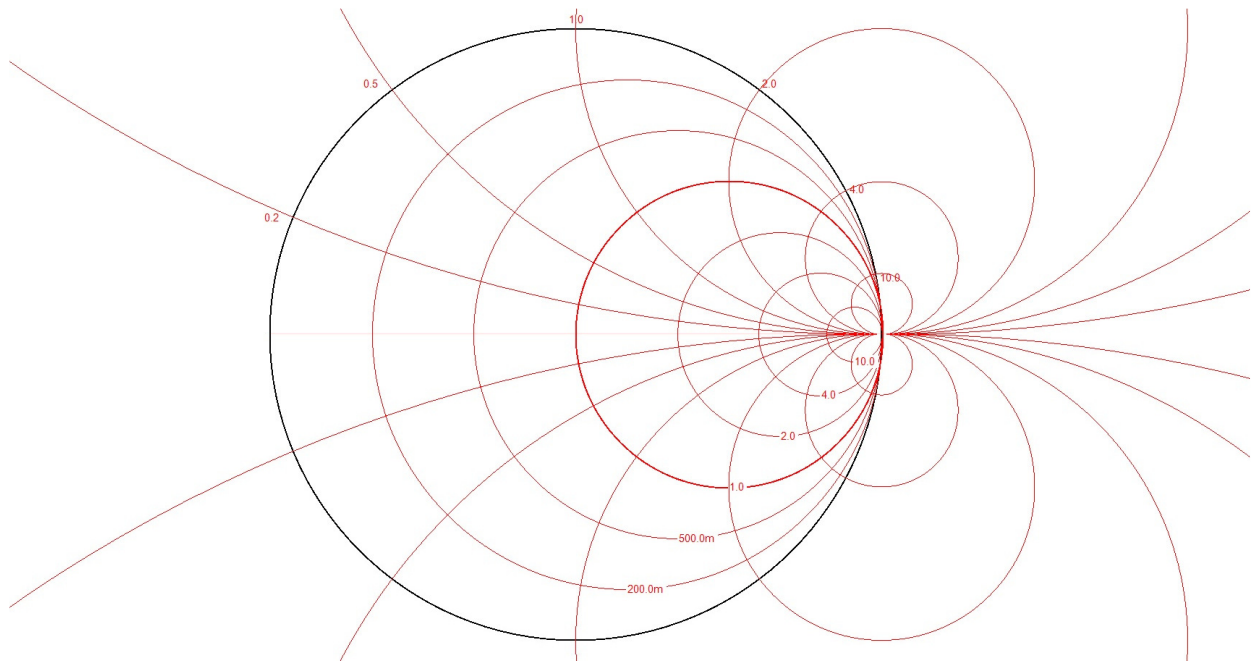


Figure 3.1: Bundle of circumferences

For this reason, from now on we will only consider and represent the circumferences and circumference arcs contained within the borders of the black unit radius circumference shown in figure 3.1, therefore only the ones with a positive real part.

In order to make it as generic as possible, The Smith chart has been normalised to the unit. This means that the central point “C” of figure 3.2 corresponds to the characteristic impedance  $Z_0$  of the line under examination, usually of 50 Ohm. All the impedance values that we will draw on the chart will have to be divided by the module of the characteristic impedance  $Z_0$  of the line used to supply the energy. Therefore, the coordinates of the points that we’ll draw on the chart will be the result of the following equation:

$$Z_N = R/Z_0 + j X/Z_0 \quad 3.1$$

On the other hand, to reverse it to the non-normalised form we will have to multiply the same equation by  $Z_0$ :

$$Z = R_N \cdot Z_0 + j X_N \cdot Z_0 \quad 3.2$$

For example, if we wanted to represent the impedance  $Z=10+j25$  normalised with a line at  $Z_0=50\Omega$ , using the formula 3.1 we would obtain:

$$Z_z = 10/50 + j 25/50 = 0,2 + j0,5 \quad 3.3$$

The point “Z” on the Smith chart will have coordinates (0,2 ; 0,5), as shown in figure 3.2. The central point “C” on the normalised Smith chart is always (1 ; 0), which represents the perfect fit to the line  $Z_0$ ; point “A” on the left always represents the short circuit (0 ; 0) and the one to the right “B” (common point to all the circumferences) always indicates the open circuit ( $\infty$  ;  $\infty$ ), with infinite resistance. On the central diameter that goes from “A” to “B” lay all the purely resistive impedances, meaning the ones that have null imaginary part ( $X=0$ ), while the red circumferences that pass through said diameter (figure 3.2) are circumferences that have constant resistive part  $R$ . Travelling along one of these circumferences will change the reactive part  $X$  of the impedance, while the resistive part  $R$  will remain unchanged.

The chart was created to simplify operations with transmission lines. From the theory, we know that if we move along a line without losses, starting from its load towards the generator, we find the same impedance “Z” for each electrical half lambda. On the Smith chart the process is clearly visible; the load of the line without losses is represented by the point “Z” in figure 3.2. Moving along the blue circumference in a clockwise direction, the more we get away from the load  $Z$  towards the generator, the more the impedance will change. Every half wave ( $\lambda/2$ ) the impedance will equal the load  $Z$  again. On the chart it equals to

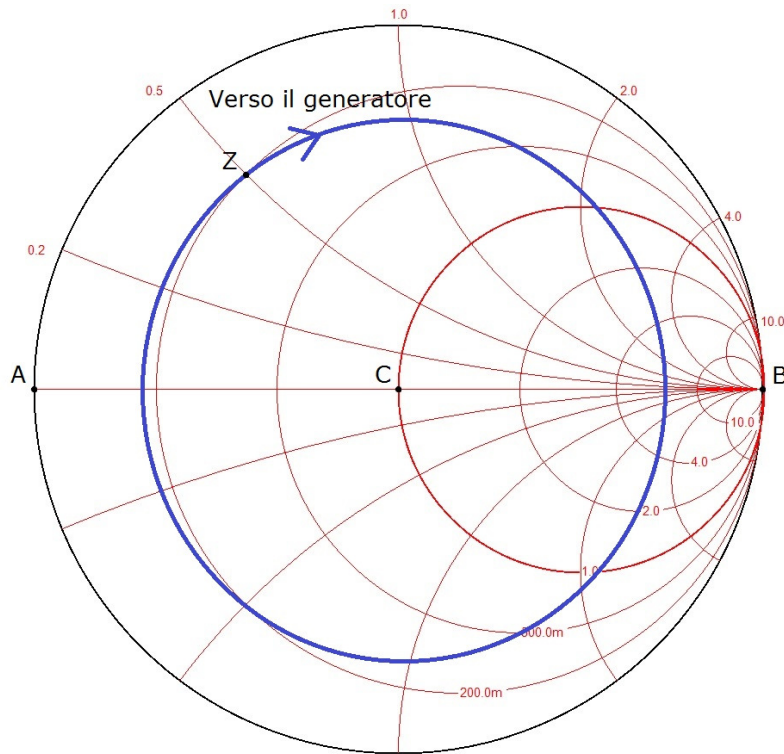


Figure 3.2: Impedance Smith chart

moving along the blue circumference of figure 3.2 in a clockwise direction from point “Z” back to point “Z”. The blue circumference has constant reflection coefficient  $\Gamma$  (that’s because we hypothesised a line without losses) and the radius of the blue circumference is also the well know SWR. It goes without saying that if we only cover half of the circumference, we will have covered  $\lambda/4$  of the line.

You might have noticed that I’ve always talked about  $\lambda$  without specifying its value; this is because another characteristic on the Smith chart is its independency from the frequency used, making it even more universal. Naturally, when we will put it to work, we will need to assign a value to the electric wavelength according to the known formula:

$$\lambda = c/f_m \quad \mathbf{3.4}$$

The electric wavelength  $\lambda$  ( $\lambda$  without subscript) is expressed in meters if the speed of light is  $c=300 \cdot 10^6$  m/s and if the frequency  $f$  is expressed in Hz (in practice we use  $c=300$  and  $f$  in MHz).

With transmission lines, and the dielectric being not equal to the air, the wave speed diminishes and therefore, to find the physical wavelength  $\lambda_f$  ( $\lambda$  with  $f$  subscript) in the line we will have to multiply the electric wavelength by the speed factor  $VF$  (formula 3.5).

$$\lambda_f = VF\lambda_m \quad \mathbf{3.5}$$

The speed factor  $VF$  depends on the dielectric placed between the two conductors.

It all seems complicated, but this accuracy is needed to be able to use the chart with every line or frequency and carry out complicated calculations with just ruler and compass. Of course, if we move along the blue circumference of figure 3.2 in an anti-clockwise sense, we will move from the generator towards the load (verso il generatore). Another choice can be to look up a Smith chart at  $50 \Omega$  (or not normalised) online; in that case the central point will be  $50 \text{ Ohm}$  and it could be used only with lines having a characteristic impedance of  $Z_0=50\Omega$ . The most complete Smith chart is the one published by the author in 1949, which you can see in figure 3.3:



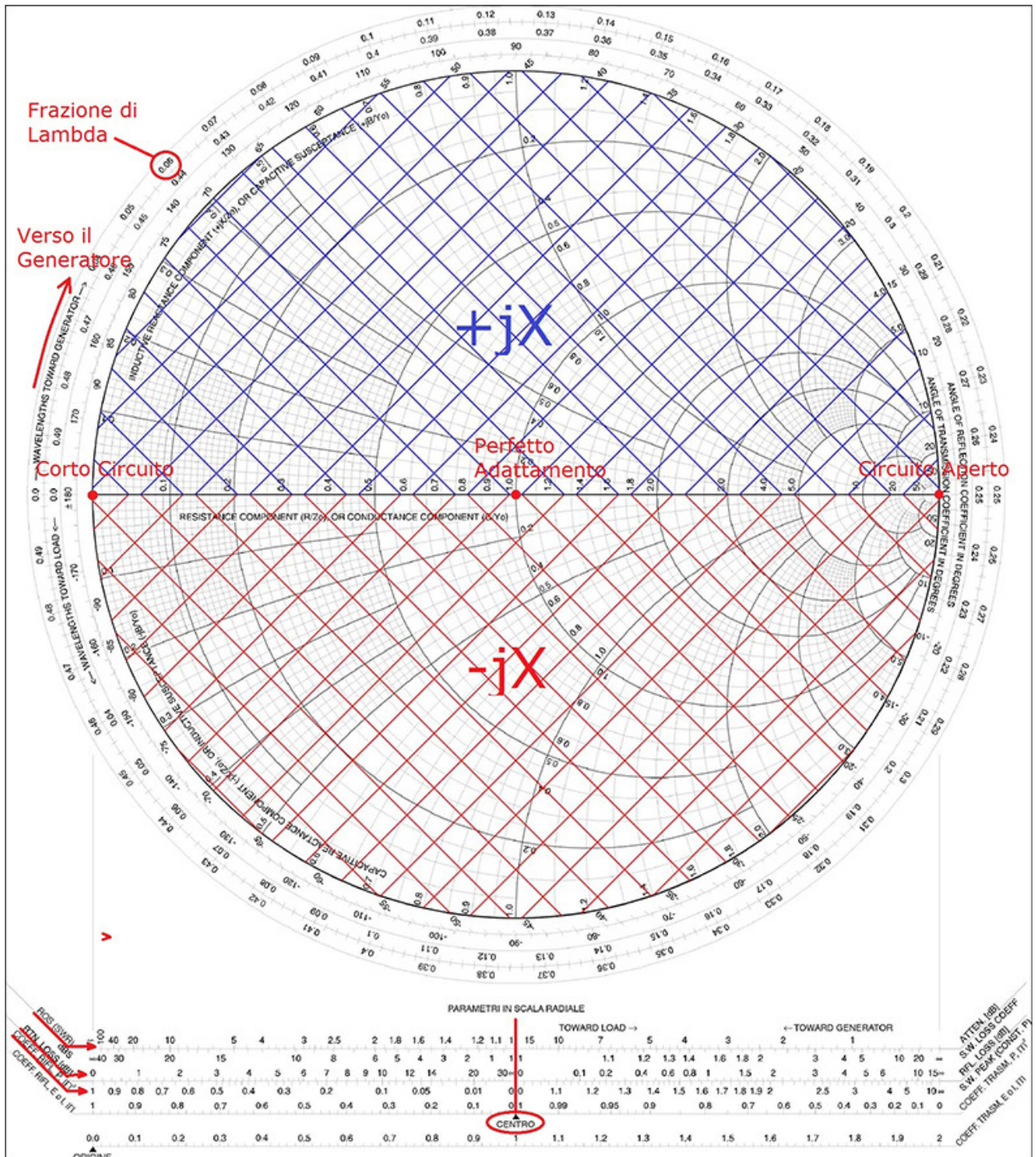


Figure 3.3: Smith chart of impedances (1949)

The amount of information contained in figure 3.3 might be confusing, but the most important ones are the ones on the external circumference, because it shows the fractions of lambda (frazione di lambda). The first circumference represents the direction towards the generator, while the second shows the one towards the load. On the central axis we can find the values of the real part (pure resistances) of the impedance and on the bold circumference we will find the arcs with the pure reactance values. Each impedance, both with resistive and reactive part, is represented by the intersection between the circumference of its resistance and the arc of its reactance. For this reason, in the upper half of the circle we'll have all the positive reactances (blue dashed line), while in the lower half of the circle there will be all the negative ones. On the central axis of figure 3.3 I indicated the main points: short circuit on the left, perfect match in the middle and open circuit on the right. Below figure 3.3 you

can find a regulus with numerous parameters, such as: the module of the reflection coefficient, the return loss RL (RTN in figure 3.3) and the SWR (ROS in figure 3.3). The distances on the regulus are always to be taken as a centre, which corresponds to the centre of the Smith chart.

To fix the concepts I've just explained, let's make an example.

Let's suppose we're measuring an impedance of  $Z=30.8 - j5.5$  Ohm at our antenna connector inside our shack, that our antenna is working at 1.83MHz and that the coaxial supply cable, which is 20m long, had  $Z_0=50$  with speed factor  $VF=0.66$  (without losses). We would like to know the impedance of the antenna at the end of the supply line.

Firstly, we need to calculate the length of the electric wave of the sinusoid at a frequency of 1.83MHz. We can do that with the formula 3.4:

$$\lambda = c/f = 300/1,83 = 163,93m \quad 3.6$$

Once we have that, we then need to calculate the electric length  $l_e$  of the coaxial cable, by dividing the physical cable length by the VF:

$$l_e = l_f/VF = 20/0,66 = 30,30m \quad 3.7$$

The electric length has increased because the wave travels at a slower pace inside the cable. Now let's express the electric length of our cable as a fraction of the wavelength  $l_{\%}$ , at the desired frequency:

$$l_{\%} = l_e/\lambda = 30,30/163,93 = 0,184 \quad 3.8$$

Let's normalise the impedance  $Z=30.8 - j5.5$  at 50 Ohm with the 3.1:

$$Z_A = R/Z_0 + j X/Z_0 = 30,8/50 - j 5,5/50 = 0,61 - j0,11 \quad 3.9$$

Let's mark point A (0.61; -0.11) on the Smith chart, which represents our starting point. Now, let's trace a radius, which, starting from the centre of the chart, passes through point A and intersects with the outside circumference Toward Load; if we read the value of the wavelength, we will certainly find a value of  $\lambda_A=0,028$ . To find the ending point  $\lambda_B$ , which corresponds to the terminal of the coaxial line, we will have to add the length of the cable expressed as a fraction of wavelength  $l_{\%}$  to  $\lambda_A$ :

$$\lambda_B = \lambda_A + l_{\%} = 0,028 + 0,184 = 0,212 \quad 3.10$$

Then, with a compass, we have to draw an arc of a circumference (with constant  $\rho$ ), with center coinciding with the center of the chart. Said arc of a circumference has to pass through point A and intersect the radius ending in  $\lambda_B$ . The latter intersection will become point B of coordinates (1.54; -0.4), which corresponds to a normalized impedance  $Z_B=77-j20\Omega$ .

This is only an exercise for practice purposes, I didn't even trace it to allow everyone to solve it alone and get more acquainted with handling the chart.

If it was the case of an antenna working at 7100 kHz instead of 160m, things would radically change. The electric wavelength would in fact be:

$$\lambda = c/f = 300/7,10 = 42,25m \quad 3.11$$

The fraction of lambda would become:

$$l_{\%} = l_e/\lambda = 30,30/42,25 = 0,717 \quad 3.12$$

Value that ends up being over 0.5, therefore removing an entire turn ( $\lambda/2=0,5$ ) from the fraction is necessary to bring the value back to being lower than 0.5 and to be able to find it on the external crown of the chart.

$$l_{\%} = 0,717 - 0,5 = 0,217 \quad 3.13$$

Point B will intersect the radius that passes through the center of the chart and  $\lambda_B$ , which is:

$$\lambda_B = \lambda_A + l_{\lambda} = 0,028 + 0,217 = 0,245 \quad 3.14$$

The intersection of the radius towards  $\lambda_B$  with its circumference passing by A meets point B, of coordinates (1.67; 0.05), corresponding to a non-normalized impedance  $Z_B=83,5-j2,5\Omega$ : an antenna almost at resonance.

The example teaches us that every time we put a line segment without losses between the load  $Z$  and the generator, the impedance of the load seen by the generator changes significantly at the changing of the length of the line, but every  $\lambda/2$  it goes back to be



equal to  $\mathbf{Z}$ . On the chart it equals moving along a circumference having center in the center of the chart and passing by point A (which is  $\mathbf{Z}$  normalized); after every turn on said circumference we go back at the initial point A, which corresponds to a line segment that is  $\lambda/2$  electric long. The radius of that circumference is the module of the reflection coefficient  $\rho$  of the load (or the SWR, if you prefer), module that remains constant all along the line segment (hypothesized ideal) up until the generator. Since the reflection coefficient  $\Gamma$  is a complex number, not only does it have a module  $\rho$ , but also a phase: the phase can be read on the outside circumference, in correspondence of the extension of the radius, which passes by point A, in our case being  $-160^\circ$  (green in figure 3.4). Moving along the line the phase changes, while the module  $\rho$  doesn't; for this reason, the SWR measured along the ideal transmission line remains constant along the entirety of it.

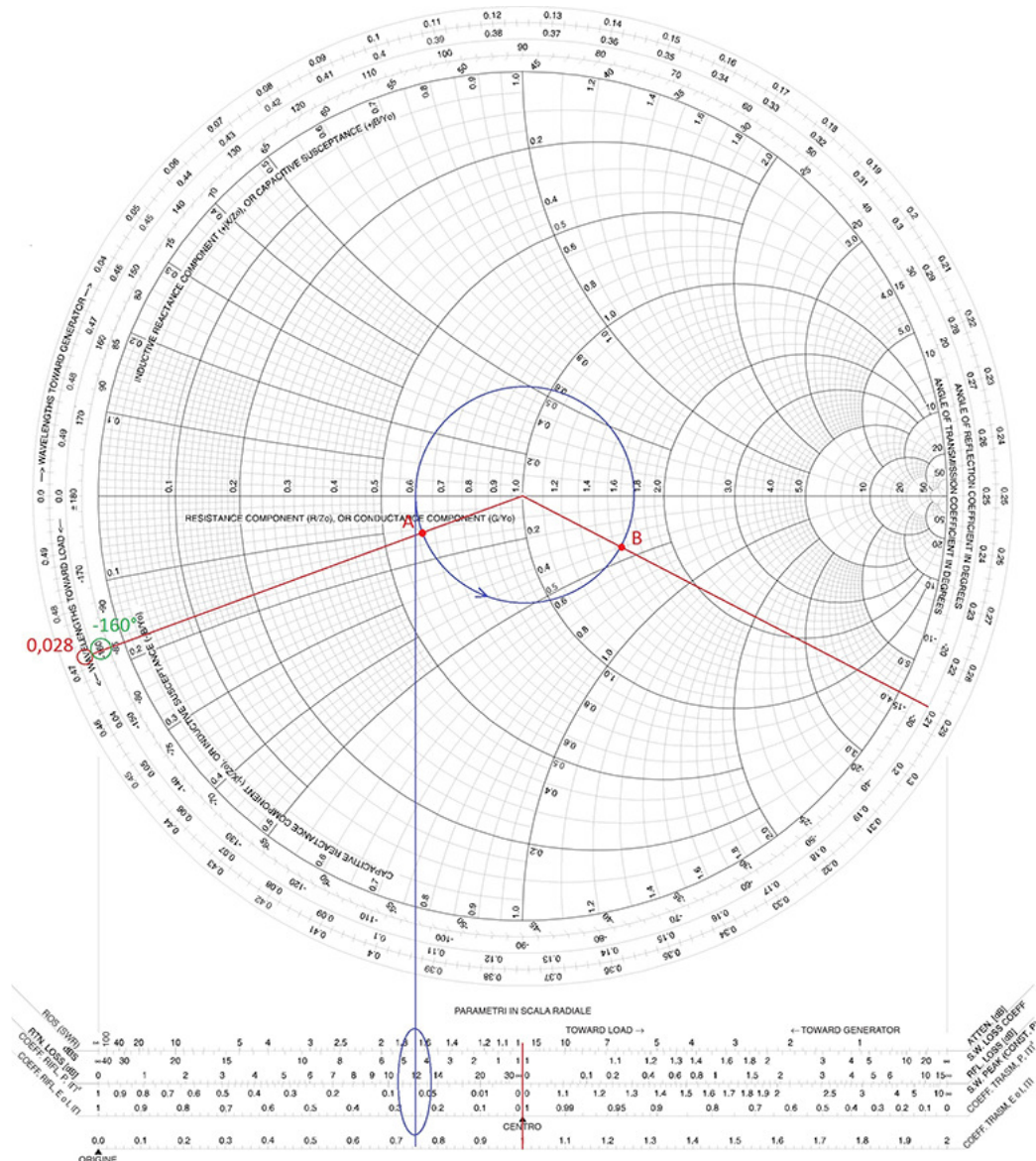


Figure 3.4: Example at 1830kHz

Up until now we have spoken in terms of complex impedance  $\mathbf{Z}$ , assuming that the model was resistance and reactance in series (Thevenin), while it is possible to represent the impedance with its equivalent parallel circuit (Norton). In this case we'll talk about admittance  $\mathbf{Y}$ , conductance  $\mathbf{G}$  and susceptance  $\mathbf{B}$  (see figure 3.5).

