

PRACTICAL INSTRUCTIONS FOR THE RF AND MICROWAVE MEASUREMENT TUTORIAL

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Abstract

For this practical tutorial using (vector) network and spectrum analysers detailed instructions are given on how to operate these sophisticated instruments, avoiding the need for reading heavy instruction manuals. The tutorial is basically split into two parts, namely network analysis and spectrum analysis. In the network section the reflection and transmission characteristics of passive and active one- and two-ports are measured with a particular emphasis on resonators and cavities as well as calibration procedures. Spectrum analysers are used to examine amplitude- and frequency-modulated (AM, FM) signals, signal strength measurement, non-linear behaviour of amplifiers via intermodulation and the evaluation of noise and noise-figure properties.

1. INTRODUCTION

Radio-frequency (RF) spectrum analysers (SPA or SA) can be found in virtually every control room of a modern particle accelerator. They are used for many aspects of beam diagnostics, including Schottky signal acquisition and RF observation. We discuss here only the application of classical super-heterodyne SPAs and not systems based on the acquisition of time-domain traces and subsequent Fourier transform (FFT analysers). Such a super-heterodyne SPA is very similar (in principle) to any AM or FM radio-receiver. The incoming RF signal is moderately amplified (sometimes with adjustable gain) and then sent to the RF port of a mixer. A mixer is a non-linear element (containing one or more fast diodes) and acts like an analogue multiplier. The output signal of a local oscillator (LO) is connected to the LO input of the mixer and superimposed on the RF. As a consequence we obtain (non-linear) mixing products at the sum and difference frequency between RF and LO which appear at the intermediate frequency (IF) port of the mixer. This IF signal is subsequently band-filtered and sent to the vertically deflecting plates of the cathode ray tube (CRT) in the SPA (or via a demodulator to the loudspeaker in a radio-receiver). The word ‘super-heterodyne’ is a mixture of Latin and Greek meaning adding (super) a different (hetero) force (dyne), where the different force is no other than the LO signal, superimposed (added) in a non-linear fashion on the RF signal.

A network analyser can often be found in a control room but its real home is the RF laboratory. There are two kinds of instrument, namely the Scalar and the Vector Network Analysers (SNA and VNA, respectively). The SNA can only display the magnitude of some property (e.g. reflections or transmission coefficient) versus frequency. It basically consists of a tunable RF generator and a power detector. Often SPAs with a tracking generator (tracking to the actual frequency shown in the display) are used for scalar network analysis. However, with its additional phase display capability and much higher dynamic range with a simple power detector, the VNA offers a much wider range of applications than the SNA. The higher dynamic range (around 100 dB) is also linked to the fact that VNAs apply the mixing concept mentioned above. We are only using VNAs in this tutorial, but if a spectrum analyser with tracking generator is available one may also try using it as a scalar analyser. A typical RF laboratory application of a VNA is the measurement of the scattering parameters (S-parameters) of some device under test (DUT). In this tutorial we will perform most of the standard measurements on objects such as cavities, complex impedances including non-ideal resistors,

capacitors and inductors, filters and amplifiers. It will also be possible to do some ‘fun’ experiments (not described in this paper), e.g. using a coaxial waveguide transition as an antenna and building a simple radar system.

2. EXPERIMENTS WITH THE VECTOR NETWORK ANALYSER

2.1 General information about the VNAs used here

We use three different models of the HP8753 VNA, namely types B, C, and E. All have similar functions and frequency, power, and dynamic range, but use different software and have slightly different control menus. So, for example, the desired output power of the HP8753C is obtained by applying an external attenuator to an output power from -10 dB to $+20$ dB. For the HP8753D an output power range is selected (which internally switches attenuators) and then the desired output power is chosen. However, the instructions given here are also ‘in principle’ applicable to comparable instruments from other manufacturers.

The HP8753 has two kinds of control buttons: hard keys on the front panel, which are used to set the kind of measurement, the frequency range and other functions, and soft keys on the right-hand side of the display. In the following descriptions bold letters are frequently used to denote hard keys while normal letters denote soft keys: **MEAS**, REFL., FWD S_{11} (A/R), etc.

The HP8753 analysers have two display channels, which can be displayed in parallel in two modes: split screen and common screen. Each channel can be processed using mathematical functions, memory, data/memory display and other features.

Only one channel at a time responds to control inputs. Changes to some settings act only on the active channel, whereas frequency range and other principal parameters remain unchanged for both channels. For critical measurements the noise level can be improved either by averaging or reducing the IF bandwidth or by increasing the output power to the maximum allowed level. The VNAs have powerful marker functions to simplify search operations such as 3 dB bandwidths, maxima and minima and desired power levels. These can be combined with special mathematical functions to allow the calculation of the Q -factor of a resonator, for example.

All analysers used in the tutorial have a time-domain option, which can be found under **SYSTEM**, TRANSFORM MENU. A frequency range up to 2 GHz with 801 data points is selected for these measurements and **FORMAT**, MORE, REAL is used to adjust the display format. Note that the frequency range is automatically adapted to the values required for the transformation (chirp–Z type) when using the ‘low pass’ time-domain mode.

The ‘power-sweep’ function for a fixed Continuous Wave (CW) frequency can be applied for measuring the 1 dB compression point of an amplifier.

All new measurements are started with a **PRESET** of the VNA, and the desired type of measurements (S_{11} , S_{21} , or other) are to be keyed in. Then the frequency range is set and the measurement is calibrated with the proper choice of the calibration kit, for example N 50 Ω or 3.5 mm for SMA, depending on the connector required to connect the DUT.

2.2 Becoming acquainted with the HP 8753 (first VNA lesson)

First of all, it is very important to define the kind of measurement (e.g. reflection, transmission) to be performed.

The VNA provides the following basic choices:

MEAS S_{11} (A/R) REFL.: FWD	for impedance, VSWR, etc. at port 1.
S_{21} (B/R) TRANS.: FWD	for attenuation or gain, phase, group delay etc. between ports 1 and 2 (not from 1 to 2!).

S_{12} (A/R) TRANS.: REV for attenuation, gain etc. between ports 2 and 1.
 S_{22} (B/R) REFL.: REV for impedance, VSWR, etc. at port 2.

The notation A/R and B/R refers to the coaxial connections between the S-parameter test set and the actual VNA. These connections are externally accessible for certain instruments and permit special configurations. Here the letter A stands for the input to channel A and similarly B denotes the input of channel B. R is the reference channel. The instrument will essentially display (in complex formats such as real and imaginary or magnitude and phase) the (complex) ratios A/R or B/R. The S-parameter test basically set contains two directional couplers and several remotely controlled RF relays in order to establish the required connections between the two S-parameter ports (port 1 and port 2 at the front panel) and the connections A, B, R, and signal source. FWD stands for ‘forward’ measurement and REV for ‘reverse’.