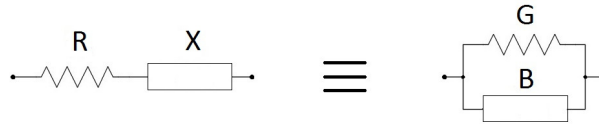


Figure 3.5: Series and parallel impedance

The two models are perfectly equivalent, and it is possible to move from one to the other with the following formulas:

$$G = \frac{R}{R^2 + X^2} \quad B = \frac{-X}{R^2 + X^2} \quad Y = G + jB \quad 3.15$$

While the inverse formulas are:

$$R = \frac{G}{G^2 + B^2} \quad X = \frac{-B}{G^2 + B^2} \quad Z = R + jX \quad 3.16$$

These are not impossible formulas; they just require some calculations. The Smith chart allows for the transition from a series configuration to a parallel one in a graphical way, without requiring any calculation, since as I mentioned before, it was created in a period of time where calculators didn't exist yet. The method is simple, it is just enough to find the point of interest with respect to the center of the chart (figure 3.6).

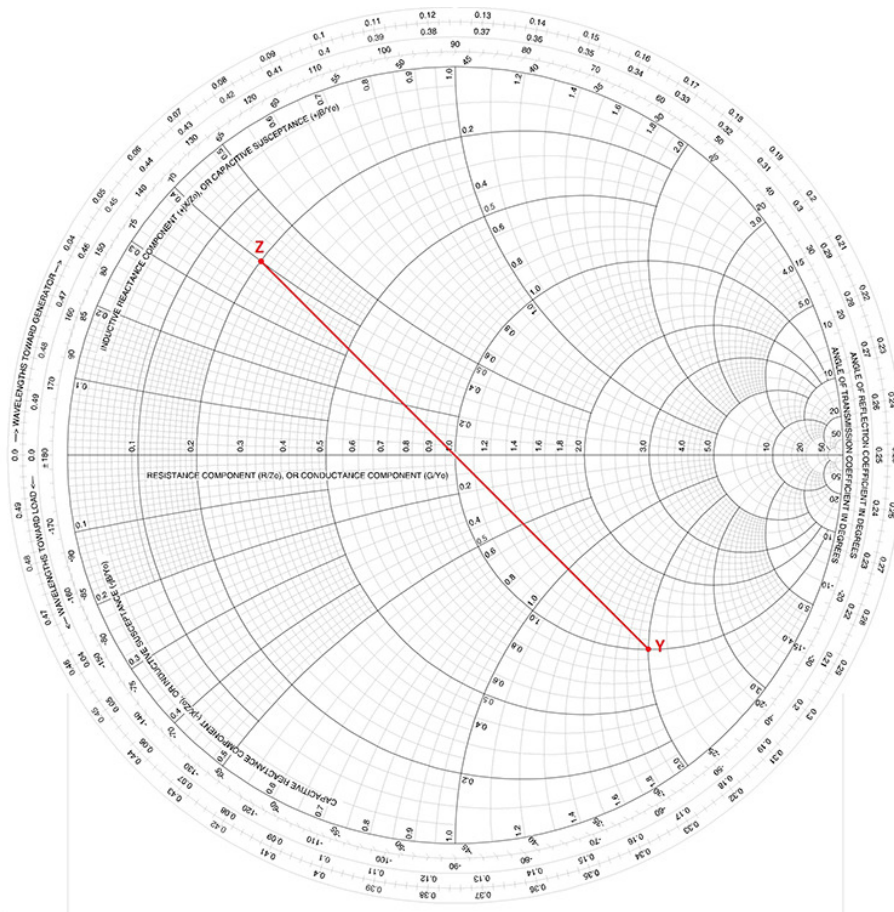


Figure 3.6: Transition Impedance - Admittance

Let's consider point  $Z = 0.2 + j0.4$  (normalized values), which, on the chart, is located on the upper semi-circumference to the left, then let's trace the opposite with respect to the center. After doing the calculations with 3.16, we will find on the chart that the result will be  $Y = 1 - j2$ , exactly like the equivalent reactance.

The position of **Y** on the chart is the one that we would obtain on an identical chart that is rotated by 180° around the centre. A similar chart does exist, even though it came out way after, and it is usually superimposed to the usual Smith chart (figure 3.7).

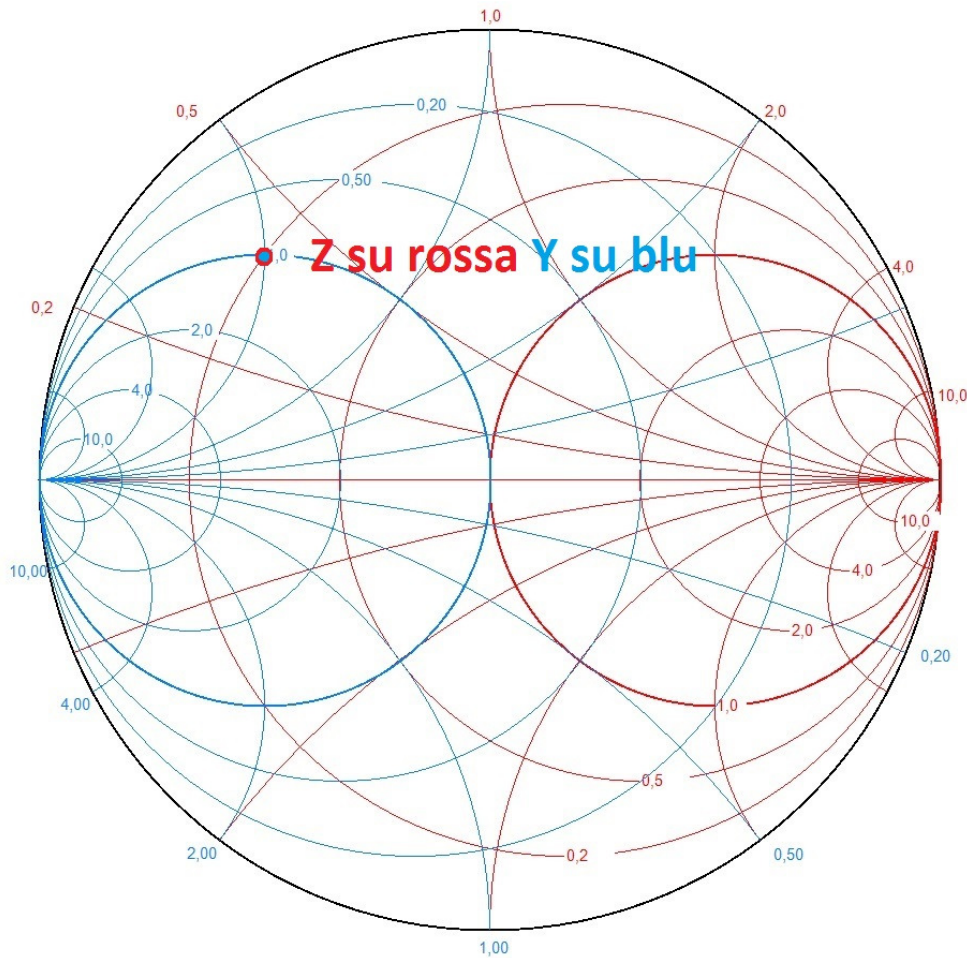


Figure 3.7: Smith Chart of the impedances (red) and the admittances (blue)

If the chart in figure 3.7 is used, there is no need to find the opposite point with respect to the center; it is enough to read the impedance **Z** on the Red chart (**Z su rossa**) and, remaining on the same point, read the value of **Y** on the Blue one (**Y su blu**).

You might wonder what the utility of this is. Every time we put a reactance **X** in series to our load we will move along the red circumferences (at constant resistance), while when we put it in parallel, we will move along the blue circumferences (at constant conductance). The objective is always to arrive at the center of the Smith chart, where we have the perfect match.

Lastly, I want to remind you that inside the lower half-circle we have the negative reactances (therefore capacitive) and inside the upper half-circle we have the positive reactances (therefore inductive), while on the horizontal diameter, that marks the border between the two half-circles, we have pure resistances.

I am going to end the explanation of the Smith chart here, if you want to dig deeper in the topic you will probably be able to find a lot of tutorials and articles on the internet.

#### 4. The device VNA.

Like I said in the first paragraphs, the Vector Network Analyzer (VNA) is a device available with one, two or more ports and it can measure direct and indirect tension waves and it can detect the tensions transmitted on every port. This device is particularly apt to analyze the



behavior of known networks (meaning the ones we already know the frequency behavior of, like an antenna, a load, a quartz, a filter...) in the context of radio frequencies and microwaves, by obtaining the S-scatter parameter on the measurement plain of the various ports. The internal structure of a two-port VNA - the most common one among the OMs - is shown in figure 4.1 and it can be PC based – when the processing and visualization part is delegated to an external PC – or it can be an autonomous unit, equipped with everything – called stand-alone.

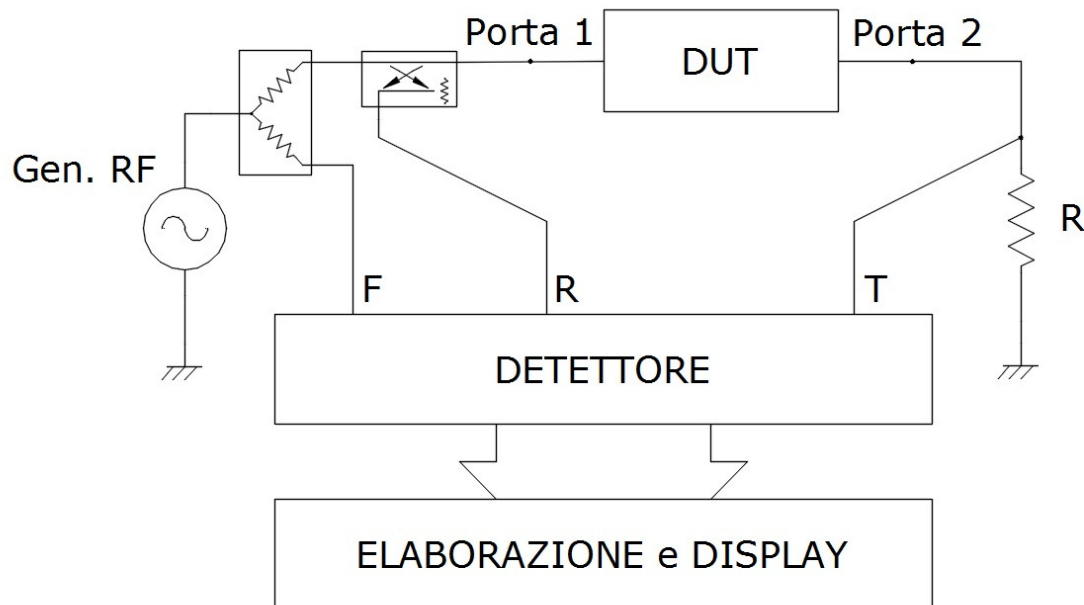


Figure 4.1: Minimal two-port VNA

As you can see from figure 4.1, inside there is a sinusoidal-tensions generator CW (usually a DDS) that produces a direct signal (F forward). The direct signal is made up by a succession of CW sinusoids at increasing frequency (called sweep), and it goes from an initial frequency – or start - to a final one – or stop – both adjustable. Before actually arriving at the connector located on port 1, the single tension waves CW go through a directional coupler that reads the corresponding reflected tension wave (R reflected) on the same port [4]. This is what happens inside the one-port VNAs, while the two-port devices have a downstream fictitious load at port 2, on which we can read the transmitted tension wave (T thru).

Said device is able to measure in module and phase only the parameters  $S_{11}$  e  $S_{21}$ . If we wanted to measure the parameters  $S_{22}$  e  $S_{12}$ , we would have to turn the DUT around. The most sophisticated devices have a system inside that allows for the automatic exchange of the ports, while others allow this type of exchange through another device called “test-set” (figure 4.2).

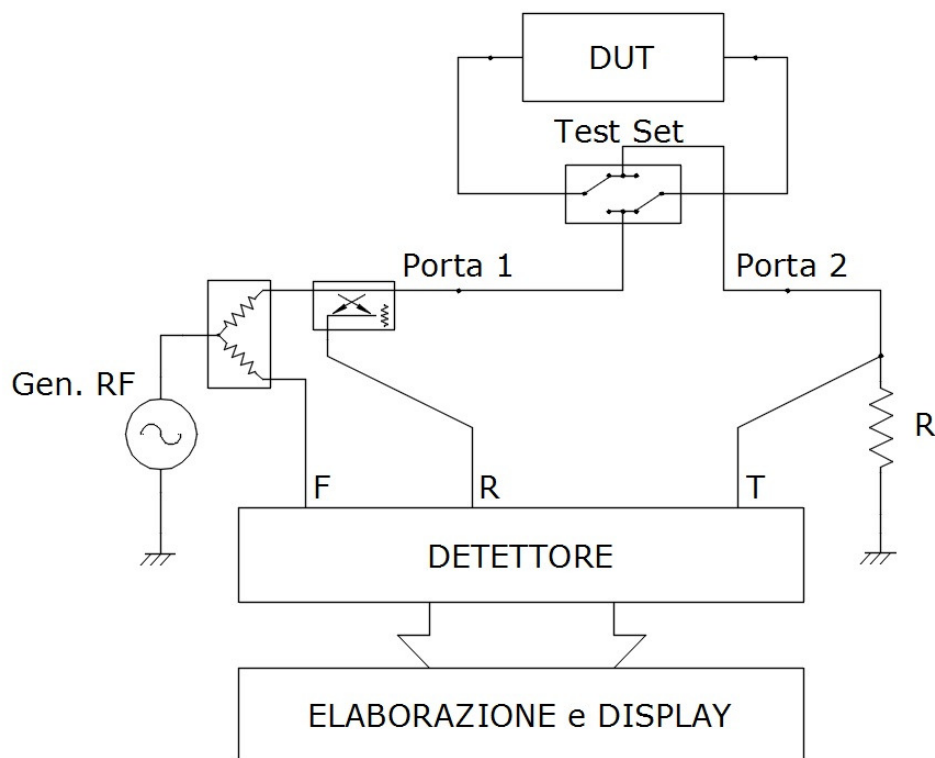


Figure 4.2: two-port VNA with external test-set

As you can imagine, the measurement of the S-scatter parameters is non-invasive and allows for very accurate measures. It wouldn't be possible to measure tensions or currents with the oscilloscope, because inserting the probe inside the circuit would be disrupting for the network. Disruption that transforms into reflections, which can distort the measure to the point of making it inaccurate, especially at high frequencies.

Even the VNA can suffer from undesired reflections, and for this reason the connections need to be checked almost maniacally, especially as the frequency increases. For example, the SMA connectors need to be tightened with the appropriate torque wrench to have perfect connections. In the field of MF and HF usually the directional coupler is not used - because it doesn't work well at low frequencies (MF) – and it is often substituted with a RF bridge that works well starting from continuous current but, compared to the coupler, it is deafer [4] and it is not accurate when measuring impedances that deviate from the nominal value (50 Ohm). The measure of the impedance remains very accurate anyway when it stays within values that are four times bigger and four times smaller (that is between 12.5 and 200 Ohm) and it is still acceptable between 5 and 500 Ohm (10 times bigger or smaller). It follows that all the derived parameters (capacity, inductance, etc...) won't be as precise outside the indicated parameters.

Additionally, the extra care for connections for the devices that work up to the HF limits is less important but is highly suggested. It goes without saying that from the VHF and beyond the meticulous attention is mandatory. The transmission waves F, R and T - that are non-invasively collected – are then measured by appropriate detectors and transformed into amplitudes and phases. Then, they get elaborated and the result of that gets graphically represented on a display, frequency by frequency. The graph can be a Smith chart or a cartesian or logarithmic graph; in any case it is possible to insert one or more markers that will show the values of the parameters on the desired point.

The instrument, in addition to the setting of a maximum and minimum frequency (interval called span), needs some other important basic settings, such as: scanning time of the span (sweep) and calibration.

The scanning time  $t_w$  depends on two factors: the number of measure points  $N$ , in which the span is divided (called data points) and the assessment time setting of the measure  $t_s$  (settling time). The result of the product between the points of measure number  $N$  and the assessment time  $t_s$  determines the time that's needed to carry out the whole scanning  $t_w$ .

$$t_w = N \cdot t_s \quad [\text{seconds}]$$

4.1

The number of points  $N$  in which the span gets divided determines the minimum interval of frequency  $\Delta f$  discriminable:

$$\Delta f = \frac{f_{stop} - f_{start}}{N} \quad [\text{Hz}]$$

4.2

The settling time  $t_s$  is important because it allows the direct tension wave to propagate up to the load and come back to the port where it is measured as reflected. In particular, if the interposed transmission line is long, some wait is needed for the R wave to settle and be able to make various round trips; the longer the waiting time, the more the measure will be accurate. At the same time, waiting for too long would be pointless, mainly because the EM wave travels, in the worst conditions, at 180,000 km/s (about 2/3 of light speed).

To the calibration I dedicated the entirety of the next chapter, since it is a crucial step to carry out errorless and repeatable measures. We'll see that calibration will have to be executed every time the measure setup or parameter setup (frequency interval, number of points, settling time, etc...) is changed. The device signals when it detects that it is not being calibrated, but if we change anything in the external measure setup (connectors, cables, etc...) it won't notice, and it will be on us to recalibrate.

## 5. VNA calibration

As already mentioned, to carry out accurate measures, calibration is an indispensable operation. I want to recall that accuracy is the ability of an instrument to get close to the real value, which, of course, is unknown.

The VNA is not able to understand what's going on externally, but it limits itself to compare the incident tension wave with the reflected and with the transmitted one, for every point of measure; then it calculates the  $S_{11}$  or the  $S_{21}$ , without knowing what is causing them; for this reason, calibration is necessary to make the device acknowledge the external situation. Calibration is done by connecting known loads to the plane of measure, upon request of the VNA, which allows to fix the reference points for subsequent measures. The loads to use for this process are the ones that have a known scatter parameter  $S_{11}$  (or, if you prefer, the reflection coefficient  $\Gamma$ ; see chapter 2, definition 2.3) or, short circuit  $S_{11} = -1 + j0$ , open circuit  $S_{11} = 1 + j0$  and dummy load  $S_{11} = 0 + j0$  (Load:  $Z_0 = 50 + j0$ ). To calibrate the  $S_{21}$  we use a bullet (Thru); this is why the calibration kits are called SOLT, from the initials of the words Short, Open, Load and Thru (if it was the case of a one-port VNA, a SOL kit would be enough). With three known loads it is possible to find the necessary correction parameters for the  $S_{11}$  measures and still have a good accuracy up to 6GHz. Beyond 6GHz the error matrix has to include more parameters, and this is why the price of SOLT kits is this high; it goes without saying that even the VNA has to be able to consider them, otherwise knowing these pieces of information would be completely pointless.

Naturally, the calibration process has to be carried out for the  $S_{22}$  and  $S_{12}$  parameters as well, since even with these there is an error matrix to calculate.

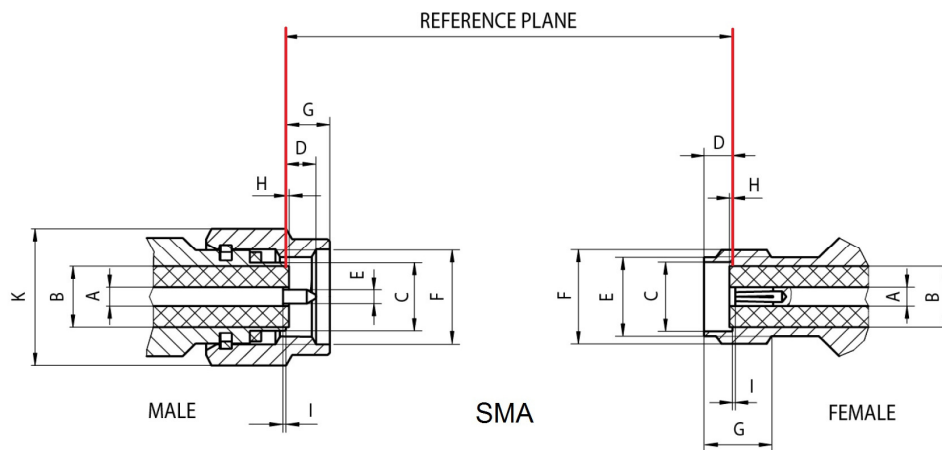
Notice that the three loads correspond to the points A, B and C (figure 3.2), all located on the real axis ( $j0$ ) of the Smith chart.

SOLT kits exist on the market for every type of connector (male and female), but the most affordable ones for the OM pockets are the ones made with SMA connectors.

Once the VNA is calibrated, we can build other SOLT kits, but the first one must be bought, since the calibration data to insert into the SW of the VNA is in the certificate. Indeed, when we make an open, we'll inevitably introduce a minimal capacity; even when we put it in short there will be a minimal parasite inductance. Those will become noticeable at very high

frequencies. In HF and lower, the accuracy is quite good even without considering the propagation delays, but it is equally important to calibrate the device and solve the error matrix. In VHF and up to 6GHz, the model that includes the delay and the first parasite coefficient allows for decent accuracy, however, as already mentioned, beyond 6GHz the models have to be way more precise, with the insertion of a 3<sup>rd</sup> degree polynomial model for which it is necessary to know many more corrective coefficients. A SOLT calibration kit for frequencies beyond 6 GHz with N connector can easily cost more than 6000€, while the one up to 6GHz with an SMA connector will cost less than 100€, shipping included. After this premise, I will now delve into the topic of measurement planes, which have already been introduced in the first chapter.

All the connectors have a conventional plane of reference (red line); in figure 5.1 I highlighted the conventional reference plane of an SMA connector. In the case of a Male, it coincides with the point where the Female crown leans against.



	Male		Female	
	min.	max.	min.	max.
A	Ø 1.245	Ø 1.295	Ø 1.245	Ø 1.295 <sup>1)</sup>
B	–	Ø 4.178	–	Ø 4.178
C	–	Ø 4.59	Ø 4.60	Ø 4.67
D	–	2.54	1.88	1.98
E	Ø 0.902	Ø 0.940	Ø 5.28	Ø 5.49
F	1/4-36 UNS-2B		1/4-36 UNS-2A	
G	–	3.43	4.32	–
H	-0.18	+0.05	-0.18	+0.05
I	0.00	–	0.00	0.41
K	hex 8		–	–

Dimensions in mm  
<sup>1)</sup> Contact diameter refers to 50 Ω

Figure 5.1: Reference plane in the Rosemberg SMA connectors taken from [5]

The conventional reference plane (CRP) is not, however, the point where the reflection phenomena of the waves happen; in fact, they take place where we decided to place the short circuit or the open. The plane where the reflection phenomena happen takes the name of physical reflection plane (green line) or PRP and it rarely coincides with the conventional reference plane; figure 5.2.

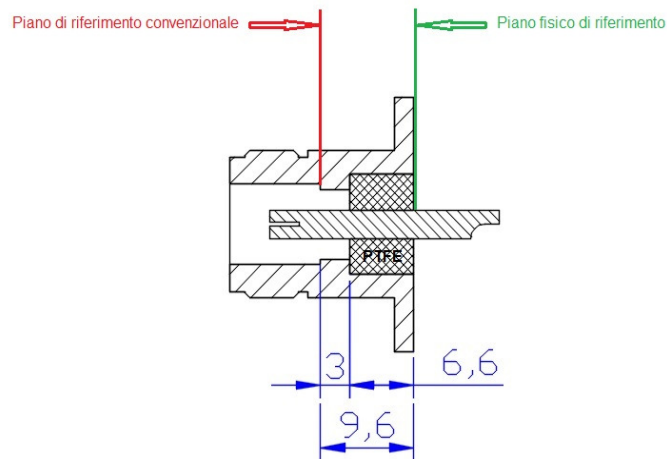


Figure 5.2: Reference and reflection planes of a female N connector

On the technical sheet of the SOLT kit, the distance covered by the waves and the values of the parasite parameters are indicated for people to be able to insert them in the VNA; that's how the SW will be able to correct the difference and bring back the physical plane (Piano fisico di riferimento) to coincide with the conventional reference one (Piano di riferimento convenzionale). The distances usually are expressed through the necessary time for the wave to travel the connector section/tract (or more generally, the transmission line) that divides the two planes. The calculation of the travel time is derived from the law of motion:

$$time = space/speed \quad 5.1$$

The speed of the wave is the one of light  $c$  (celeritas) multiplied by the speed factor VF, which is specific of the material where the wave goes through:

$$v_f = VF \cdot c \quad [\text{m/s}] \quad 5.2$$

In turn, the speed factor VF depends on the relative dielectric permittivity  $\epsilon_r$  (epsilon with r) of the insulator used according to the formula 5.3 (where  $\mu_r \approx 1$  for the insulators).

$$VF = \frac{1}{\sqrt{\epsilon_r \mu_r}} = \frac{1}{\sqrt{\epsilon_r}} \quad 5.3$$

The most used parameter, the inverse of the speed, is expressed in ps/mm (Pico seconds= $10^{-12}$ s) and it is the time it takes to the wave to travel one millimeter of line.

$$\frac{1}{v_f} = \frac{1}{VF \cdot c} = \frac{\sqrt{\epsilon_r}}{c} = \frac{\sqrt{\epsilon_r}}{0,3} \quad [\text{ps/mm}] \quad 5.4$$

Materiale	$\epsilon_r$	VF	ps/mm
Aria	1,0	1,00	3,333
Foam	1,6	0,79	4,216
Teflon	2,1	0,69	4,830
Polyetilene	2,3	0,66	5,055
PVC	3,5	0,53	6,236

Table 1

Therefore, summing up, the physical distance  $d_f$  (the one measured with a meter) is the distance that the wave travels on the real line:

$$d_f = d_e \cdot VF \quad 5.5$$

The electric distance  $d_e$  is the one the wave would travel if it was in the vacuum, and can be obtained from formula 5.5 in the following way:

$$d_e = \frac{d_f}{VF} \quad 5.6$$

The time  $t_d$  it takes the wave to travel  $d_e$  becomes:

$$t_d = d_e \cdot \frac{1}{c} = d_f \cdot \frac{1}{VF \cdot c} \quad 5.7$$

In the case of figure 5.2, for the wave to go from the physical reference plane (green line) to the conventional reference plane (red line) it takes:

$$t_d = t_{air} + t_{PTFE} = \frac{d_{air}}{v_{air}} + \frac{d_{PTFE}}{v_{PTFE}} = 3 \cdot 3,333 + 6,6 \cdot 4,830 = 41,88 \text{ ps} \quad 5.8$$

Since the incident wave encounters the conventional reference plane first and then the physical one, a time  $t_f$  will have to be subtracted and therefore taken with a minus sign  $t_f = -41,88 \text{ ps}$ . Lastly, the incident wave travels once through the connector tract between the conventional reference plane and the physical one, when it arrives on the physical plane, and a second time when the incident wave continues as a reflected wave (two way); on the other hand the wave goes through it only once when the incident wave continues as a transmitted wave (one way). In other words, when one calibrates to obtain the parameter  $\mathbf{S}_{11}$ , doubling the traveling time is needed, while when calibrating the  $\mathbf{S}_{21}$  doubling is not needed. It seems complicated, but the VNAs know this and do not get confused.

Once the  $\mathbf{S}_{11}$  parameter is calculated one can move to the calibration of the  $\mathbf{S}_{21}$  and then, to the parameters  $\mathbf{S}_{22}$  and  $\mathbf{S}_{12}$ . Of course, if it's the case of a one-port VNA, the  $\mathbf{S}_{11}$  is enough. Inserting the correction factors is part of every VNA and it strictly depends on the SW used. For the purpose of this essay, we will reference DG8SAQ's device, a VNA version 3, and in particular to the texts [6] and [7].

Everything I said until now regarding the calibration might sound excessive, but it is actually the bare minimum for the VNA to be able to measure the phase of the scatter parameters, together with their module. This is because this device is very powerful but also quite hard to use. As a matter of fact, if we calculate how much a degree of physical wavelength  $l_{1^\circ}$  ( $\lambda_f = VF \cdot \lambda$ ) corresponds to, we will notice right away of how much the lengths are important (if  $f$  is in MHz, then  $c=300,000 \text{ mm/s}$ ).

$$l_{1^\circ} = \frac{\lambda_f}{360^\circ} = \frac{VF \cdot \lambda}{360^\circ} = VF \cdot \frac{c}{360^\circ \cdot f} \quad [\text{mm}/^\circ] \quad 5.9$$

Let's consider the limits of the HF (30MHz), VHF (300MHz) and UHF (3GHz) bands from 5.9, in air; we'll obtain that a degree of wave  $l_{1^\circ}$  corresponds respectively to: in HF, 27.7 mm; in VHF, 2.77mm and in UHF, 0.27mm. With lower speed factors VF (like with coaxial lines) the lengths decrease further. Therefore, up to the HF limits there are no big problems, in the VHF even one mm is important and, in the UHF, and superior, the SOLT model we have presented until now has to be used following the technical sheet and inserting all the data. In figure 5.3, you can see the components of a SOLT kit SMA-f from Rosemberger for calibration (the SMA calibration ones that are economically accessible to us OMs are from Rosemberger or Amphenol), which are: a short circuit f, a bullet f-f and a fictitious load of  $50\Omega$  f.

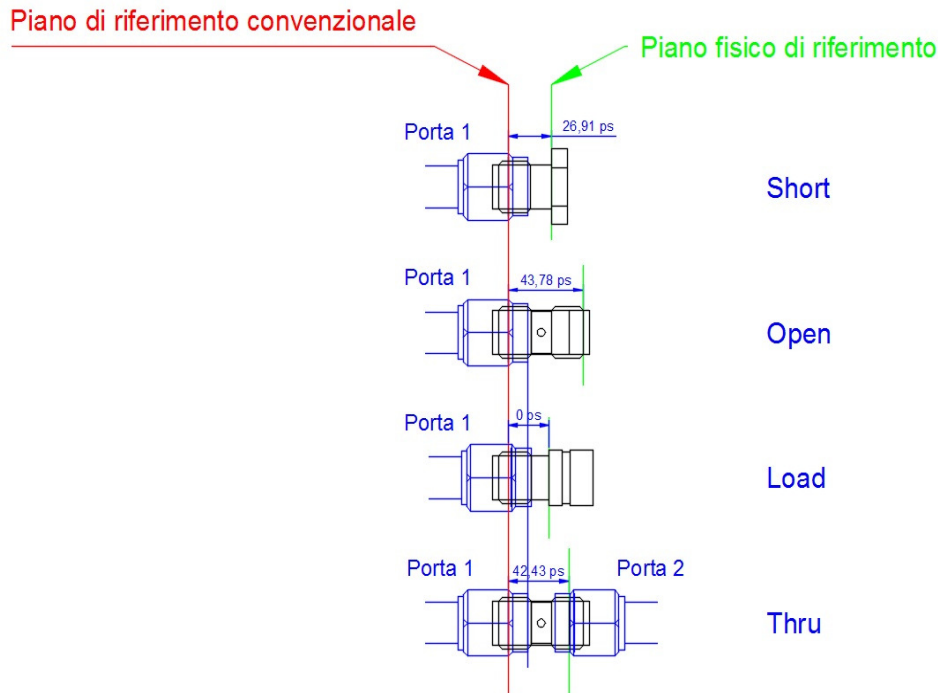


Figure 5.3: Female Rosemberger SMA calibration kit

The kit has to be used exactly as described in the figure above, otherwise the time frame that brings the physical plane of reference (green) to coincide with the conventional reference one (red) won't be correct.

It is a good practice to tighten the connectors with the appropriate dynamometric wrench, to guarantee a constant tightening torque of the connector (I built one myself and I set a tightening torque of 0.8 Nm). Remember that when the parameter  $S_{11}$  (or  $S_{21}$  at port 2) will be calibrated with the Short, the Open and the Load, the time frames will have to be doubled and will have negative sign compared to the ones in figure 5.3.

In the calculation of the Open time frame the effect due to the parasite capacity has already been considered, by adding an additional delay (in my case 1.35 ps have been added to the delay that the bullet would have had if it hadn't been left open on one side).

In the Load, it is necessary to measure with a 4-wire precision ohmmeter how much the resistance located in the fictitious load measures – a couple of decimals are sufficient. Furthermore, we need to indicate how much the parasite capacity put in parallel to the resistance in the fictitious load measures; usually it is indicated in the data that come with the kit. Said capacity is expressed in fF (femto Farad, or  $10^{-15}F$ ). In my case the resistance value is  $49.26 \Omega$ , the parallel capacity is 2fF and the delay is 0 pico seconds.

The Load delay existing between the conventional reference plane and the physical plane is often set to zero because it is considered as ideal (delay=0, loss=0 and  $Z_0=50$ ), therefore there is no reflected wave  $S_{11}=0$ . For this reason, the eventual minuscule reflected wave due to the non-ideal Load will be compensated by making it null, which will make the delay insignificant. This means that the accuracy of the device will never be superior to the one of the calibration kit.

Lastly, in the Thru, the physical plane is located halfway through the bullet, so it will be sufficient to indicate the distance between the two planes of the bullet in ps, since the SW of the VNWA will cut it by half on each side automatically.

The calibration of the Thru has to necessarily be preceded by the SOL calibration, since by knowing solely the delay introduced by the bullet, it won't be possible to zero out the line segment (of random length) that connects the bullet to port 2 with the detector at internal 50 Ohm.



The calibration of the VNWA also includes the zeroing of eventual crosstalk that might happen on the reception line connected to port 2.

Having done so we move on to the zeroing of the line between the bullet and the detector (Thru) and then to the measure of the internal detector impedance (Thru Match), to increase the accuracy of the measure of the  $S_{21}$ .

At the end of the calibration, the device will be able to correctly measure anything that gets connected to the conventional reference plane or, in the case of a two-port measure, between the two conventional reference planes. The correct traces that are generated on the conventional reference plane when it is left open are the ones shown in figure 5.4.

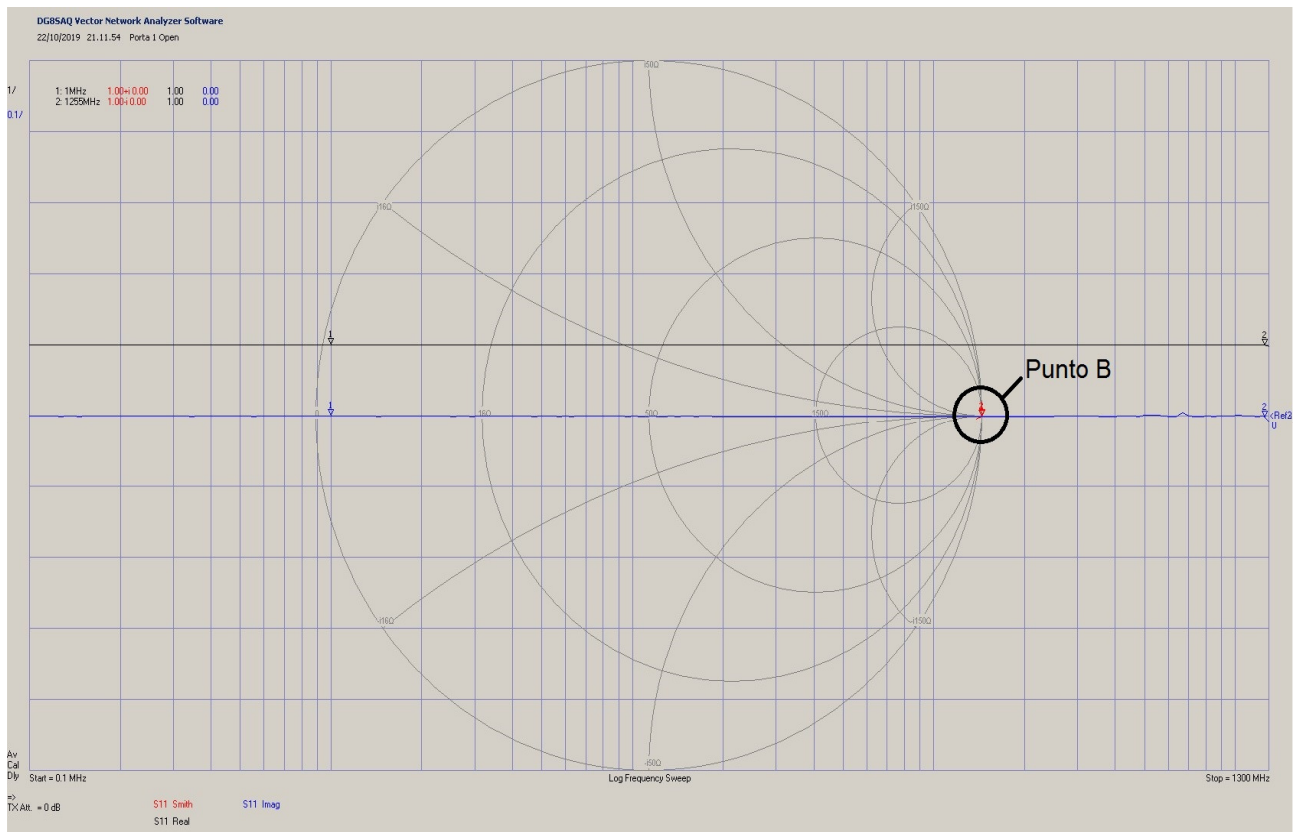


Figura 5.4: Open sul piano di riferimento convenzionale

Notice that the red trace is almost point-like around point B on the Smith chart (real impedance is infinite). The black trace represents the real part of the scatter parameter  $S_{11}$  while the blue one represents the imaginary part of  $S_{11}$ . Aside from small oscillations due to the small scale ( $0.1^\circ$ ), the traces are flat. This indicates that the parameters relative to the SOL kit in question are accurate and therefore the device is calibrated successfully.

To check for the calibration of the Thru, we need to connect port 1 to port 2 with the cables used for the calibration, interconnected with the same bullet, then we need to launch a scan  $S_{21}$ ; this will have to generate a trace  $|S_{21}|=0$  dB, flat everywhere on the measurement range. Later, when measuring with the VNA, we will see how to find the parameters of a homemade calibration kit, starting from the physical measures of the homemade components and then perfected with the ones measured with the calibrated VNA.

## 6. Measuring with the VNA

Once the device is calibrated correctly, we will be able to measure anything that is on the conventional reference plane, which, in figure 6.1, coincides with the measurement plane; in particular, we can see a  $100\Omega$  resistor insertion point on the measurement plane (orange).



Figure 6.1: R 100 Ohm on the conventional measurement plane

Let's now carry out a scan to measure the real part and the imaginary part of the impedance that is derived from the  $S_{11}$  parameter.