The desired frequency range must be set once the type of measurement has been defined. The analysers start with their full range sweep, which is likely to be too high for measuring lumped elements like resistors and capacitors.

The following possibilities are available for setting frequencies:

either

START and
STOP

or

CENTER and
SPAN

Having set the frequency range, the influence of the mismatch in the connecting cables between the ports of the VNA and the DUT should be eliminated, as well as the impact of generator mismatch and finite directivity of internal couplers of the VNA. In addition, the reference plane will be moved to the end of the cable. In other words, the typically applied application for a reflection measurement (open, short, load) eliminates errors resulting from generator and cable mismatch, electrical length and losses, as well as finite coupler directivity. Of course the imperfections mentioned above cannot be suppressed by themselves, but by inferring the results from the calibration measurements a software routine allows for numerical correction.

The calibration procedures will be described in more detail in the first measurements, but one should always remember that *the calibration can only be done AFTER having defined the frequency range* (except for the interpolation mode when using frequency ranges smaller than the calibrated range)

- As an exercise, **PRESET** the instrument (push the green button on the HP8753).

The VNA starts with full range sweep, CH1 active in S_{11} mode with logarithmic magnitude (log mag) and reference level at 0 dB (REF). Dial a frequency span from 0.5 to 5 MHz using the Start and Stop controls.

Set start frequency to 0.5 MHz **START; 0.5; M/u.**

Set stop frequency to 5 MHz **STOP; 5; M/u.**

- Calibrate S_{11} (port 1) with the reference plane at the end of the cable. Typical RF cables are equipped with M-type (male) N-connectors on either end. Hence components in the calibration kit have F-type (female) N-connectors. During the calibration procedure a question will appear on the screen of the instrument asking which component of the calibration kit is being used. It should be

remembered that the kind of calibration is defined for the type of connector you are doing the calibration on (i.e. the end of the cable), NOT for the calibration-kit component itself!

By the way, the term N for the type of cable connector we are using here has its historical roots in the word navy, since this kind of connector was first used by the US Navy more than 50 years ago. The frequently applied competitor, the BNC connector, comes from the same shop (Bayonet Navy Connector). But the BNC has much lower performance than the N-connector.

- Now follow the menu on the calibration page and attach the open, short, and load as requested by the system.
- Don't forget to confirm that the calibration is done; you may want to store it for later use.
- For very critical and accurate measurements you may calibrate the system using the trace average function, but this is rather time consuming. However, you may reduce the IF bandwidth from 3 kHz (standard setting) to 100 Hz (in the average menu) and note the difference.
- As a next step select in the display menu the split screen display and then activate channel 2. For channel 2 we are measuring now in transmission from port 1 to port 2, i.e. S_{21} . Connect the end of the cable where you just did the S_{11} calibration to port 2 of the VNA.
- Carry out a 'response' calibration for transmission (not full two port) and store the result.

Now you are well prepared to measure simultaneously some DUT in reflection and transmission with a calibrated system. As a DUT you can use the 10 Ω resistor in a SUCO (blue) box. This DUT contains a simple 10 Ω carbon resistor between the inner conductors of the input and output connector. You can check now up to what frequency this test object looks like a pure resistor and what kind of parasitic effects occur. Despite the fact that the geometrical length of the DUT is just a few centimetres you will already see a considerable inductive component at 5 MHz. The free space wavelength of 5 MHz amounts to 60 m! For the next exercise disconnect the cable from port 2, attach the DUT and terminate the open connector of the DUT with a short (from the calibration kit).

- Using channel 1, select as display format the Smith Chart and look at the read-out of the marker on top of the screen: it gives a reading of the resistance and inductance of the lumped resistor. Which read-out changes significantly with frequency? Discuss the results.
- Exchange the blue test box for another one, look at the results in different formats, and turn the dial to measure at different frequencies.
- Now you may try a further method to measure the frequency-dependent complex impedance of the 10 Ω test box (leaving the DUT connected to the cable to port 1 with a short at its end). The instrument provides a conversion menu that allows direct display from the S_{11} measurement of the real and imaginary parts of the DUT's complex impedance as a function of frequency. This kind of conversion is possible from both the reflection and the transmission measurements. Try both techniques and discuss the reason for possible discrepancies. Play with the electrical-delay correction in the transmission-type test. How can you explain the (of course unphysical) negative real parts for certain electrical-delay settings?

2.3 Demonstration of calibration effectiveness

- Display the locus of S_{11} of a 25 Ω DUT in the Smith Chart for the frequency range 600–1200 MHz after having done a reflection calibration. This 25 Ω DUT may be made by using two 50 Ω terminations and a coaxial T-piece to connect them 'in parallel'.
- Now produce a severe generator mismatch with another coaxial T-piece inserted at port 1, between port 1 and the coaxial cable. The open port of this T-piece will be terminated with another 50 Ω load. In this configuration we have artificially modified the generator impedance from about 50 Ω to around 25 Ω .

- As a next step perform the usual open, short, load calibration at the end of your (say 1 m long) test cable. Reconnect the 25 Ω DUT at the end and display its characteristic in the Smith Chart. You should not now be able to see any significant difference from the result found in the first step.
- Then insert the triple-stub tuner between the end of the calibrated test cable and the 25 Ω load and by adjusting the length of the three stubs (cut and try method) try to achieve an impedance match ($S_{11} = 0$) around 900 MHz.

Sometimes one is not really sure whether a certain calibration procedure, in particular in reflection, has been done well. There are many possible ways of making a mistake and an independent cross-check is desirable. A rather sensitive test consists of connecting a short piece of coaxial cable with its end left open or shorted and then displaying the reflection coefficient in the Smith Chart. In case of a poor calibration you may find the locus of S_{11} exceeding the boundary of the Smith Chart, which would only be valid for an active element with a reflection coefficient larger than unity. For a passive device this kind of negative resistance response is impossible and indicates a calibration or display error. However, in the past reflection amplifiers were used for special applications (certain parametric amplifiers or negative resistance devices containing tunnel diodes).