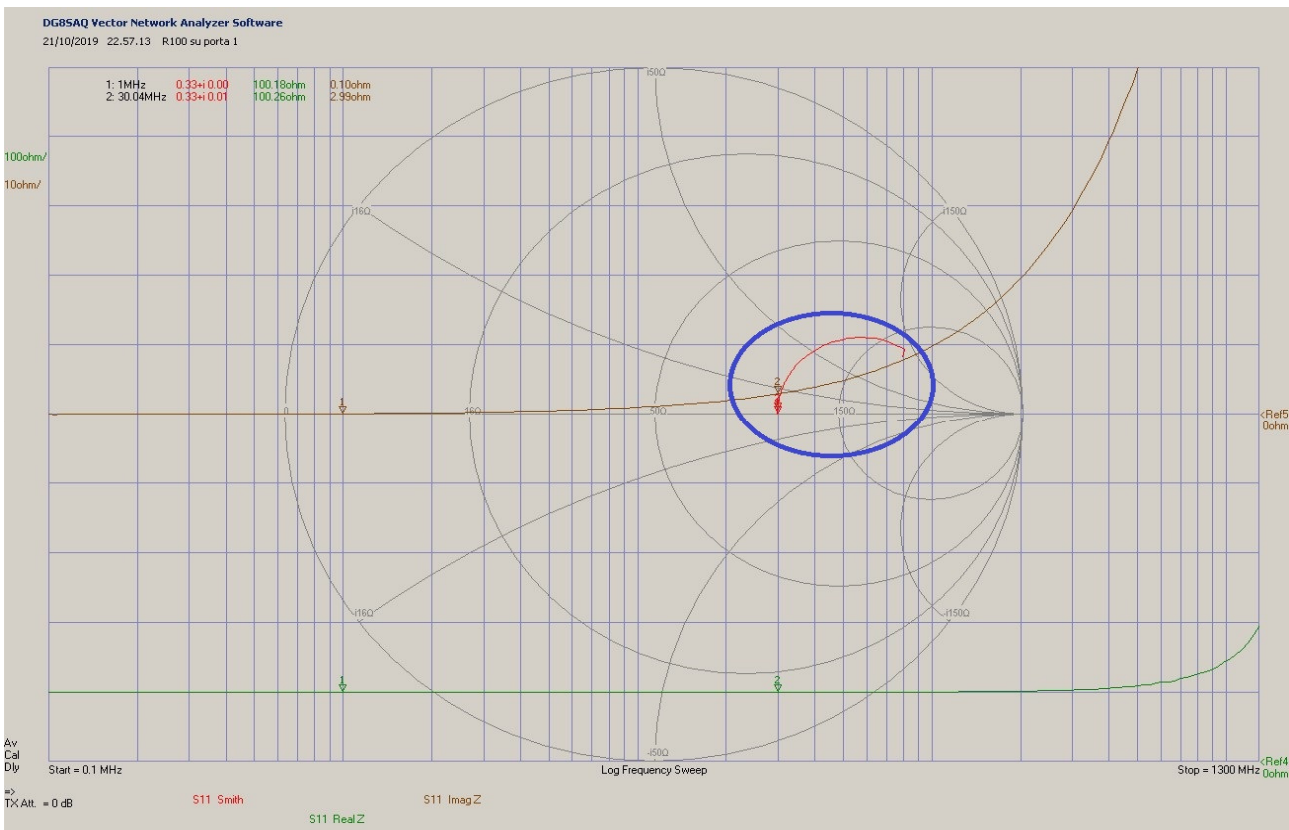


Figura 6.2: Resistore da 100 Ohm posto sul piano di riferimento convenzionale

The green trace in figure 6.2 represents the real part of the resistance, while the brown trace represents the imaginary part. Inside the blue circumference, on the Smith chart, the impedance  $Z$  is reported; it is calculated from the  $S_{11}$  parameter at the varying of the frequency. The red trace starts from a point on the real axis of the Smith chart, and it enters the upper semi-circle of the positive impedances at the growing of the frequency. This underlines the presence of a parasite inductance that gets more prominent as the frequency increases. Either way, we can consider the resistance under examination as acceptable until 100MHz. It is not convenient to carry out the measures directly on the conventional

reference plane; it's simpler, in fact, to have a connector onto which we weld our 100 $\Omega$  resistor. To do so we need to insert an SMA m-m adapter and a SMA f connector for PCB (figure 6.3).



Figure 6.3: SMA-F connector for PCB with SMA M-M adapter

By doing so, we have moved the measurement plane by 25.4mm from the conventional reference plane; let's then see what happens when we launch a scan.

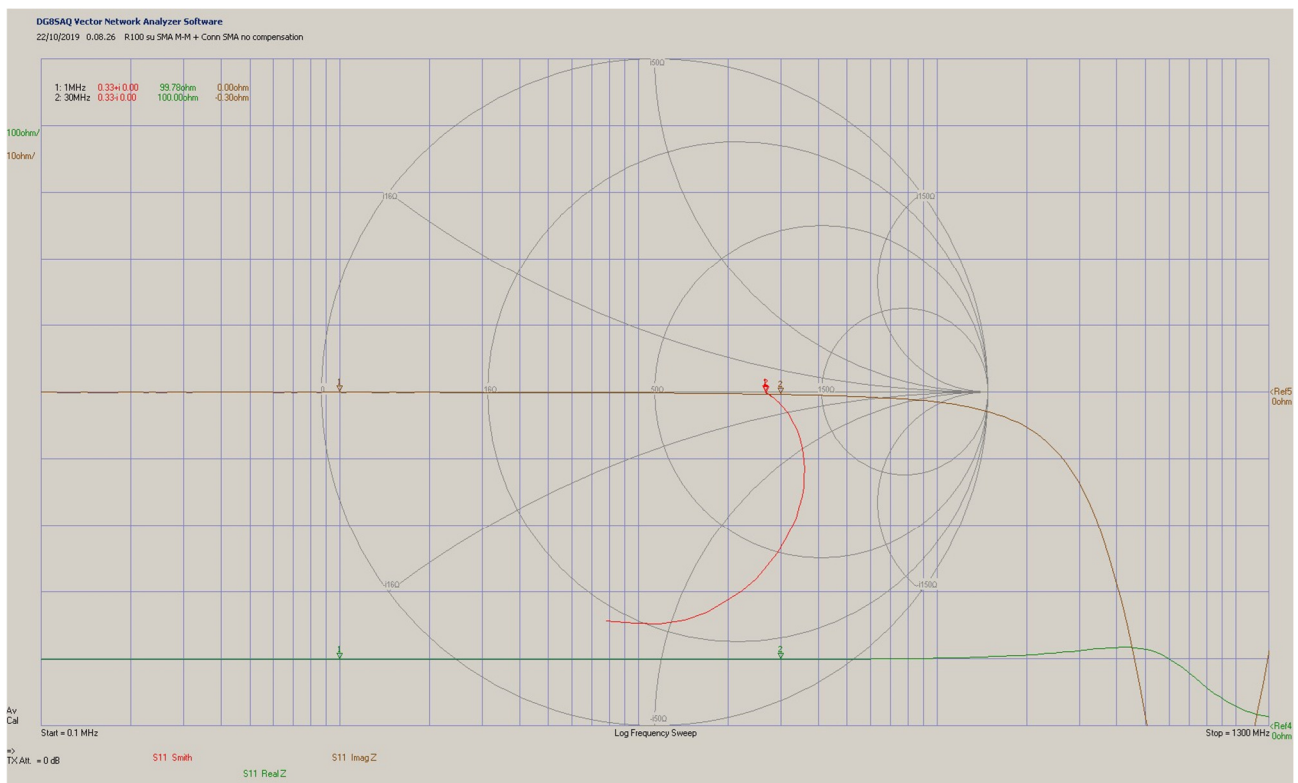


Figure 6.4: 100 Ohm resistor placed 25,4mm from the conventional reference plane

As you can see in figure 6.4, the trace is completely inside the lower semi-circle, therefore showing that the parasite effect seems to be capacitive. Nevertheless, our resistor isn't affected until 30MHz; at higher frequencies the effect of the parasite becomes noticeable (the brown trace represents the reactance).

The measure in figure 6.4 confirms what I stated in the previous paragraph, where we discussed how delays up to 27mm do not affect measures in HF; as a matter of fact, we introduced a line segment of about 25.4mm, constituted by a SMA m-m adapter and an SMA



f connector for PCB. The delay introduced is calculated with the equation 5.7, which becomes:

$$t_d = d_f \cdot \frac{1}{VF \cdot c} = 25,4 \cdot 4,830 = 122,7\text{ps} \quad 6.1$$

The delay we obtain is of about 122 pico seconds on the incident and reflected wave, delay which, as you can see from figure 6.1, heavily affects the measure starting from VHF. To “fix” the measure we need to compensate the delay by telling the VNA that the measurement plane moved away from the conventional reference one by 122.7ps; the + sign indicates that it moved away from the TX port.

The simplest method to compensate the delay in DG8SAQ’s VNA is to insert the delay in “Port Extension” under the menu bar “Measure”; this way the measure goes back to the original (figure 6.5). This method is quite practical, and it is convenient in every occasion in which the position of the measurement plane changes in a non-permanent way, meaning just for a few measures.

The most stable alternative to the above solution is recalibrating the VNA with the SOL kit on the new measurement plane and consequently making it so that the conventional reference plane repositions itself on the new measurement plane. In both cases the measure becomes the one in figure 6.5.

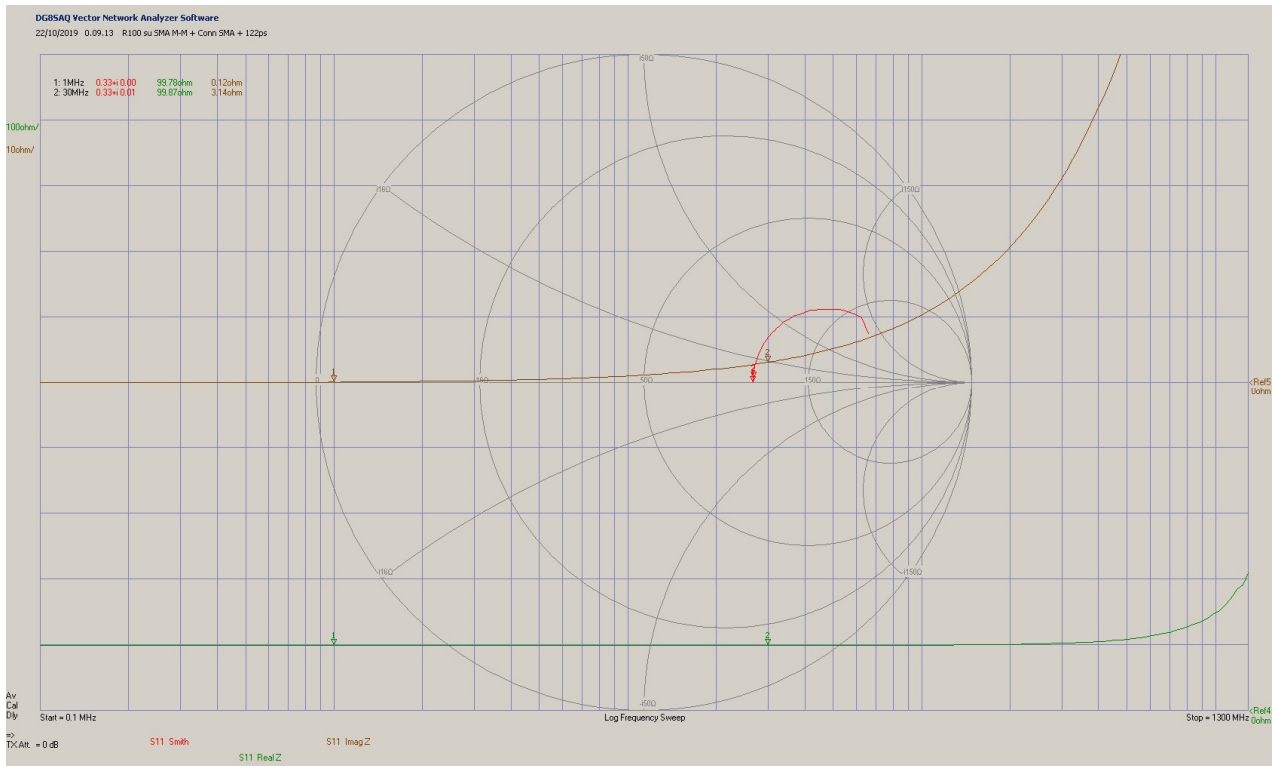


Figure 6.5: Delay compensation between the measurement and conventional reference plane

## 7. Misurare il delay di una linea di trasmissione

When measuring a device, it often gets connected at the end of a transmission line, therefore it is necessary to move the conventional reference plane onto the new measurement plane. This operation requires inserting the delay caused by the introduction of a section of line, in picoseconds (ps).

To obtain the delay of a transmission line you need to know its speed factor VF and the length of the segment that goes from a connector to the other, measured with a meter. As an example, I prepared a section of a SUCOFORM-141-50 deformable cable (figure 7.1), headed with two SMA-m connectors, its speed factor being VF=0.71 [8] and the length between the two connectors 170.2 mm (measured with a calliper). From the technical sheet

of the SMA connector of figure 5.1 it emerges that the distance from the connector border and the conventional reference plane is of 3.43 mm (quote G), therefore, to get to the two conventional reference planes we need to remove 6.86 mm. The length of the lines then becomes  $d_f = 163.3$  mm.

From equation 5.7 we can calculate the delay  $t_d$  the wave will have in the cable only:

$$t_d = d_f \cdot \frac{1}{VF_{PTFE} \cdot c} = 163,3 \cdot 4,695 = 766,7 \text{ ps} \quad 7.1$$

After this, I inserted a SMA f-f bullet in the connector left open, since the SMA male connector does not guarantee the 50  $\Omega$  impedance when it is left open, while the female one behaves better; then I added the typical delay of my SMA f-f bullet left open, which is 43.78 ps.

The total delay introduced by the section of cable + the bullet is:

$$t_d = 766,7 + 43,78 = 810,5 \text{ ps} \quad 7.2$$

Always sum the delays together because in the delay calculation the type of dielectric is already taken into account.

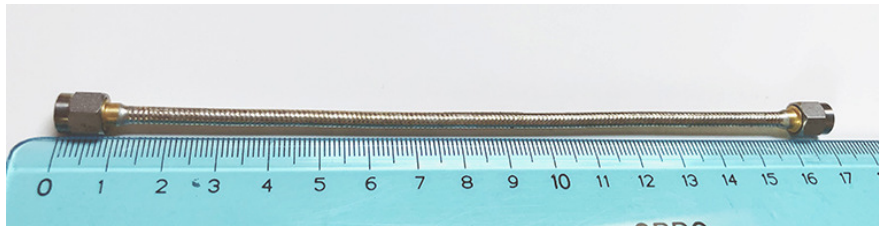


Figure 7.1: Sucoform 141-50 cable, headed with an SMA-m

Let's now connect the section of cable to port 1 and do a scan  $S_{11}$ , leaving the SMA connector on the other side open, with the bullet inserted. The result is the one in figure 7.2.

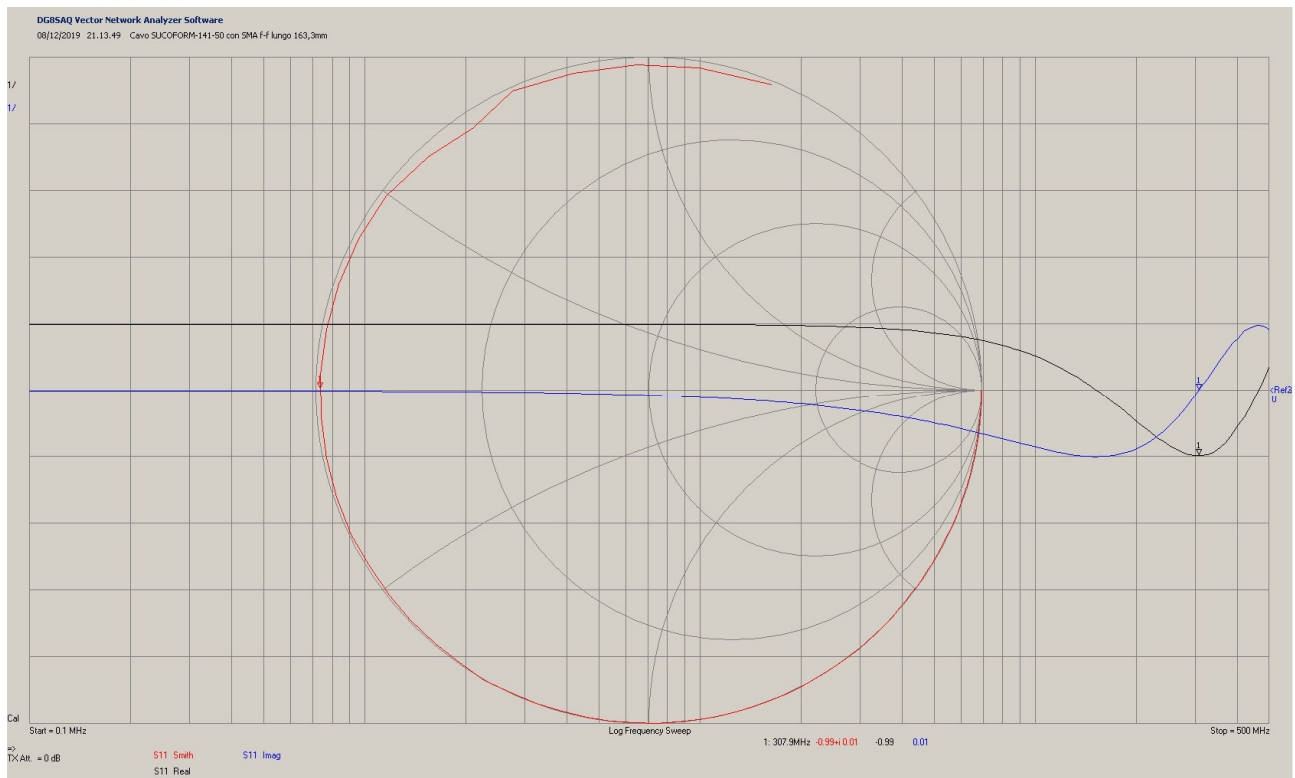


Figure 7.2: Segment of Sucoform 141-50 cable, open

As you can see by looking at figure 7.2, at low frequency the module of the  $S_{11}$  (black trace) is equal to 1, but the more the frequency grows closer to  $\lambda_e/4$ , the module of the  $S_{11}$  reaches -1; then it gradually goes back to the value of 1 when the frequency becomes  $\lambda_e/2$ . At the growing of the frequency, the trace of the module will keep swinging between -1 and 1 in

correspondence of multiples of  $\lambda_e/4$  and  $\lambda_e/2$  respectively. This is a sign that the conventional reference plane is not placed at the end of the coaxial line.

Now let's insert the value of the delay of figure 7.2 in the VNA extension port and see what happens.

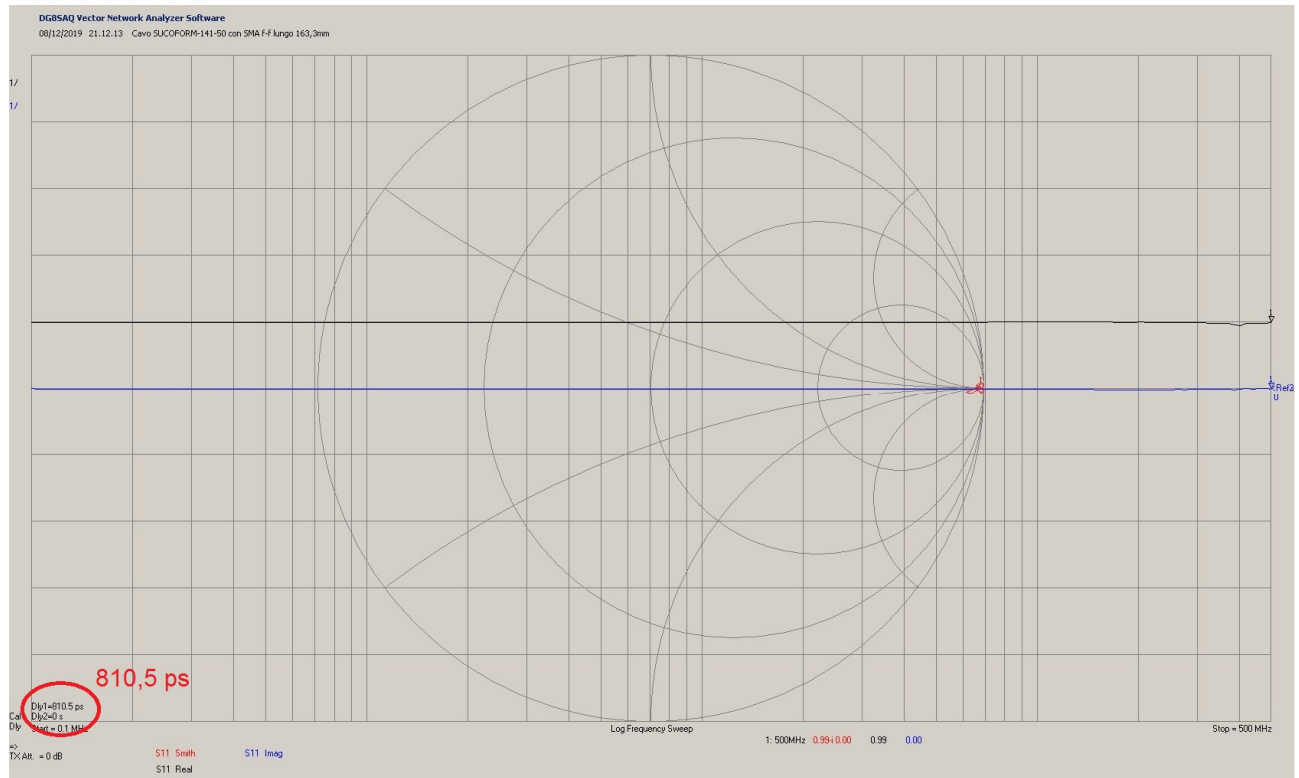


Figure 7.3: Sucoform 141-50 cable segment, open, with corrected delay

The correction adopted has made the imaginary part of the reflection coefficient flat (blue trace, figure 7.3). There is no difference between the value calculated and the one needed, since deriving the delay from the physical measure is always very accurate. The small irregularities are due to imperfections in the cable dielectric. As a matter of fact, the electric length of the coaxial (even if of good quality) depends on the frequency and varies according to the length due to manufacturing differences.