2.4 Time-domain measurements

When using the time-domain option of the VNA the ‘low-pass’ or the ‘band-pass’ modes are available. The low-pass mode can only be used for equidistant sampling in the frequency domain (equidistant with respect to DC), since the Fourier transform of a repetitive sequence of pulses has a line spectrum with equidistant spacing of the lines including the frequency zero. This implies that for a given frequency range and number of data points the instrument must first work out the exact frequencies for the low-pass mode (by using the soft key: set frequency Low-pass). Once these frequencies are defined, calibration can be applied. For a linear time-invariant system, frequency- and time-domain measurements are basically completely equivalent (excepting signal-to-noise ratio issues) and may be translated mutually via the Fourier transform. Note that the Fourier transform of a spectrum with constant density over a given frequency range (rectangular spectrum) has a $\sin(f)/f$ characteristic in the time domain. This characteristic shows undesired ‘side-lobes’ and thus an (amplitude) weighting function (= window) is applied in the frequency domain before entering the FFT. This weighting function is typically \sin^2 or Gaussian and helps to strongly suppress side-lobes in the time domain. Within the low-pass mode we can use the pulse and step function. The step function is nothing other than the integral over the pulse response. When using the gating function, keep in mind that gating is a non-linear operation and thus may artificially generate frequency components that were not present before gating. In the band-pass mode the spectral lines (frequency-domain data points) need no longer be equidistant with respect to DC but just within the frequency range of interest. The corresponding time-domain response for the same bandwidth is twice as long as in the low-pass mode and we get, in general, complex signals in the time domain. These complex signals are equivalent to the I and Q signals (I being in phase and Q quadrature) often found in complex mixer terminology. They can be directly displayed using soft keys ‘real’ or ‘imaginary’ in the format menu. The real part is equivalent to what one would see on a fast scope, i.e. an RF signal with a Gaussian envelope. The meaning of the time-domain band-pass mode response in linear magnitude format is the ‘modulus of the complex envelope $[\text{SQRT}\{(\text{re}^2(t)+\text{imag}^2(t))\}]$ of a carrier modulated signal’. Note that the time-domain mode can also be applied for CW excitation from the VNA, but to analyse a slowly time-variant response of the DUT (up to the IF bandwidth of 3 kHz).

To start with the time-domain option, follow these instructions.

- Preset the instrument, dial a frequency range of 300 kHz to 3 GHz (801 data points) and go into the time-domain, low-pass mode, step function. By this operation the VNA sets a frequency range, which is required for the low-pass mode Fourier-transform calculation. Check the frequency reading.

- Now you have to calibrate S_{11} as you did before, with the only difference being that you have to use the ‘OPEN’ from the calibration kit, as you are now measuring up to about 3 GHz. (Refer to the above descriptions of calibrating S_{11} if you do not remember.)
- Read out the pulse amplitude with the end connector of the cable open.
- Now connect a SHORT and read the signal.
- Discuss the meaning of the sign from the read-out for the reflected wave.
- Connect the 10 Ω test box with a SHORT at the end to the THRU.

Can you calculate the resistance from the read-out?

Remember the definition of S_{11} and which simple formula you have to apply: $\rho = (Z - Z_C)/(Z + Z_C)$; this is the reflection coefficient as seen in the reference plane of the DUT. Z_C stands for the characteristic impedance of the cable and usually amounts to 50 Ω . *Note that Γ and r can also be used to denote the reflection coefficient.*

- Repeat the experiment with the 100 Ω blue box using a SHORT at the end.

Look at the 12(18) pF and 100 pF capacitor boxes (end = open). Discuss the traces and remember that it is a single-step response. The two numbers (12 and 18) for the capacitance indicate that a 12 pF capacitor mounted inside the test box returns a total capacity of 18 pF because of the connector feed-throughs and other parasitic capacitances.
- Now apply a 25 Ω DUT in the calibrated reference plane instead of the capacitor, using two 50 Ω loads connected in parallel via a coaxial T-piece.
- Use an appropriate vertical scale factor to obtain a good resolution on the screen.
- Store the result in the memory and display memory and data.
- Put a SHORT on the end of the calibrated cable (instead of the previously used 25 Ω) and compare the data trace with the memory trace.
- Plot the results (if a plotter is available) and discuss them.
- You may also try building a simple notch filter by attaching the T-piece to your calibrated reference plane and a 50 Ω load to one connector of the T-piece and an open (and later shorted) stub (say about 1 m of cable) to the other connector of the T.
- If you are interested in going further, repeat all the time-domain measurements mentioned above in the pulse mode (low-pass) instead of the step mode.

There is a wide range of applications for this synthetic pulse time-domain technique. A VNA in the time-domain, low-pass step mode has a very similar range of applications to a sampling scope (Figs. 1 and 2). However, it must always be kept in mind that carrying out a measurement in the frequency domain and then going via FFT or similar into the time domain implies strict linearity of the DUT. Thus a transient on a non-linear system, such as the onset of oscillations on some microwave oscillator with active elements after turn-on of the supply voltage, would not return meaningful results when using the technique mentioned above. The dynamic range of a typical sampling scope is limited at about 60 dB with a maximum input signal of 1 V and a noise floor around 1 mV. The VNA can easily go beyond 100 dB for the same maximum level of the input signal of about +10 dBm. Both instruments are using basically the same kind of detector, either a balanced mixer (four diodes) or the sampling head (two or four diodes), but the essential difference is the noise floor and the average power arriving at the receiver. In the case of the VNA we have a CW signal with bandwidth of a few hertz and thus can obtain with appropriate filtering a very good signal-to-noise ratio, since the thermal noise floor is -174 dBm/Hz. For the sampling scope we get a short pulse with a rather low repetition rate (typically around 100 kHz) and all the energy is spread over the full frequency range (typically 20 GHz bandwidth). With this low average power (around 1 μ W) the spectral density is orders of

magnitude lower than in the case of the VNA and this ultimately makes the large difference in dynamic range (even without gain switching). Furthermore, the VNA permits a wide range of band-limited RF pulses to be tailored in the band pass, which would be very tedious with a sampling scope.

2.5 Amplifier measurements (second VNA lesson)

The description here is for the HP8753C. If you use another instrument you may have to adapt certain measurement parameters.

Whenever you measure medium- or high-power amplifiers, be sure that the power level cannot destroy the input of the VNA. For example, even measuring the input impedance of an amplifier may destroy the VNA, if the amplifier produces parasitic (self) oscillations.

- Preset the VNA. To protect the VNA against overload from the amplifier output, start with the following set-ups:

Output power: –10 dBm

Attenuator port 1: 20 dB; this leads to an input power of –30 dBm for the amplifier.

Never remove the fixed 30 dB attenuator from the output of the amplifier. Assume that its attenuation is exactly 30.0 dB, constant over the complete frequency range.

- Carry out a response calibration in S_{21} .
- Measure the transmission coefficient from port 1 to port 2 ==> S_{21} (B/R) not S_{12} !
- Display the response using the auto-scale function; select trace averaging with an averaging factor of 10; measure the response (in this case the isolation) at 1 GHz; produce a hard copy (plotter) of the screen.
- Reduce the IF bandwidth (IF-BW) to 100 Hz; compare the result with the above hard copy.
- Go back to IF-BW = 3000 Hz and connect or turn on the DC voltage of 15 V to the amplifier. Use the full-screen display for a single channel; determine the small signal gain of the amplifier at 1 GHz; with marker1. Produce a hard copy.
- Measure the 3 dB bandwidth (and also 1 dB BW) of the amplifier (determine the frequencies where the gain is 1 dB lower than at 1 GHz) using marker functions. Use two markers and the statistic function. Produce a hard copy!

Test other possibilities to determine the 1 dB and 3 dB bandwidth of this amplifier:

- Determine the 1 dB compression point of the amplifier at 1 GHz. Go via MENU to CW frequency and power sweep. Use marker functions. Produce a hard copy!
- Measure the frequency range over which the gain compression of 1 dB occurs, first in the specified frequency range of the amplifier using the CW mode at the lower, mid-band and upper frequency of the amplifier. Apply the display features of the 8753 for this measurement and the marker search function. Produce a hard copy!
- Now return to the frequency sweep display, set a suitable range (response calibration) to cover the 3 dB bandwidth of the amplifier and store the response trace for a small input signal (small enough that the output power is well below the 1 dB compression point). Select a vertical resolution of 1 dB/div and an adequate reference level to position the trace approximately in the middle of the screen.
- Now increase the input power until the read-out is approximately 1 dB below the previous trace. Note that the 1 dB compression level is frequency dependent. Copy the result.

What we see on a network analyzer

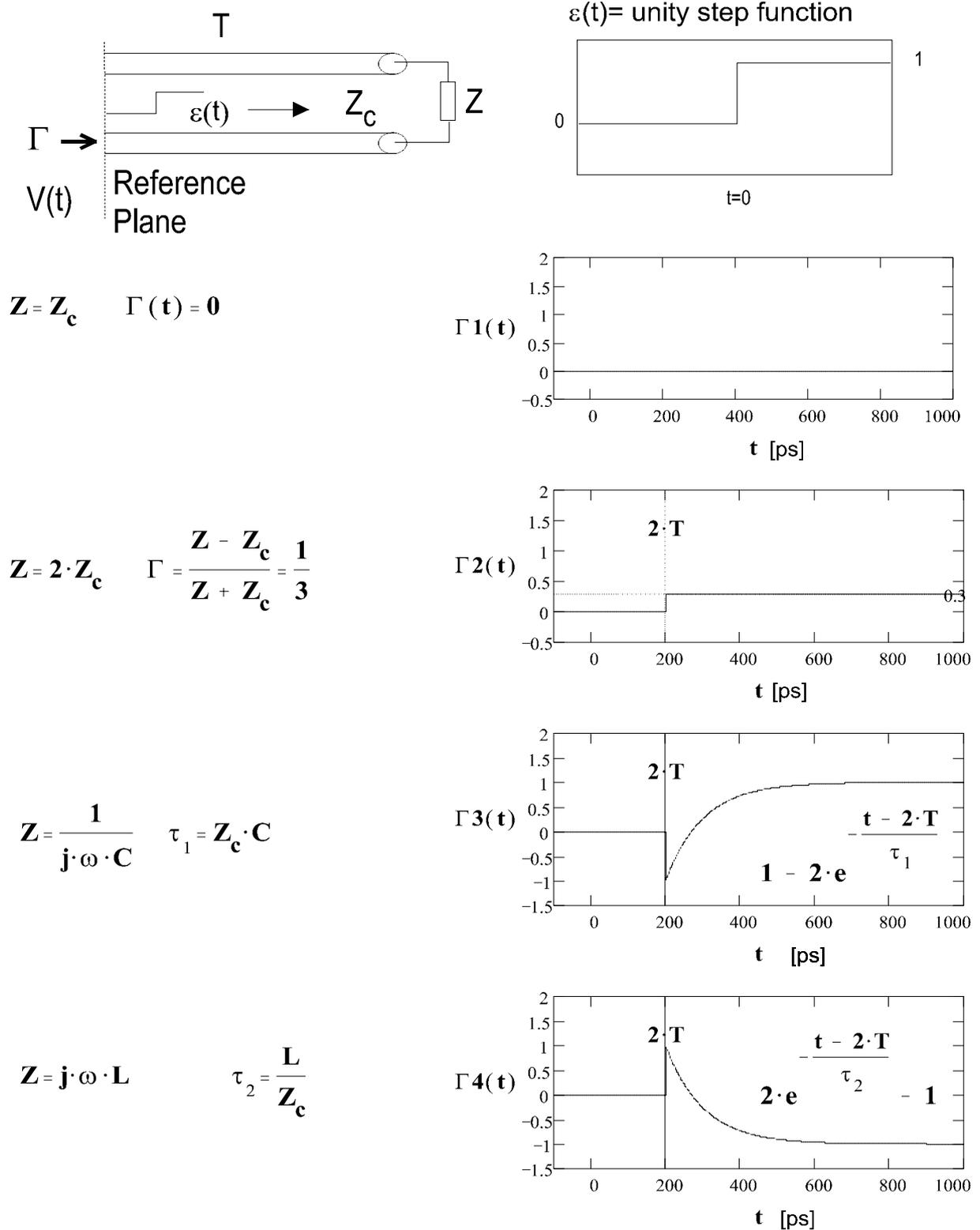
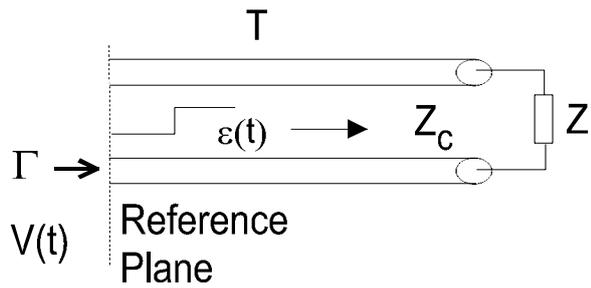