Ultimately, moving the measurement plane in the new position after adding a section of a cable is rather complex and does not give optimal results, albeit them being overall acceptable, especially for small movements (if compared to the wavelength). For big movements it is better to carry out a new calibration with the SOL kit on the new conventional reference plane to eliminate any error without much hassle.

This latter technique is the same technique that is used to take measures at the end of down cables of our antennas without necessarily having to learn how to fly, or how to position an antenna analyser (aka one port VNA) on top of a post and observe it with a drone.

## 8. Measuring a homemade SOL kit.

Building a SOL kit connector N-f requires first, to get N-f connectors of good quality, such as ones with a four-hole flange; then, we need to measure the physical distances with a good caliper or find the quotas on the data sheet provided by the manufacturer. In my case, I used three connectors (taken by [9]), from which I cut the rear pins (figure 8.1).



Figure 8.1: N-f, 4-hole connector

The quotas detected with the caliper are reported in figure 8.2.

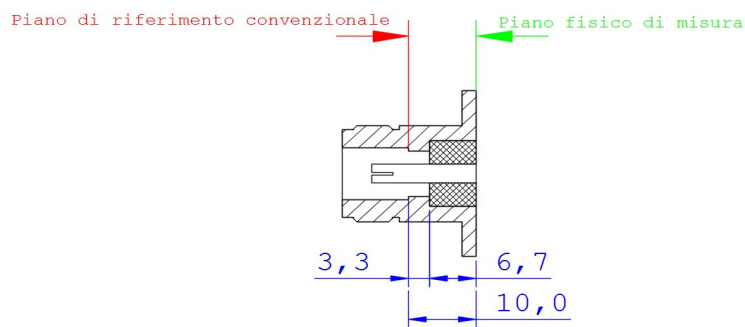


Figure 8.2: Quotas of N-f connector with 4-hole flange

From the equation 5.7, as seen before, we obtain the delay  $t_N$  of our connector N:

$$t_N = t_{air} + t_{PTFE} = \frac{d_{air}}{v_{air}} + \frac{d_{PTFE}}{v_{PTFE}} = 3,3 \cdot 3,333 + 6,7 \cdot 4,830 = 43,36 \text{ ps} \quad 8.1$$

Now let's get a thin copper square platelet (10x10 mm, some tenth of millimetres thick, like the coaxial screen RT5020), and in the middle, let's poke a hole with a diameter of 3mm; we also need two 100  $\Omega$  SMD 0805 resistances with a 0.1% precision thin film (RS 666-2437). With a powerful welder, let's weld everything as shown in figure 8.3.



Figure 8.3: SOL kit with N-f

We have obtained a N-f SOL kit; now let's see how to characterise it.

Let's calibrate our VNA with the SOL kit we bought (or borrow it from someone that has it calibrated), the closest to port 1, putting the minimum frequency equal to 100KHz and the maximum one to 500KHz (hypothesised range of use). Then, let's set 200 points of measure and keep the settling time at  $t_s=100$  ms.





Figure 8.4: Measurement setup

To allow for the homemade SOL kit to be screwed on the SMA f of port 1, it is necessary to insert an adapter from SMA-m to N-m. The measurement setup prepared uses an adapter from Radiall that you can see in figure 8.4. The two red traces indicate the old and new conventional reference planes. Obviously, we will need to tell the VNA how many picoseconds the new plane has moved. To do so, let's measure the distance between the two red lines of figure 8.4 in mm with the calliper; while doing so we have to distinguish between the one in air and the one in the PTFE. With the 5.7, let's transform the physical distances  $d_i$  (the ones measured with the calliper) in the respective delays, hypothesising a  $VF = 0.69$  in the PTFE. Then, let's sum the delays (like we've done in the 5.8) to find the delay  $t_A$  of the entire adapter:

$$t_A = t_{air} + t_{PTFE} = \frac{d_{air}}{v_{air}} + \frac{d_{PTFE}}{v_{PTFE}} = 11,9 \cdot 3,333 + 16,1 \cdot 4,830 = 117,42 \text{ ps} \quad 8.2$$

Then we have to screw the homemade Short into the adapter; we're starting from the short since it is not affected from the parasite capacity typical of the Open. We expect having to add another 43.37 ps (calculated with 8.1) to the delay of the adapter calculated with the 8.2. This is due to the fact that the four-hole connector with short circuit at the end, introduces a further delay; for this reason, the total delay expected on port 1, between the calibration plane and the physical one will be:

$$t_T = t_A + t_N = 117,42 + 43,36 = 160,78 \text{ ps} \quad 8.3$$

Let's insert the delay calculated with the 8.3 into the extension port 1 and let's launch the scan. The result is represented in figure 8.5.

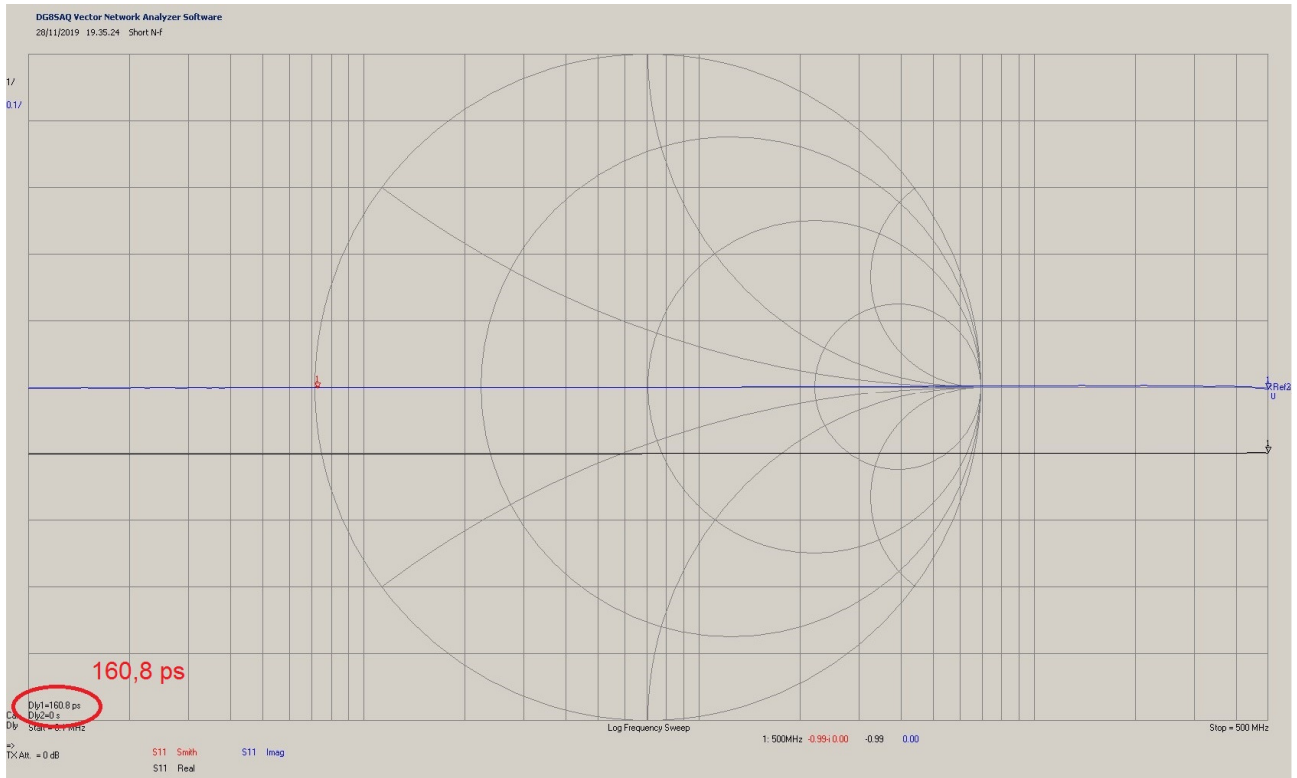


Figure 8.5: N Short without movement of the conventional reference plane

To obtain a flat/horizontal trace we had to correct the delay taking it to  $t_{sm}=160.8$  ps. The scale of the imaginary part (blue trace) is set at 0.1, while the real part (black trace) one is set at 1. The difference between the delay of the Short measured and the one calculated  $t_{ds}$  is very small:

$$t_{ds} = t_{sm} - t_T = 160,80 - 160,78 = 0,02 \text{ ps} \quad 8.4$$

This is probably due to the really small parasite inductance.

However, the theoretical result and the measured one are almost coincidental, and this confirms that the value calculated from the physical measures is extremely accurate.

Let's take the theoretical delay as a delay of the Short, plus the 0.02 ps measured:

$$t_s = 43,37 + 0,02 = 43,39 \text{ ps} \quad 8.5$$

Let's deal with the Open in the same way.

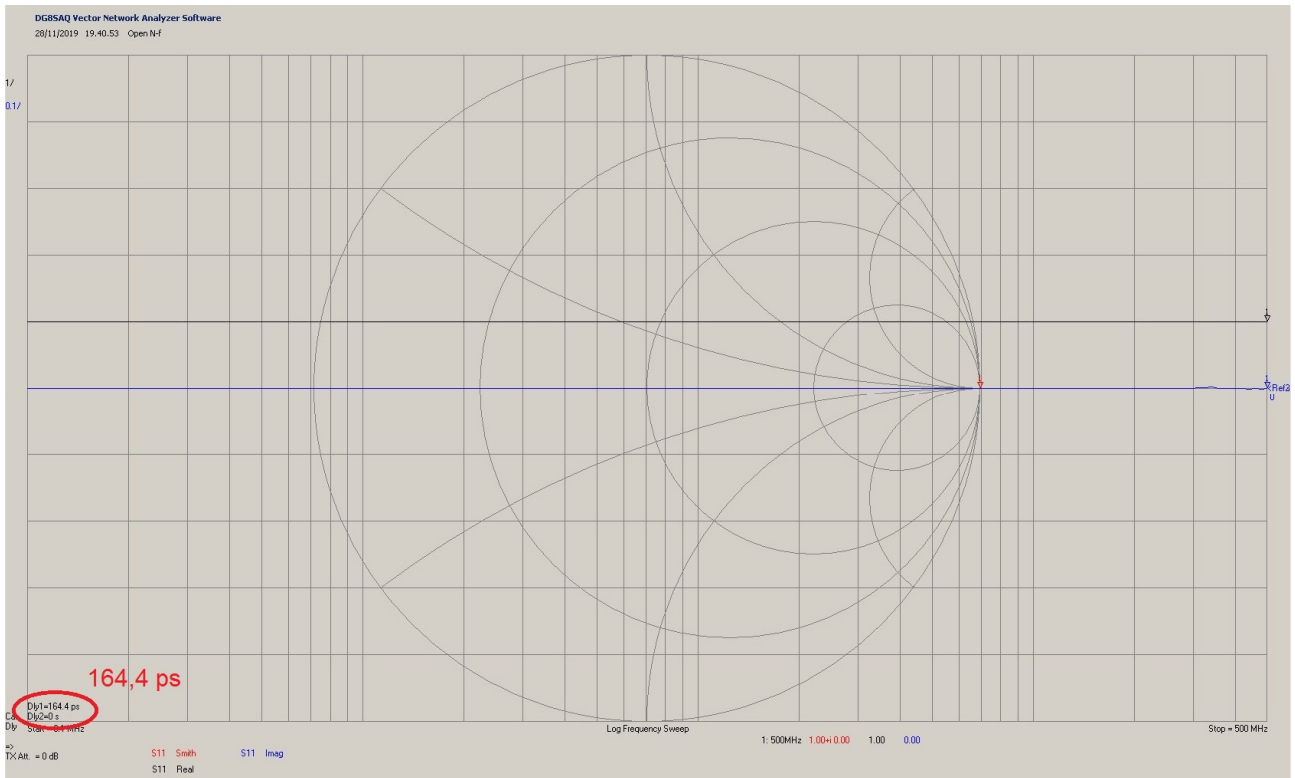


Figure 8.6: Homemade OPEN: delay measured after calibration with itself

In this case, the delay necessary to flatten the blue line is  $t_{Om}=164.4$ . For this reason, the difference between the delay of the Open and the  $t_T$ , calculated with the 8.3 is:

$$t_{dO} = t_{Om} - t_T = 164,4 - 160,78 = 3,62 \text{ ps} \quad 8.6$$

Therefore, the total delay of our Open becomes:

$$t_O = t_N + t_{dO} = 43,36 + 3,62 = 46,98 \text{ ps} \quad 8.7$$

We added the difference of 3.62 ps because it is attributed to the parasite capacity of the Open. Now, with a 4-wire multimeter (I used an HP 34401A) let's measure the resistance of our Load which, in my case, presents an  $R=50.006 \text{ Ohm}$ .

With the data above, we can dare to hypothesise the parameters for our homemade SOL kit. As a matter of fact, with the delays  $t_S$  and  $t_O$ , we know the distance between the physical plane and the conventional reference plane (that will also be the measurement plane). Additionally, we know that the physical plane is located further away from port 1 and that the tension wave travels through its path twice (two way) when the  $S_{11}$  calibration is carried out. Therefore, the parameters that characterise our SOL kit are the following:

$$SHORTdelaytwo\text{way} = -2 \cdot t_S = -2 \cdot 43,39 = -86,78 \text{ ps} \quad 8.8a$$

$$OPENdelaytwo\text{way} = -2 \cdot t_O = -2 \cdot 46,98 = -93,96 \text{ ps} \quad 8.8b$$

$$LOADresistance = 50,006 \text{ Ohm} \quad 8.8c$$

For now, let's ignore the parallel capacity  $C_{//}$  of the Load.

Once we put the values in the VNA, let's carry out a calibration with the SOL kit that we've just characterised, so that the measuring plane moves stably on the adapter crown of figure 8.4 (red circumference).

Be careful to reset the extension port that we inserted beforehand, otherwise the calibration won't be accurate.

Having done so, let's screw the SOL kit Short (the one we've just used for the calibration) onto the adapter and do an  $S_{11}$  measure to see how much delay we have to insert to move the conventional reference plane from the adapter crown (conventional reference plane) to

the physical plane of the Short. Said delay, if we did things properly, will have to be equal to the  $t_0$  of 8.7.

The trace of the imaginary part of the  $S_{11}$  becomes flat with extension port= 46.98 ps; just perfect.

Let's do the same thing with the Short too, where I found that the extension port= 43.39 ps. If the values are too different or if they start to fluctuate at high frequencies, it means we didn't calibrate the VNA or the delays correctly. Don't get too hung up on it, though, since the measure depends on many factors such as the temperature and/or the coupling torque of the connectors; moreover, remember that you won't be able to be more accurate than the calibration kit with which you carried out the first calibration.

The measure of the Load, instead, is more problematic. It requires inserting through attempts the parallel capacity  $C_{//}$  and fine tuning the value multiple times before obtaining the one that allows to get the flat imaginary line of the  $S_{11}$  (blue trace of figure 8.7). With the version 36.5.0 of the VNWA it is possible to use the function OPTIMIZE, that allows to find the values of the parameters automatically (in a recursive way); our homemade Load has  $C_{//}=2.7$  fF.

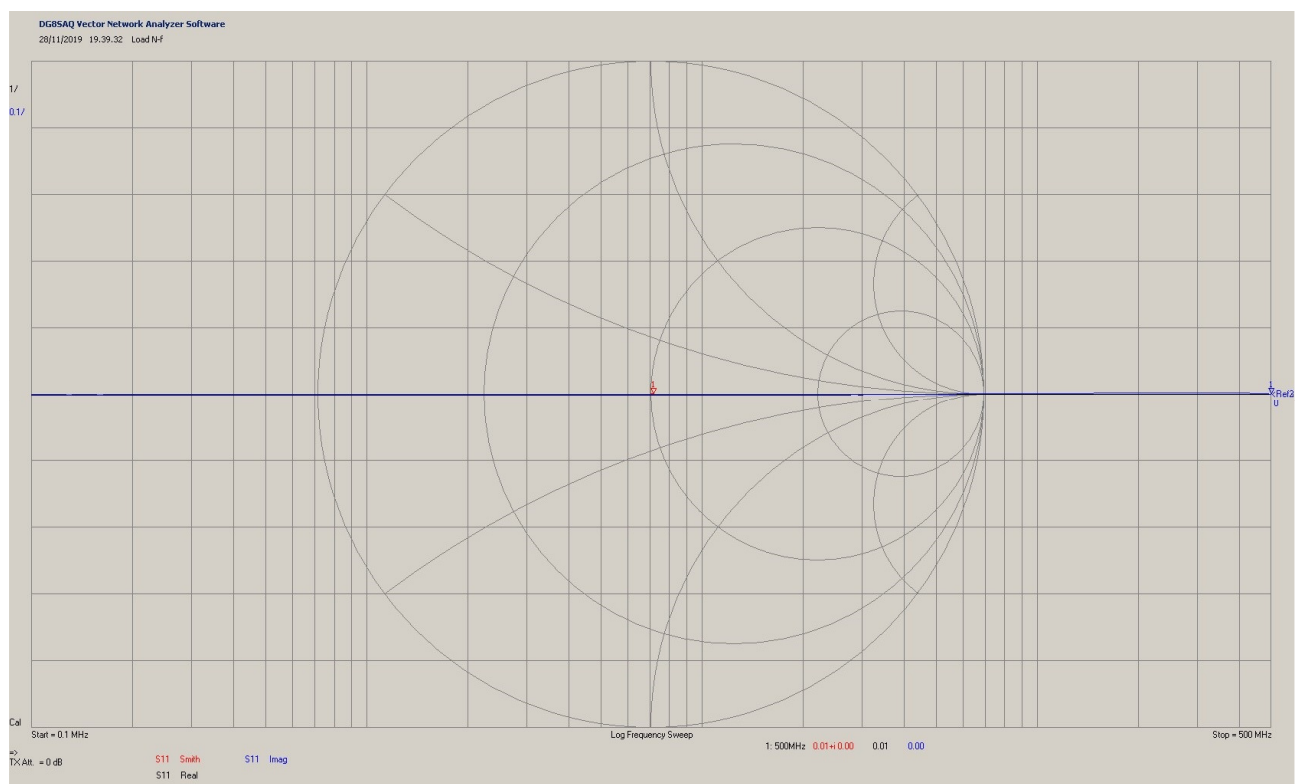


Figure 8.7: homemade LOAD, with  $C_{//}=2.7$  fF

I want to remind you that varying the delay in the extension port with the Load inserted does not modify at all the  $S_{11}$  trace; as we said before, the Load that we use to calibrate gets taken by the VNA as if it was ideal, therefore the reflected wave for the VNA gets compensated to zero.

Once the SOL kit is completed, we are adding the Thru. For this purpose, let's take the adapter N-f/N-f of figure 8.8 and with a caliper, let's measure the distances in air and in the PTFE.



Figure 8.8: Thru N-f/N-f

From the measures found, we can see that the segment in air is 10.53 mm long, while the one in the PTFE is 9.25 mm long. From this, with the 5.7, we obtain the following delay:

$$t_T = t_{air} + t_{PTFE} = \frac{d_{air}}{v_{air}} + \frac{d_{PTFE}}{v_{PTFE}} = 10,50 \cdot 3,333 + 9,30 \cdot 4,830 = 79,91 \text{ ps} \quad 8.9$$

Therefore, the delay of the Thru - being passed through by a wave only once - becomes:

$$THRU_{oneway} = 79,91 \text{ ps} \quad 8.10$$

With the same technique, we could build other Sol kits, but in general, the SMA, the N and the PL are enough. The latter does not require big characterizations because it works up to HF. In the last chapter we'll see an example of construction of a SOL kit with PL female connector.