Fig. 1: Step-function response for different terminations on a VNA

What we see on a sampling scope



$$Z = Z_c \quad \Gamma(t) = 0$$

$$V1(t) = \varepsilon(t)$$

$$Z = 2 \cdot Z_c \quad \Gamma = \frac{Z - Z_c}{Z + Z_c} = \frac{1}{3}$$

$$V2(t) = \varepsilon(t) + \Gamma \cdot \varepsilon(t - 2 \cdot T)$$

$$Z = \frac{1}{j \cdot \omega \cdot C} \quad \tau_1 = Z_c \cdot C$$

$$Z = j \cdot \omega \cdot L \quad \tau_2 = \frac{L}{Z_c}$$

$\varepsilon(t) =$ unity step function

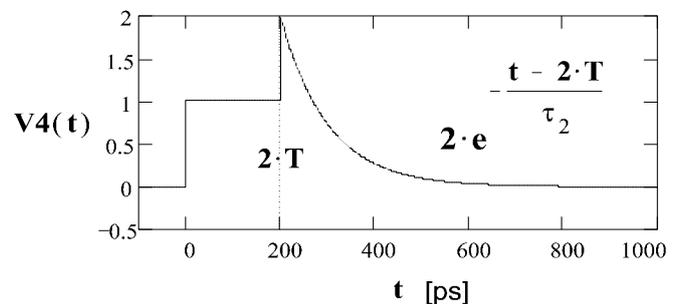
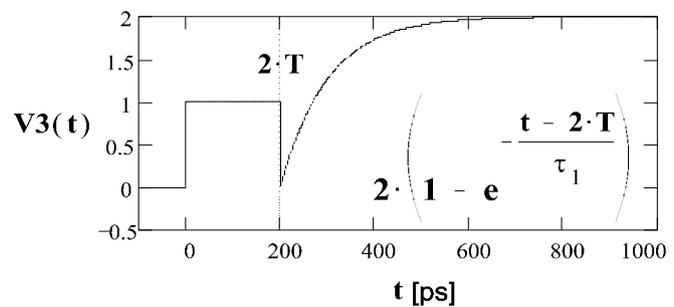
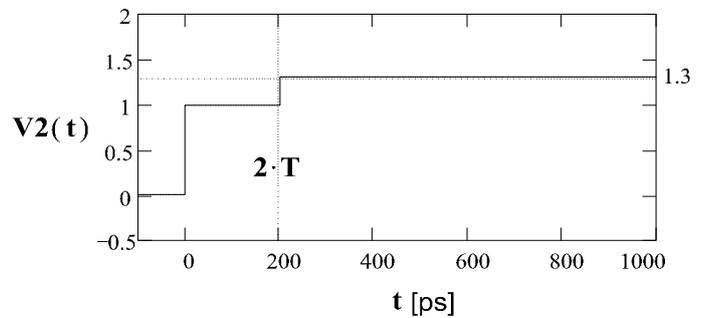
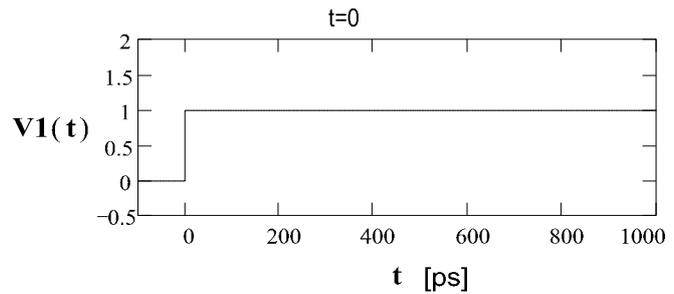
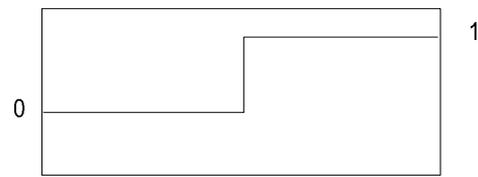


Fig. 2: Step-function response for different terminations on a sampling scope (through sampler)

Preset the VNA, set the power and attenuator of port 1 as at the beginning, set the frequency range and carry out an S_{11} one PORT calibration. Measure the Standing Wave Ratio (SWR) of the input of the amplifier in the specified frequency range. Determine the maximum SWR or Voltage Standing Wave ratio (VSWR).

Measure the deviation from linear phase and the group delay in transmission.

2.6 Directional couplers and cavities (third VNA lesson)

Now you have the choice of carrying out measurements on directional couplers, on cavities, or on a transmission line over a ground-plane (with an EMC probe).

In the directional-coupler section you will become familiar with the working principle of loop couplers, how to perform a precision calibration and how to measure the relevant parameters of different couplers.

Cavity measurements include different ways to determine the loaded and unloaded Q of a cavity; tuning the coupling loop to critical coupling; measurements of the resonance frequencies of a pillbox for different modes in comparison with MAFIA calculations, and perturbation measurements.

With the EMC probe you can measure the effectiveness of RF shielding and the standing waves on a wire over a ground-plane. Decide yourself whether you will do a little of each or focus on your particular points of interest.

2.6.1 Directional couplers

The principle of a directional loop coupler is very simple. The capacitive coupling and the inductive coupling of the loop should be tuned to have exactly the same magnitude at the outputs of the coupling loop, terminated with 50 Ω .

From the capacitive part of the coupling we get equal output voltage at either end of the loop. The induced current, however, flows through the loop and leads to a positive voltage (with respect to ground at this location) at one end and a negative voltage at the other (Fig. 3).

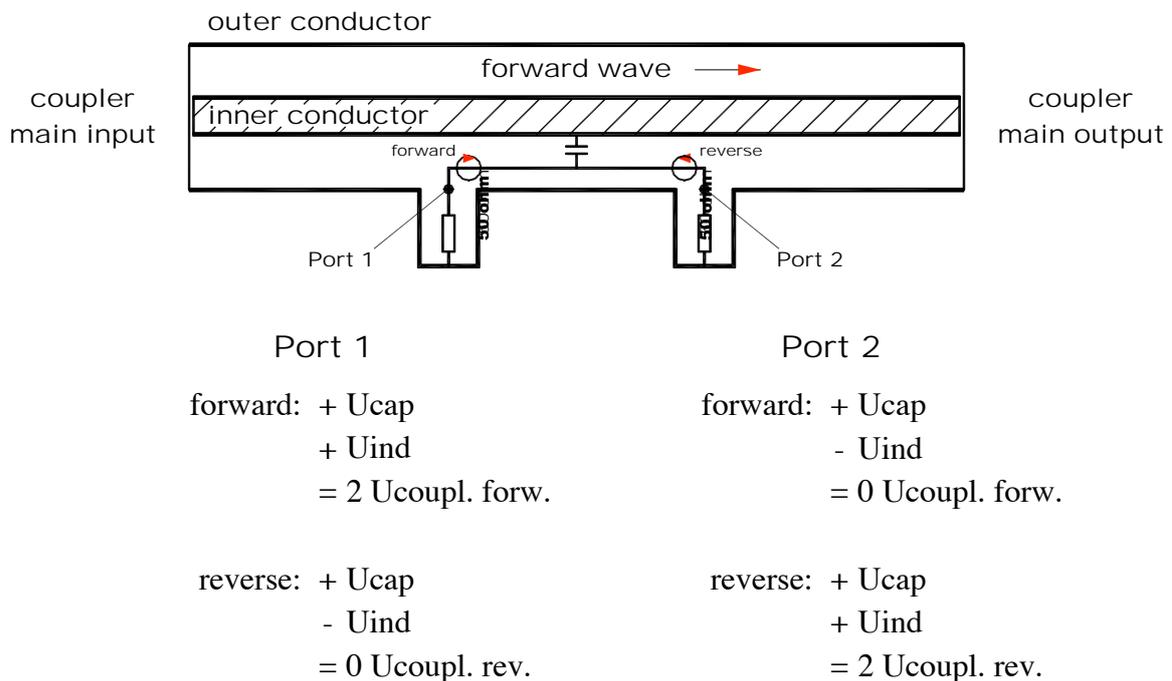


Fig. 3: Construction principle of a directional coupler

At the positive end (port 1), capacitively coupled and induced voltages combine, while on port 2 they cancel each other. This is valid for an incident (from the coupler main line input) wave in the forward direction.

Let us say that port 1 of the coupler then measures a proportional part of a wave in the forward direction, while port 2 does not see it. For a wave in the reverse direction it is just the opposite, as the

induced current in the loop flows in the opposite direction to the capacitively coupled part. Thus a backward-moving wave is proportionally measured on port 2, while port 1 does not see it and the directional coupler is ready.

The directivity is a measure of how close capacitive and inductive coupling match in magnitude and also in phase (or 180 degrees offset). For perfect balance the directivity is infinite. In this case you measure only the forward wave on port 1 and only the reflected wave on port 2.

In practice, you will always have a finite directivity, which implies that you measure a (small) part of the forward wave on port 2 and vice versa. The directivity of a directional coupler in your amplifiers should be better than 30 dB at the centre frequency.

Note that the 50 Ω terminations are also required to have a matching better than -30 dB as they directly influence the effective directivity!

The measurement error of the coupling coefficient should be below ± 0.1 dB.

As directional couplers of high-power amplifiers have coupling coefficients in the range from -30 to -60 dB, it is not easy to calibrate these levels to 0.1 dB to absolute accuracy, as the strongly different input levels of the VNA have an impact on the uncertainty. Therefore it is necessary to calibrate the response with a known attenuator in the vicinity of the coupling coefficient.

Thus, for example, if you do not have attenuators calibrated in accordance with the Physikalisch Technische Bundesanstalt (PTB) in Germany or the National Institute of Standards and Technology (NIST) in the US, with a certificated attenuation at the desired frequency, you can help yourself by measuring attenuators up to 10 dB directly and add them until you get close to the coupling coefficient.

2.6.1.1 Precision calibration of a 108.406 MHz coupler with -30 dB coupling coefficient

(This procedure also applies, of course, for other adjustable, narrow-band couplers at different frequencies.)

- Preset the VNA. For all precision measurements the VNA should have had a warm-up time of at least 10 minutes, but two hours is better.
- Set the CW frequency to the centre frequency of the coupler.
- Disconnect the short cables from ports A and B of the analyser (not from the S-parameter test set).

This procedure allows simultaneous reading of the coupling AND the directivity when changing the orientation of the coupling loop. It is applicable to VNAs where the connections between the S-parameter test and the VNA unit are accessible. However, there also exist instruments that have two receiver input ports and thus allow true three-port measurements.

- Select input ports A/R for CH1 (channel 1) and B/R for CH2 (channel 2). (You have to monitor coupling and directivity at the same time.)
- Display: Dual channel on, split screen off.

The following section on the calibration of attenuators may be skipped or postponed if time is short.

- Do a response calibration for CH1 (A/R), Cal Type N 50 Ω , with port 1 of the S-parameter test set connected directly to input A of the VNA unit.
- Now insert (only one at a time) the three 10 dB attenuators into this signal path. Measure the attenuation of the three 10 dB attenuators and add the measured values numerically.
- Measure the attenuation of all three attenuators mounted together in the same signal path.

- Measure the attenuation of all three attenuators together with 0 dBm and +10 dBm output power. You will notice minor changes in the readings as a function of the generator power. Try to explain that effect.
- Repeat the last four steps, but now using channel CH2 (B/R).
- Average the result of the two channels (i.e. find the mean value between both channels). The purpose of this procedure is to get the maximum precision possible.
- Measure the loss between coupler input and output (main line) and note the value.
- Connect all three attenuators to the output of the coupler.
- Menu power: +20 dBm.
- Calibrate CH1 and CH2 (response) with the coupler and the three attenuators inserted. You may wish to use the average function for highest precision.
- Which reading from the VNA do you expect to give exactly -30.0 dB on your DUT?

Now we are ready for the actual tuning of the coupler.

- Connect the VNA input A to port 1 of one of the two coupling loops of the coupler.
- Connect the VNA input B to port 2 of the same coupling loop.
- Remove the three 10 dB attenuators at the output (main line of the coupler) and terminate this output with the 50Ω load from the calibration kit.
- Turn the average off in both channels and set the REFERENCE POSITION to 9 vertical units.
- Now rotate slowly the coupling loop until you get the desired reading of _____ dB at CH1.
- The reading on channel CH2 gives the directivity. (It is not worth correcting the small difference of the reference to -30.0 dB.)
- Do the same for the other coupling loop.
- Reconnect input A to the S-parameter test set.
- Calibrate S_{11} in the frequency range 20 MHz to 220 MHz (801 points).
- Measure the input matching of the coupler at 108.4 MHz using the termination resistor of the calibration kit at the output. Write down the matching in dB. How is the matching at other frequencies?
- Measure the coupling from 20 MHz to 220 MHz; note that the coupling is increasing by approximately 6 dB per octave. Can you explain why? Measure the remaining couplers (if time allows).