## 9. Measuring an inductor.

An ideal inductor is characterized only by the inductance  $L$ , which remains constant at the varying of the frequency and therefore, it entails a complex impedance  $Z_L$ ; the module of the impedance grows linearly with the frequency.

$$Z_L = j\omega \cdot L = j2\pi f \cdot L \text{ Ohm} \quad 9.1$$

If we connect an ideal inductor at the end of an ideal line having impedance  $Z_0 = 50 \Omega$ , it will present a complex reflection coefficient of  $\Gamma = S_{11}$  at port 1:

$$\Gamma = S_{11} = \frac{Z_L - Z_0}{Z_L + Z_0} = \frac{j\omega L - 50}{j\omega L + 50} \quad 9.2$$

The equation 9.2 is obtained from 2.7, by making the scatter parameter  $S_{11}$  explicit (parameter which I remind you being equal to the reflection coefficient). As you can observe, the reflection coefficient varies at the varying of  $\omega$ , and when  $\omega$  grows, it goes from a value of -1 (short circuit) to a value of 1 (open circuit). Basically, it moves following the external upper semi-circumference of the Smith chart, starting from the left (short circuit), figure 3.3. For this reason, the module of the  $S_{11}$  parameter remains equal even when  $\omega$  grows, while its phase goes from +180 to 0 degrees. As you can see, the behavior of the  $S_{11}$  parameter, aka, the reflection coefficient  $\Gamma$ , is very different compared to the one of the impedance  $Z_L$ . Furthermore, the impedance  $Z_L$  does not get measured directly, but it gets calculated from the  $S_{11}$  parameter (formula 2.7), while the  $S_{11}$  parameter is obtained with an RF bridge (formula 2.2). The sensitivity of the instrument is not the same for all the impedances that appear on the conventional reference plane, but it has its maximum around the reflection coefficients with a phase of +90 ( $Z_L = +j50$ ); however, from 42° to 135° it maintains a good stability.

Unfortunately, the ideal inductor does not exist in real life, so we will have to find a model that describes the real inductor in the best way. Up until 500MHz the most used model [2] is the one of figure 9.1.



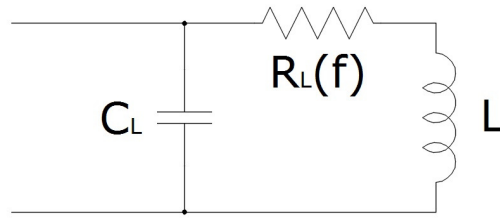


Figure 9.1: Model of a real inductor

Of course, the model of figure 9.1 comprehends the ideal inductor  $L$  with a parasite resistance  $R_L$  in series, whose value depends on the frequency ( $f$ ) because it represents the skin effect of the conductors (and/or the ferromagnetic material if present) and the parasite capacity  $C_L$ ; this is due to the leads, and the vicinity of the winding turns. I won't talk any further about this, since this is not the place, but if you want to delve deeper into the topic refer to [2]. The model of figure 9.1 presents a very different impedance compared to the pure inductance that we previously examined, and, in synthesis, it goes as follows:

$$\mathbf{Z}_L = \frac{\frac{1}{j\omega C_L}(R_L(f) + j\omega L)}{\frac{1}{j\omega C_L} + R_L(f) + j\omega L} = \frac{R_L(f) + j\omega L}{1 + j\omega C_L R_L(f) - \omega^2 C_L L} \quad \text{Ohm} \quad 9.3$$

After some calculations, the equation 9.3 turns out to be inductive up until the resonance frequency  $f_0$ , but then it becomes capacitive, becoming purely resistive at resonance.

As a matter of fact, the resonance frequency of the 9.3 is the one where the reactive part (imaginary) is nullified; it is:

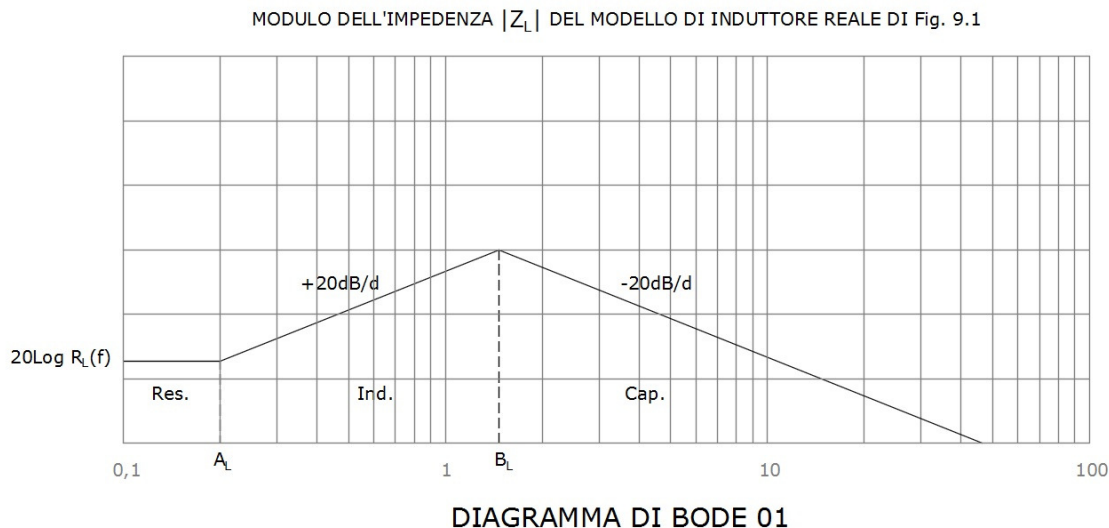
$$f_0 = \frac{1}{2\pi} \sqrt{\frac{1}{LC_L} - \frac{R_L^2(f)}{L^2}} \quad \text{Hz} \quad 9.4$$

Equation 9.4 will create a real resonance frequency only if the element under the square root is positive; this will happen only if:

$$L \geq R_L^2(f) C_L \quad 9.5$$

The conditions in 9.5 are almost always satisfied, because the value of the inductance is usually very big compared to the other term. Finally, if we observe the formula 9.3, we'll see that at low frequencies of  $f < 1$  kHz ( $\omega = 2\pi f$ ) the impedance  $\mathbf{Z}_L \approx R_L(f) \approx 0$  Ohm; I want to remind you that  $R_L(f)$  represents the parasite resistance of the coiled conductor, constituting the inductor, while the influence from the frequency  $f$  is due to the skin effect. At high frequencies, the capacity tends to shunt both the resistance and the inductance. Therefore, in those instances, the impedance  $\mathbf{Z}_L$  is a short-circuit.

The Bode plot of the impedance 9.3 is the following:



The main points are:

$$A_L = \frac{R_L(f)}{2\pi L} \text{ Hz}$$

$$B_L = \frac{1}{2\pi\sqrt{LC_L}} \text{ Hz}$$

Therefore, on the Smith chart we'll see a trace that starts from point A of figure 3.2, rotates clockwise through the upper area of the Smith chart (the inductive one) and, if the  $f$  is sufficiently big, the trace continues to rotate clockwise until it reaches again point A, going through the lower semi-circumference (the capacitive one). The point where the trace crosses the horizontal diameter of the Smith chart will represent the parallel resonance; you might notice that the frequency  $f_0$  is the one of formula 9.4. Let's verify.

Let's place our VNA at the end of a line segment (something like the one I used in figure 7.1) and calibrate it with the female SOL kit; let's set 1000 points of measure and an assessment time of 20 mS per point.

After doing so, let's take an SMA-f connector for PCS (like a Huber+Suhner code 82\_SMA-50-0-1) and measure the distance with the caliper, which is  $d_f=7,2$  mm; then, let's calculate the delay  $t_d$  with the formula 5.7:

$$t_d = d_f \cdot \frac{1}{VF \cdot c} = 7,2 \cdot 4,831 = 34,78 \text{ ps} \quad 9.6$$

After having done that, let's bring the conventional reference plane onto the measurement plane by inserting the delay  $t_d$  obtained with the 9.6 in the VNA extension port 1. Now we're ready to carry out some measurements; let's weld a random inductance onto the SMA-f connector for PCS (figure 9.2) and let's screw everything at the end of the cable segment where we made the SOL calibration.



Figure 9.2: Inductor on SMA  $f$  for PCS

Let's launch the scan and after 20 seconds the result will be the following:



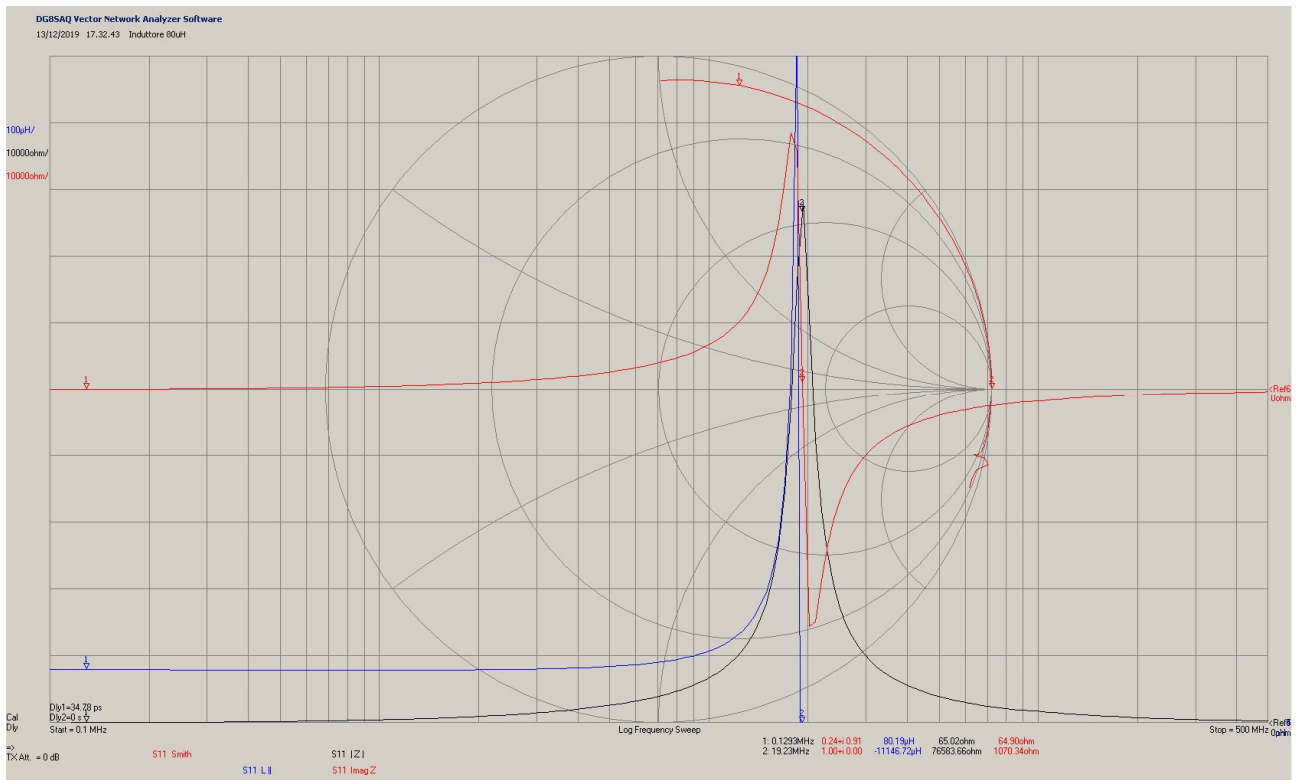


Figure 9.3: Measure of a 82uH inductor

The blue line in figure 9.3 represents the inductance, that remains constant up to 3MHz and the resonance is manifested at  $f_0=19,23\text{MHz}$  (marker 2). The nominal value of the inductance is read at low frequencies (marker 1  $\approx 100\text{KHz}$ ) 80.19 $\mu\text{H}$ . Notice that the red trace on the Smith chart starts at the top, close to marker 1, because the trace rotates inside the upper semi-circle (the inductive one), it passes the horizontal diameter of the Smith chart at the resonance point (marker 2  $\approx 19,23\text{ MHz}$ ) and then it enters the lower semi-circle becoming a capacitive reactance, as expected from the model of figure 9.1. The black trace represents the impedance  $\mathbf{Z}_L$  module, while the red trace in the middle of the graph shows its reactance. Finally, at resonance, the module of the impedance is  $|\mathbf{Z}_L|=76,5\text{ K}\Omega$ , therefore we're actually looking at a parallel resonance.

If we simplify the formula 9.4, hypothesizing that  $R_L(f)\approx 0$  at low frequencies, we could approximate the value of the parasite capacity  $C_L$ :

$$C_L \approx \frac{1}{L \cdot (2\pi f_0)^2} = \frac{1}{80,19 \cdot 10^{-6} \cdot (6,28 \cdot 19,23 \cdot 10^6)^2} = 0,854\text{pF} \quad 9.7$$

Summing up, the inductor works well up to a few MHz, then the parasite elements become too big; however, having the graph of the inductance movements, one can precisely know the value of device at each frequency included in the range of measure. An LCR meter would probably be more precise, but the measure gets carried out at a fixed frequency, usually at 1KHz, while with the VNA one can know the frequency up to which we can use an inductor and obtain its parasite parameters at the desired frequency.

Another example of inductor is the one in figure 9.4, which is an inductor in air.

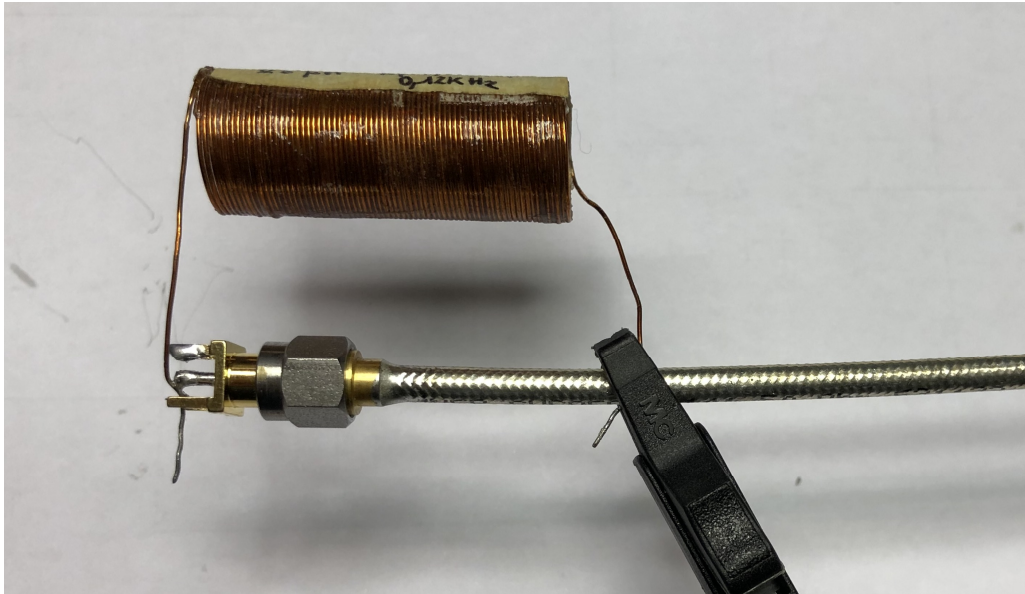


Figure 9.4 In-air inductor

Using a conformable coaxial conductor without an external sheath allows to simplify the connection of the “cold pole” of the inductance with the use of tweezers. This introduces a parasite inductance that makes the measure less accurate. Not much would have changed if the inductor would have been connected to the measurement plane, since this would have increased the parasite capacity due to the vicinity of the leads. In both cases we’re introducing an error; for this reason, if you can, try to stay as close as possible to the measurement plane, because it will be the best compromise between comfort and accuracy. The scan creates the diagrams of figure 9.5.

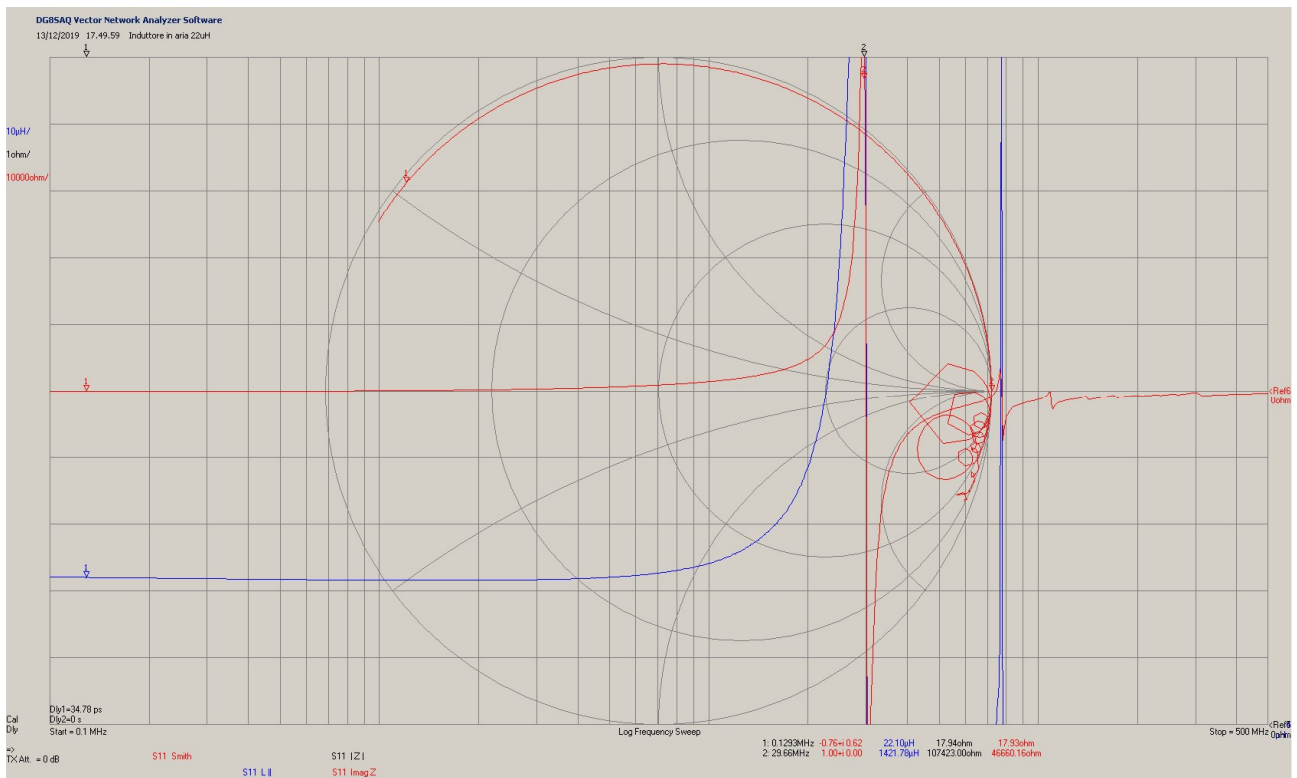


Figure 9.5: Graphs of the in-air inductor

The inductor, by and large, behaves like the previous one; it has a parallel resonance of  $f_0=29,66$  MHz (marker 2), then, after the usual capacitive tract, it presents numerous other resonances outside the scope of our circuital model. The “nominal” inductance is of  $22\mu\text{H}$ . To conclude, let's measure an inductor that I'll use to build a trap for the 80m. The coil is sized with a very good software that you can find in [10]. The physical parameters of the coil are: support in PE of 50mm, 22 coils of enameled wire of 1.3mm and length of the coiling of about 30mm. You can see the result in figure 9.6.

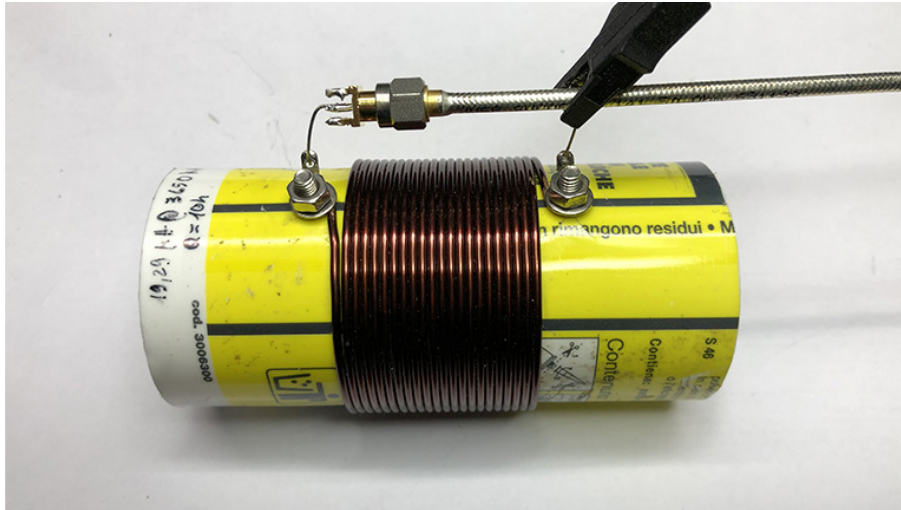


Figure 9.6: Coil for 80m trap

Since the trap will have to work at 3850 KHz, I have narrowed the measuring range from 1000KHz to 30MHz. The variation in the frequency range forces us to carry out a new calibration with the SOL kit.