2.6.2 Cavities

We have three cavities available in this course: two pillbox cavities (cylindrical resonators) and one coaxial cavity.

The first pillbox cavity has a diameter of 30 cm; the second has a diameter of 31.5 cm, and its length is variable. The coaxial resonator is short-circuited at one end, while the capacitively loaded open end is variable at the location of the capacitor plates. This resonator was built to determine the dielectric losses of isolators.

2.6.1.2 Pillbox measurements

- PRESET the instrument. Set the frequency to between 500 MHz and 1.2 GHz. (There is no resonance below 500 MHz and there is no time available here to look for higher modes above 1.2 GHz.)

- Set the number of points to 1601 (to have enough data points at a high Q-resonance).
- Choose two small probes with N-connectors, look at how they are constructed.
- Calibrate (N) S_{11} on CH1 and S_{21} on CH2.
- Set the reference position of both channels to 9 vertical units.
- Connect the cables to the two installed inductive or capacitive probes and determine the frequency of all resonances you can find at S_{21} in the selected frequency range. (Display only CH2.)
- Calculate the length (actual length for pillbox 2) using the given mode patterns in your handouts and determine the type of the modes. Compare with results obtained using Figs. 4 and 5.
- Set the marker on the peak of the E_{010} mode (TM_{010}) and adjust it to the centre of the screen (MARKER => CENTER). For pillbox 2 (with variable length), the E_{010} mode is that which does not change its frequency if the length of the cavity is altered.
- Reduce the SCALE/DIV stepwise to 0.5 dB/DIV and the SPAN to fill the screen with the resonance curve. You may use the marker functions MARKER SEARCH MAX, MARKER => CENTER and MARKER => REFERENCE to do this.
- Then press MKR; MKR ZERO to define the maximum of the resonance curve with, as reference, the marker read-out about zero.

Now you can use the width function to automatically get the -3 dB points. (Marker search, width on, width value -3 dB. You measure the nearly unloaded Q if both of the used probes are so small [noise level] that you get the result only by setting the following values: output power to 20 dBm, IF bandwidth to about 300 Hz, and the averaging on. It is also useful to insert a low-noise amplifier in front of port 2.)

- Write down the centre frequency, 3 dB bandwidth and the measured Q.
- Now try to reach a critical coupling with a bigger loop or a capacitive pick-up. Use channel 1 with increased bandwidth (as the new loop may shift the resonance frequency). Go to 10 dB/DIV and display S_{11} in the formats LOG MAG and Smith Chart. Adjust a circle that goes through point 1 (50Ω). Turn the circle with the function PHASE OFFSET to the left side of the screen until it is symmetric to the locus of real impedance (de-tuned short position). Measure the loaded and unloaded Q using the markers. The points where the real and imaginary parts are equal give the bandwidth for the unloaded Q. (See *RF cavity higher order mode measurements* by J. Byrd and R. Rimmer in the appendix given to students participating in this tutorial and *Basic concepts I and II* by Heino Henke in this Report). You can find these points in the de-tuned short position looking at the real and imaginary parts of the marker.

The loaded Q can be found at the crossing point of the circle with ± 45 degree lines starting at the zero point. This can be easily done on paper but not on the analyser screen. It helps to know that the loaded points are always (not only at critical coupling) the highest and the lowest points of the circle, if this is brought to the de-tuned short position. For critical coupling you will find the points 10Ω (or 0.2 when normalized to 50Ω) for the real part and $\pm 20 \Omega$ for the imaginary part.

But there is an even easier method for reading out the loaded and unloaded Q, if you turn the circle with the phase offset (do not use the electrical delay as it would deform the circle) to the de-tuned open position. For critical coupling the circle lies directly on the circle of the Smith Chart, where the (normalized) real part is 1. Therefore you find the points of the unloaded Q at the crossing to the lines where the (normalized) imaginary part is 1 too ($X = R$) and the loaded Q at the crossing to the lines where the (normalized) imaginary part is 2 ($X = R + 1$).

Make sure that the de-tuned short and de-tuned open positions are well adjusted by checking for symmetry of the maxima of the imaginary parts using appropriate marker functions.

- Determine the loaded Q also by the -3 dB points of S_{11} for the critical coupling in the format LOG MAG.
- Determine the loaded Q in transmission by the 3 dB points of S_{21} with one probe in critical coupling and the other very small.
- Move your coupling loop to over-critical and under-critical coupling and calculate the coupling coefficient from the formula $k = 1/(2/D - 1)$. D is the diameter of the circle; its unit is the radius of the Smith Chart. If $D = 1$, then the coupling is 1 and you have critical coupling with the centre frequency point at the centre of the Smith Chart. Weakly coupled resonators have small Q-circles, strongly coupled have large ones.