Once that is out the way, we'll obtain the graph of figure 9.7.

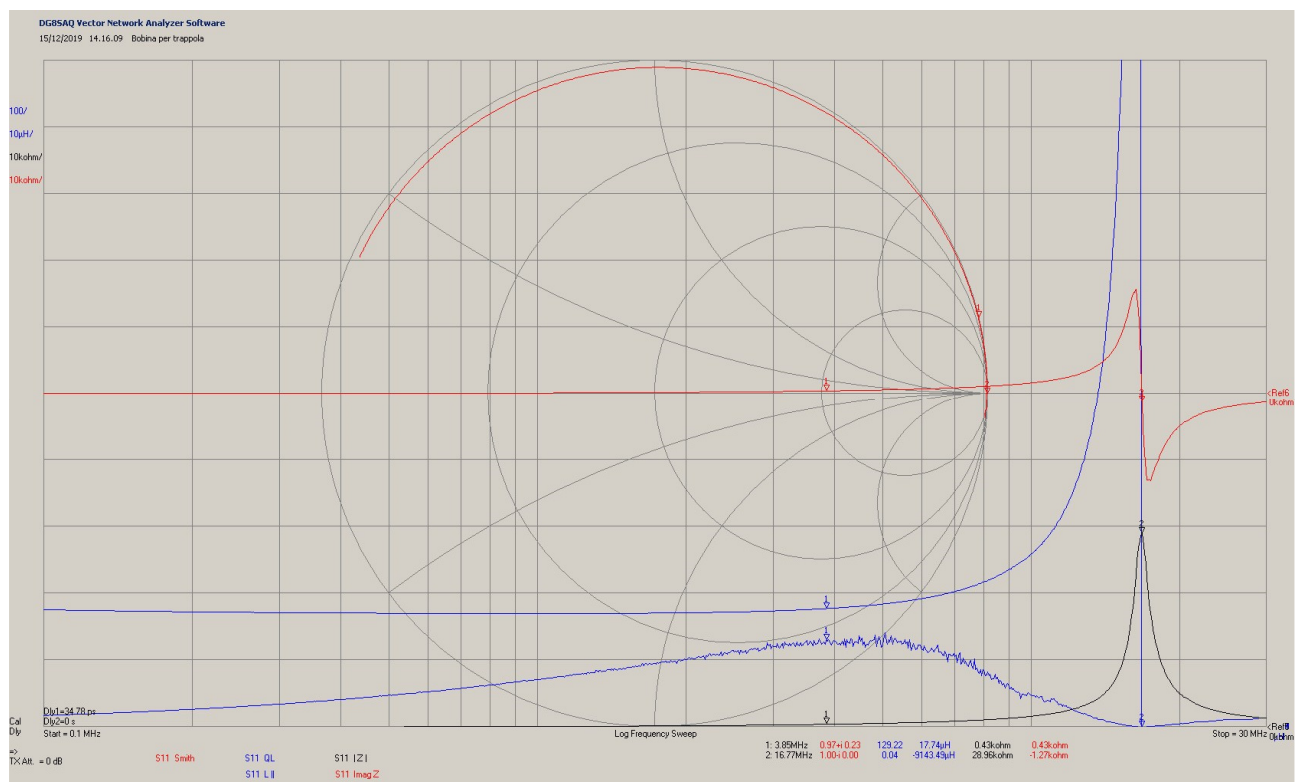


Figure 9.7: Coil graphs for 80m trap

The inductor behaves in compliance with the theoretical model; marker 1, placed at 3850KHz, indicates that the inductance is 17.74  $\mu$ H. The blue trace represents the trend of the  $Q_U$ , that is 129 in correspondence of marker 1, while the frequency of resonance is  $f_0=16.77$  MHz. With said data is possible to completely characterize the inductor. By reworking equation 2.9, with the value of the  $Q_U$  and with the imaginary part of the impedance, we can obtain the value of the parasite resistance at the desired frequency  $f=3850$ KHz.

$$R_L(f) = \left| \frac{X_L}{Q_U} \right| = \frac{2\pi f \cdot L}{Q_U} = \frac{429.1}{129.2} = 3.32 \text{ Ohm} \quad 9.8$$

If we approximate the parasite capacity  $C_L$  with formula 9.7, we will have characterized completely our real inductor.

## 10. Measuring a condenser.

An ideal condenser is defined by the sole capacity  $C$ , which remains constant at the varying of the frequency, and therefore it has a complex impedance  $Z_C$ ; the impedance module decreases hyperbolically with the frequency.

$$Z_C = \frac{1}{j\omega \cdot C} = -j \frac{1}{2\pi f \cdot C} \text{ Ohm} \quad 10.1$$

If we connect an ideal condenser at the end of an ideal line, having impedance  $Z_0 = 50\Omega$ , it will present the following reflection coefficient  $\Gamma$  at port 1:

$$\Gamma = S_{11} = \frac{Z_C - Z_0}{Z_C + Z_0} = \frac{1 - j50\omega C}{1 + j50\omega C} \quad 10.2$$

The 10.2 is obtained from 2.7, by making the scatter parameter  $S_{11}$  explicit. Even in this case, the reflection coefficient varies at the varying of  $\omega$ , but it goes from the value of 1 (open circuit) to the value of -1 (short circuit) as  $\omega$  grows. It basically travels along the outermost lower semi-circumference of the Smith chart (figure 3.3, starting from the right).

The module of the  $S_{11}$  parameter is always equal to 1, while its phase goes from 0 to -180 at the growing of  $\omega$ . As you can imagine the behavior of the  $S_{11}$  parameter is similar to the one of the inductor mentioned in the previous paragraph.

Given that the impedance  $Z_C$  does not get measured directly but it gets calculated from the scatter parameter  $S_{11}$  (formula 2.7), the sensitivity of the device is not the same for all the impedances that appear on the conventional reference plane; its highest accuracy is found around the reflection coefficients with a  $-90^\circ$  phase ( $Z_C = -j50$ ), but the sensitivity remains overall good from  $-42^\circ$  to  $-135^\circ$ .

Unfortunately, the ideal condenser does not exist, therefore we had to find a model that describes the real condenser in the best way possible.

Up to 500MHz the most common model is the one in figure 10.1 (but there are others that are more sophisticated) [2]. In addition to the real capacity, the model also includes the parasite inductance  $L_c$ , the leads and the parasite resistance  $R_c$ , which groups all the dissipative phenomena and is called Equivalent Series Resistance, or ESR.

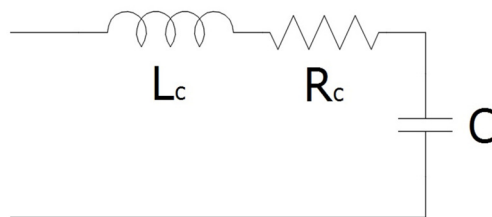


Figure 10.1: Model of real condenser

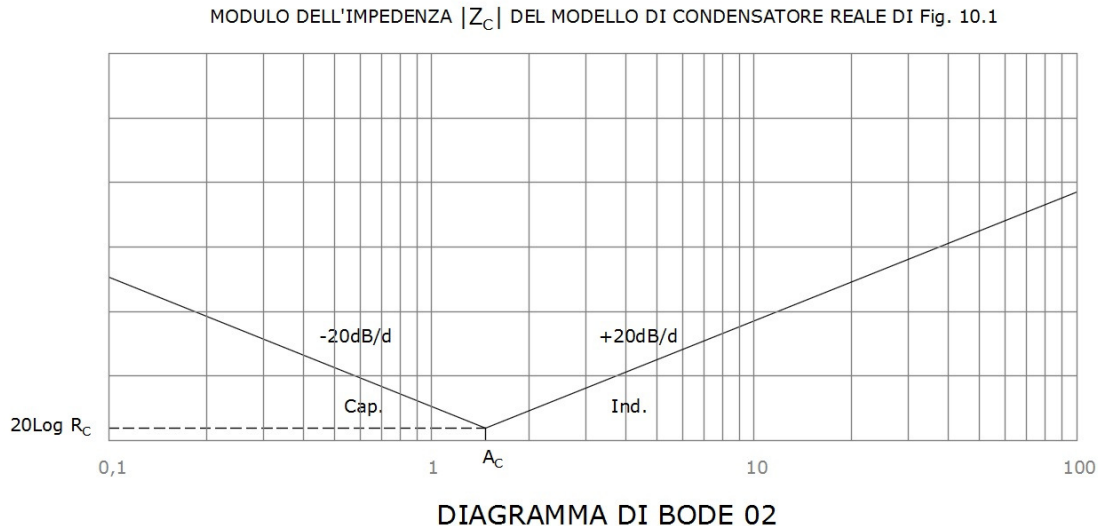
We recognized it right away; it consists of a resonator series and its impedance  $Z_C$  varies at the varying of the frequency in such way:

$$Z_C = R_C + j \left( \omega L_C - \frac{1}{\omega C} \right) \text{Ohm} \quad 10.3$$

The corresponding frequency of resonance is:

$$f_0 = \frac{1}{2\pi\sqrt{L_C C}} \text{Hz} \quad 10.4$$

If we observe the formula 10.3, we will see that at low frequencies  $f < 1 \text{kHz}$  the capacitive component of the impedance  $Z_C$  will be dominant both over the inductive component and the resistive one. At high frequencies the inductive component of  $Z_C$  prevails over the other two. Therefore, both at high and low frequencies, the impedance becomes an open circuit. The Bode diagram of the impedance 10.3 is the following:



The main point is:

$$A_C = \frac{1}{2\pi\sqrt{L_C C}} \text{Hz}$$

On the Smith chart we will see a trace that starts from point B of figure 3.2, goes through the lower part of the chart (the capacitive part) in a counter clockwise motion and, if  $f$  is sufficiently big, it continues to rotate clockwise, till it reaches point B again through the upper semi-circle (the inductive one). In the point where the trace crosses the horizontal diameter, we will find the series resonance. If you annotate the frequency, that will be  $f_0$  from formula 10.4.

Now let's measure a disc condenser with the nominal value of 47pF using the same setup used for the inductor; see figure 10.2.

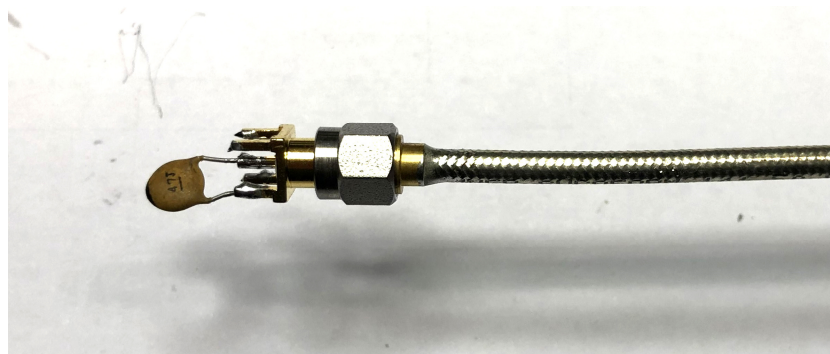


Figure 10.2: Disc Condenser

The scanning produces the graphs of figure 10.3.



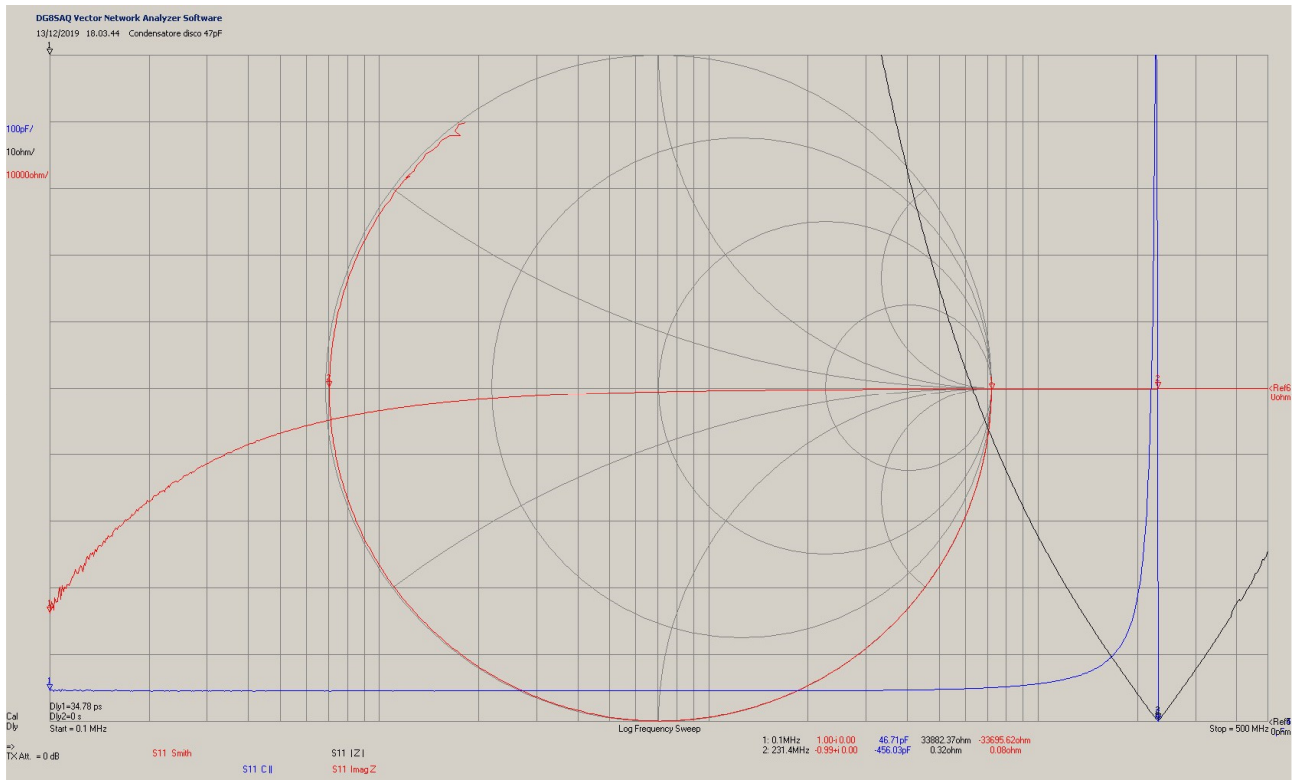


Figure 10.3: Graphs of disc condenser

Even the graphs of the condenser are following the theory. The red trace on the Smith chart starts from the open circuit (marker 1), runs through all the lower semi-circle at capacitive reactances, encounters the horizontal diameter in the resonance point  $f_0$  (marker 2 at 231 MHz) and continues in the upper semi-circumference with positive reactance where the inductive reactance prevails. The resonance  $f_0$  is a series type because the module of the impedance at the resonance is close to zero (black trace). The nominal capacity, read at 100 KHz, is  $C=46.7\text{pF}$  and it remains constant till 50 MHz.

Summing up, we have a good condenser that can be used over all the HF band till the 6m band.

Let's now characterize a knobdoor condenser, of soviet manufacture of 47 pF (6KV 7KVAR) that you can see in figure 10.4

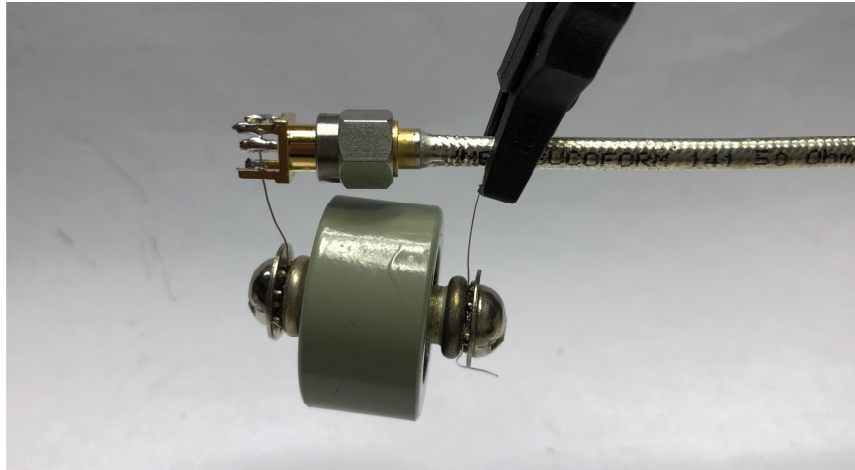


Figure 10.4: Knobdoor condenser 47pF

Figure 10.5 shows what we get after the scan.

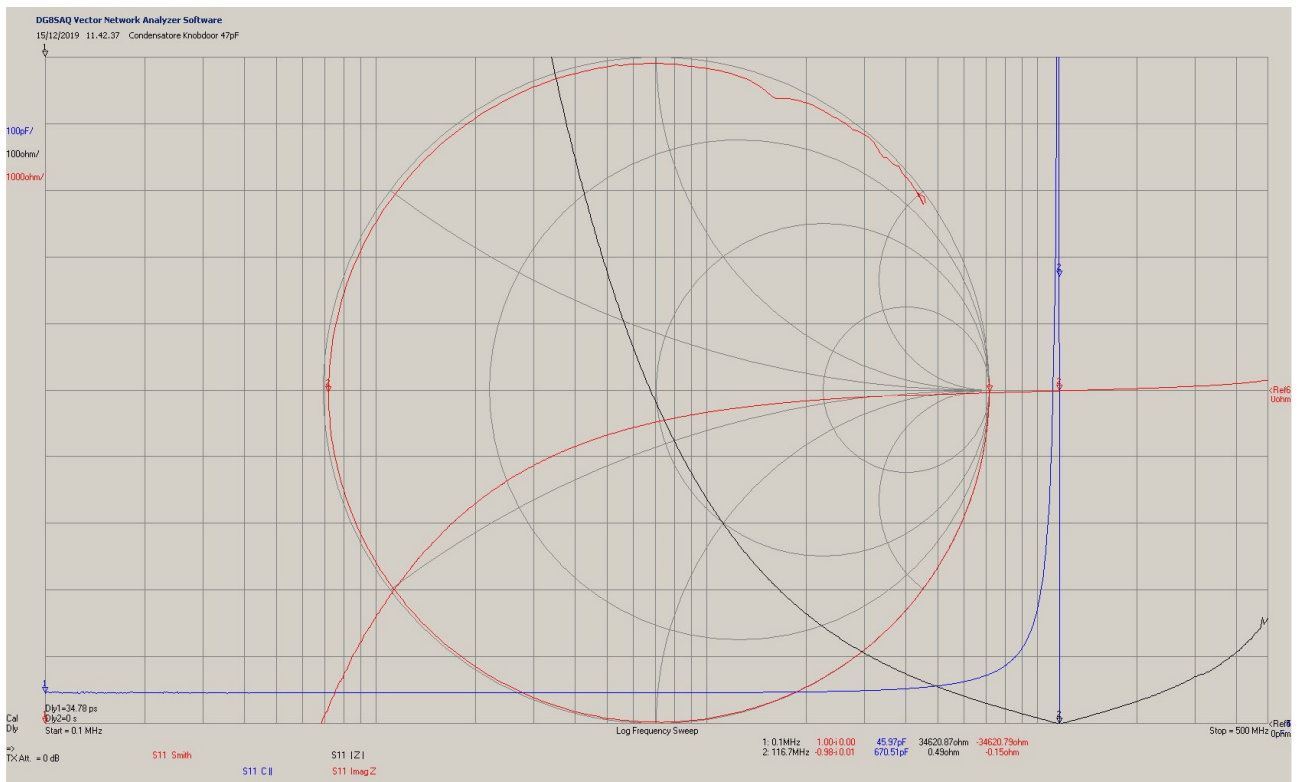


Figure 10.5: Graph of the knobdoor condenser 47pF

The nominal capacity, read at 100KHz is  $C=45.97\text{pF}$  and it maintains pretty much constant till 30MHz; the series resonance (marker 2) happens at  $f_0=116.7\text{MHz}$ .

The parasite resistance  $R_C$  is read at the resonance and is  $R_C=0.49\ \Omega$ , while the parasite inductance can be obtained by reworking the formula 10.4

$$L_C = \frac{1}{C \cdot (2\pi f_0)^2} = \frac{1}{45.97 \cdot 10^{-12} \cdot (6.28 \cdot 116.7 \cdot 10^6)^2} = 40.46\ \text{nH} \quad 10.5$$

## 11. Measuring a resistor.

An ideal resistor is characterized by the sole resistance  $R$  that remains constant at the varying of the frequency; therefore, it has a complex impedance  $\mathbf{Z}_R=R$ , module of which is fixed and real (it doesn't have a reactive component).



If we connect an ideal resistor at the end of a transmission line having impedance  $Z_0=50\ \Omega$ , it will have a constant reflection coefficient  $\Gamma$  at port 1; on the Smith chart it will correspond to a point on the horizontal axis of the resistances. If  $R$  is lower than  $50\ \Omega$  said point will be located to the left of the centre, while to the right if higher. When it is equal to  $50\ \Omega$  it will be found exactly in the centre of the chart.

The real resistor, on the other hand, is entirely different; the model that best describes its behaviour up to 500MHz is the one in figure 11.1 [2].

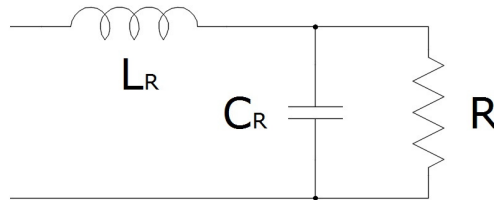


Figure 11.1: Model of a real resistor

Beyond the ideal resistance  $R$ , we will find the parasite inductance  $L_R$  and the parasite capacity  $C_R$ , both caused mainly by the leads.

The impedance  $Z_R$  of the model in figure 11.1 is the following:

$$Z_R = j\omega L_R + \frac{\frac{1}{j\omega C_R} R}{\frac{1}{j\omega C_R} + R} = j\omega L_R + \frac{R}{1 + j\omega C_R R} \text{ Ohm} \quad 11.1$$

The frequency of resonance  $f_0$  is:

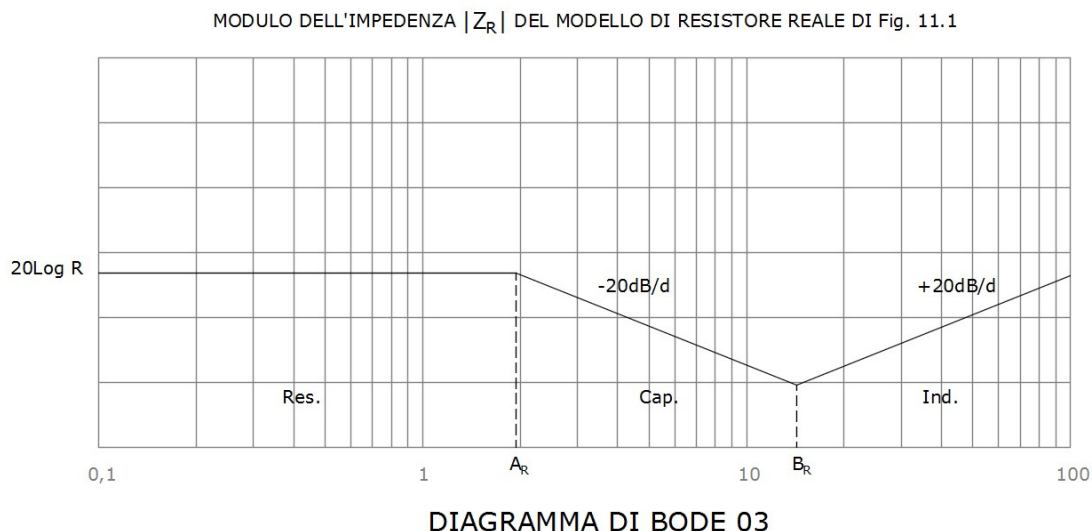
$$f_0 = \frac{1}{2\pi} \sqrt{\frac{1}{L_R C_R} - \frac{1}{C_R^2 R^2}} \text{ Hz} \quad 11.2$$

The equation 11.2 will give origin to a real frequency of resonance only if the element under square root is positive, and this happens only if the following condition is met:

$$R \geq \frac{L_R}{C_R} \quad 11.3$$

The condition 11.3 in our cases does not happen. Most of the times the parasite capacity  $C_R$  is very small, of the order of femto farad; therefore, the term to the right in the 11.3 will end up being way larger than  $R$ , hence the absence of a real resonance.

If we observe the formula 11.1, we will see that at low frequencies  $f < 1\text{kHz}$  both the capacitive component and the inductive one of  $Z_R$  are negligible, therefore  $Z_R = R$ . At high frequencies the inductive component becomes more and more important, and the condenser tends to



short circuit the resistance  $R$ . For this reason, at high frequencies the impedance will be mainly inductive  $Z_R \approx j\omega L_R$ ; therefore, the impedance will tend to become an open circuit. The Bode diagram of the impedance 11.3 is the following:

The main points are:

$$A_R = \frac{1}{2\pi R C_R} \text{ Hz}$$

$$B_R = \frac{1}{2\pi \sqrt{L_R C_R}} \text{ Hz}$$

On the Smith chart we will see a trace that starts from a definite point on the horizontal diameter, enters the upper area of the chart (the inductive one) in a counter clockwise motion and, if the  $f$  is sufficiently large, it will terminate at point B of figure 3.2. The way in which the trace reaches point B will depend on the values of the parameters  $L_R$ ,  $C_R$  and  $R$ .

Like in the previous cases, after the SOL calibration, we will connect the usual metallic film resistor of  $100 \Omega$  (RS 755-0707) on the SMA f connector for PCS like in figure 11.2.

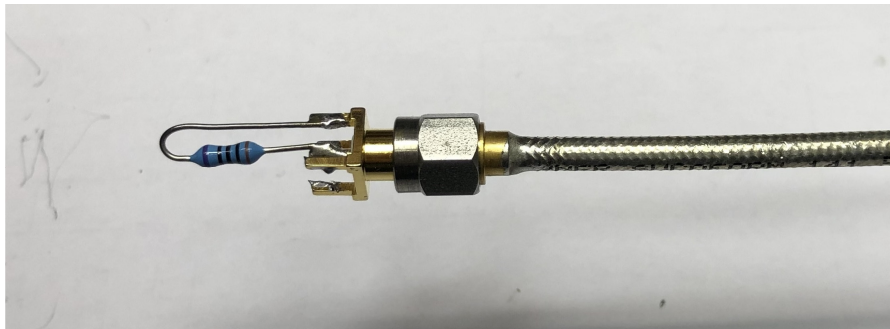


Figure 11.2: Metallic film resistor 100Ohm

The graph that results after the scan is the one in figure 11.3.

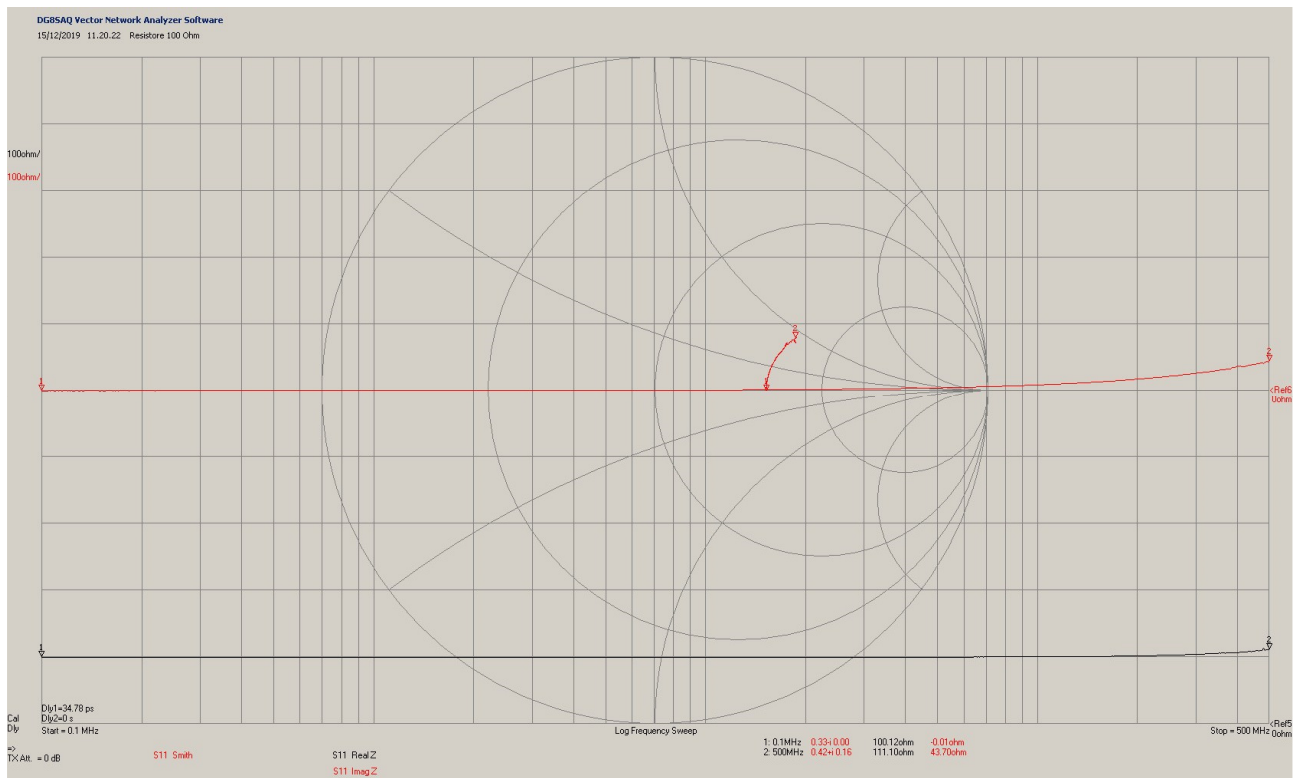


Figure 11.3: Graphs of metallic film resistor 100 Ohm

The red trace on the chart starts from point (2; 0), enters the upper semi-circle of the chart with positive reactance and tries to reach the point with infinite impedance (point B of figure 3.2), following the theoretical model of figure 11.1. The black trace indicates the module of the impedance  $|Z_R|$  that remains flat until almost 200MHz. The nominal value, measured with a 4-wire multimeter, is  $R=100.10\ \Omega$  and corresponds to what the VNA measures at 100KHz (100.12  $\Omega$ ). There are no noticeable resonances within the measuring range, therefore it becomes difficult to estimate parasite components.

To do so I used the function OPTIMIZE of the VNWA, which suggest  $L_R=139\text{nH}$  and  $C_R=0.381\text{pF}$ . With the suggested values and the resistance  $R$  measured with the multimeter, the model of figure 11.1 turns out to be the closest to the  $S_{11}$  response obtained with the VNWA. We're still talking about approximative values, linked to the model of resistor hypothesized.

## 12. Thru measure.

Up until now we have carried out measurements using only one port (port 1); now let's introduce the other port of the VNA (port 2).

The measurements with both ports are mostly used to obtain the transmission coefficient  $T$ , which is basically the  $S_{21}$  parameter (formula 2.13). To ensure that the measurement is accurate, we should make sure that the device under test, or DUT, does not produce undesired reflections from the unused ports. This means that the input and the output of the DUT have to be at the same impedance of the power line (usually 50  $\Omega$ ) [1] [2] [4]. The input and output adapted to 50  $\Omega$  only occurs in special cases, such as with amplifiers or attenuators, but not on all the nets (eg. not with the filters with high  $Q$ , like the ones made of quartz).

Therefore, when there is a mismatch, we should adopt some precautions to make sure that the measurements on ports 1 and 2 are not influenced by reflections produced by mismatched ports, which are not involved in the measurements. The most common precaution is the HW one, where some LC networks are inserted to create a temporary energetic adaptation.

The other system is the SW one, which consists in solving the system in 2.18; however, not all the VNA can do this. The VNWA of DG8SAQ has a tool that allows for the SW correction of the mismatch [6].

Having said so, let's see how to measure the attenuation of a coaxial cable with  $Z_0=50\ \Omega$ . It is a simple example that uses the  $S_{21}$  parameter and does not require special adaptation networks, since the characteristic impedance of the chosen cable is 50 $\Omega$ ; in the opposite case, it wouldn't be correct.

The coaxial cable that we will analyse is a piece of RG58 A/U MIL, 10.125m long, that I found already headed with PL connectors in my shack.

After the SOLT calibration and having checked that the  $S_{21}$  trace is flat at 0dB over all the band of measurement, I substituted the bullet with the cable segment (obviously I had to use SMA-f/PL-f adaptors). After that, I launched the scan: you can see the result in figure 12.1.

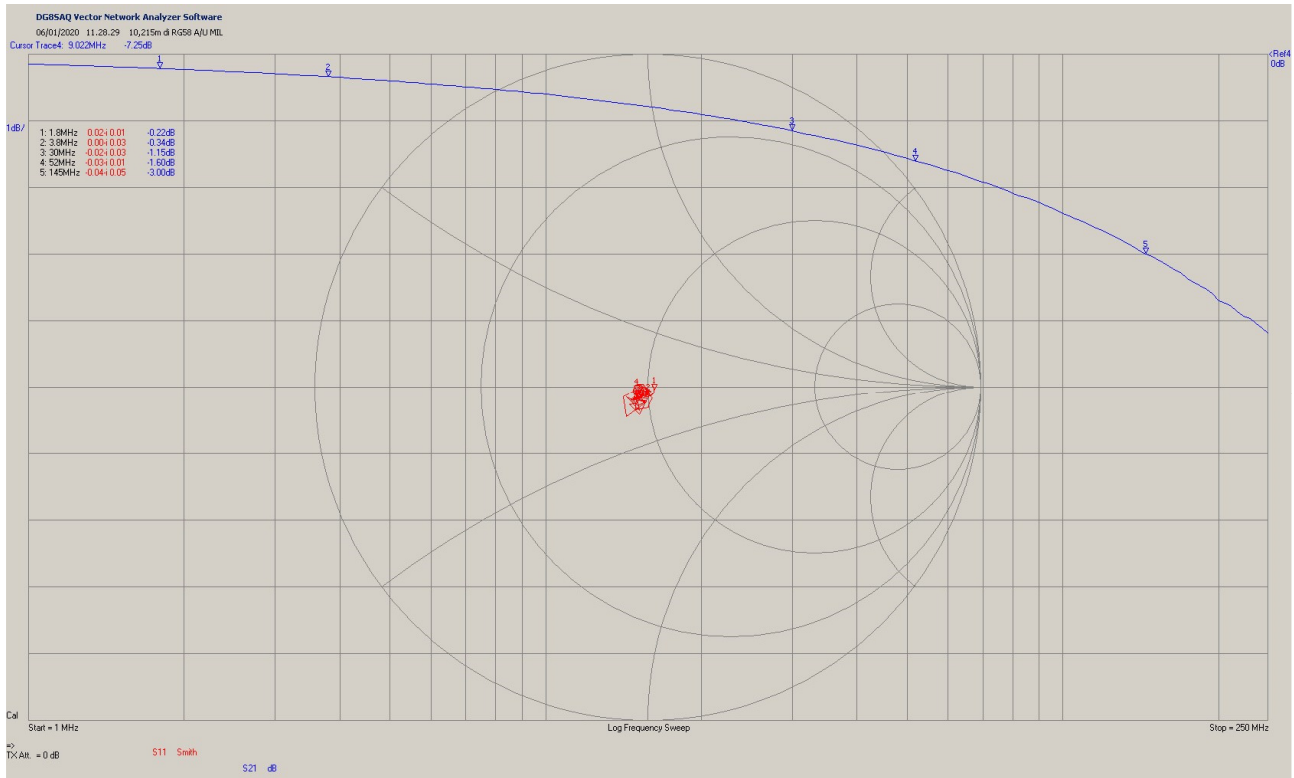


Figure 12.1: 10,215m of RG 58 A/U MIL

The blue trace represents the parameter  $S_{21}$ , meaning the attenuation of the transmitted line (see formula 2.15), while the red trace is the parameter  $S_{11}$ , that starts from 50 Ohm, but then, at the growing of the frequency, it becomes denser around point  $|Z_0|=49$  Ohm. From the absolute values of attenuation in dB, read in correspondence of the markers and with the equation 12.1, we can obtain the attenuation in dB per meter at the various frequencies:

$$A_f = |S_{21}|/l \quad 12.1$$

Furthermore, we could calculate the loss constant in the conductor  $K_1$  and the loss constant in the dielectric  $K_2$ , by solving the system in 12.2 [11]:

$$\begin{cases} A_{f1} = K_1 \cdot \sqrt{f_1} + K_2 \cdot f_1 \\ A_{f2} = K_1 \cdot \sqrt{f_2} + K_2 \cdot f_2 \end{cases} \quad 12.2$$

$F_x$ [MHz]	Att [dB]	A [dB/m]
1,8	0,22	0,022
3,8	0,34	0,033
30,0	1,15	0,113
52,0	1,60	0,157
145,0	3,00	0,294

Table 2

Once the parameters  $K_1$  and  $K_2$  are known the attenuation at the wanted frequency can be calculated. For convenience I am reporting the values measured in table 2.

Notice that at 145MHz the attenuation is already at 3dB; this means that the cable is too long for those frequencies and the losses are already relevant.

Once the system in 12.2 is solved in the range of use of the cable (therefore between 1.8 and 52MHz) we will find that the loss constant in the conductor is  $K_1=1.51694E-2$  and the loss constant in the dielectric is  $K_2=9.15612E-4$ .

With those constants and the length of the cable  $l$  we can calculate the attenuation at the chosen frequency  $f$  (in MHz) using the formula 12.3, provided that  $f$  is in the range used for the calculation.

$$A_T = (K_1 \cdot \sqrt{f} + K_2 \cdot f) \cdot l \text{ dB} \quad 12.3$$

For example, if we wanted to know how much attenuation we get with 25m of coaxial RG58 A/U MIL at 14MHz, we would find:

$$A_T = (K_1 \cdot \sqrt{f} + K_2 \cdot f) \cdot l = (1,51694E - 2 \cdot \sqrt{14} + 9,15612E - 4 \cdot 14) \cdot 25 = 1.74 \text{ dB} \quad 12.4$$

### 13. Conclusions.

In conclusion I believe that the VNA is an indispensable device for an OM who wants to know the parameters of his own devices. It is a complete device and although it requires a certain level of expertise to be used correctly, it will bring a lot of satisfying results and it will make your measures more precise and replicable.

With this article I hope I set the basis for a more informed and wise use of the device.

Lastly, I remind you that the simplest VNA, the one with one port, is the well-known antenna analyser that most of us are equipped with.

Therefore, make sure that your VNA can be calibrated with a SOL kit (even homemade) because only in that case you will be able to make the best out of its capabilities.

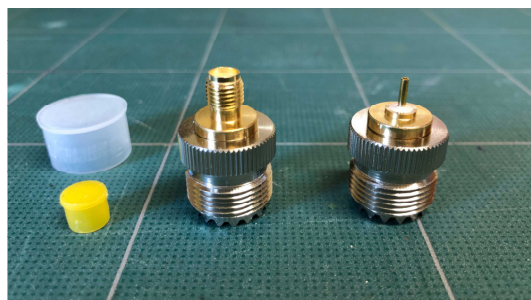


Figure 13.1: PL-f/SMA-f

The SOL kit in HF is not critical, it can easily be built with some PL-f/SMA-f adapters, by cutting the SMA part (like in figure 13.1) and inserting a short circuit for the Short and two resistances SMD 0805 of 100  $\Omega$  in parallel (RS 666-2437) for the Load.

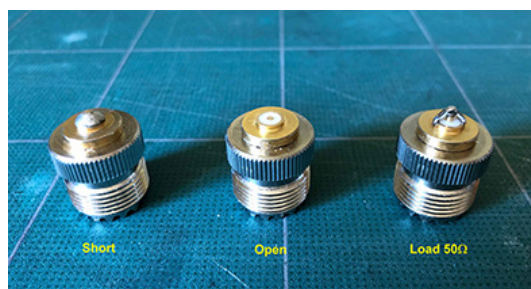


Figure 13.2: Kit SOL con PL-f

After the process the kit turns out like in figure 13.2.





Figure 13.3: Kit SOL with PL

If we wanted to overdo it, we could put them in a piece of aluminium like in figure 13.3. If we only use the SOL kit in HF, characterising the Load with a common tester will suffice, since the delays of the Open and the Short are non-significant in HF.

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