MODE LATTICE FOR CYLINDER RESONATORS

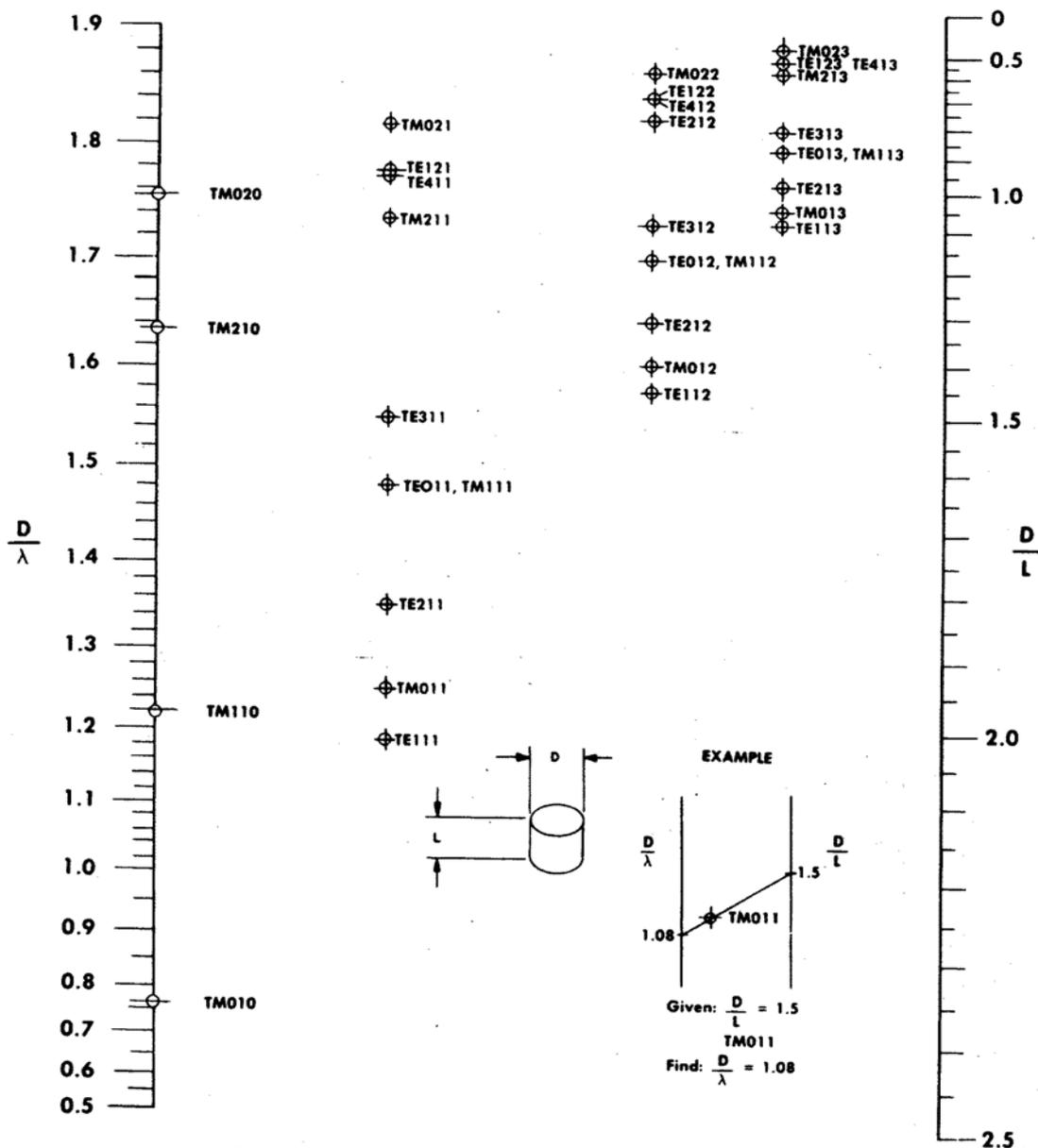


Fig. 4: Mode lattice for cylindrical resonators (reprinted from R.N. Bracewell, 'Charts for resonant frequencies of cavities', Proc. IRE, August 1947).

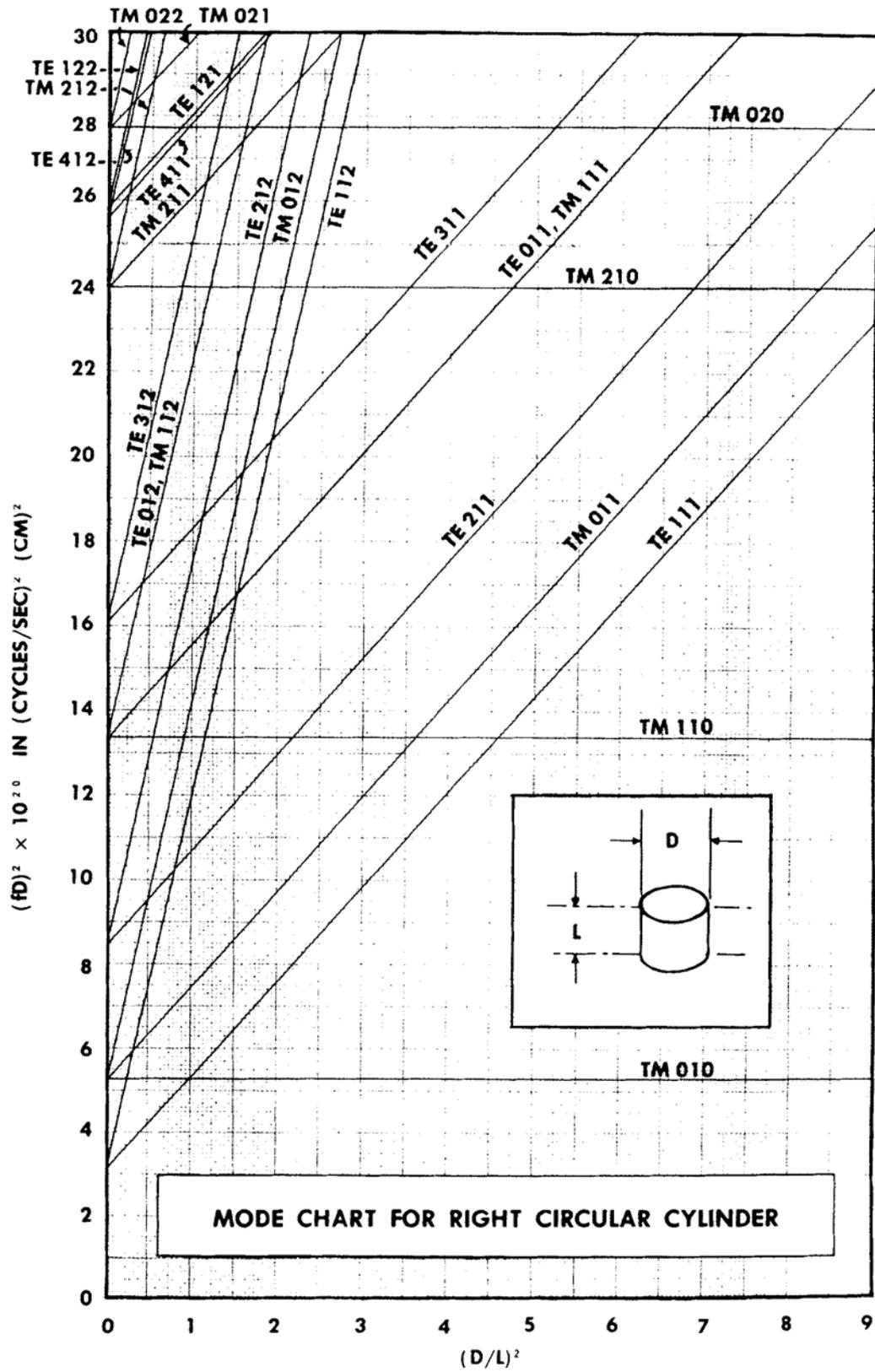


Fig. 5: Mode chart for a pillbox-type cavity (reprinted from *Microwave engineers handbook, Vol. 1*).

3. EXPERIMENTS WITH THE SPECTRUM ANALYSER

3.1 Becoming familiar with the spectrum analyser

- Display the spectrum of the signal coming from the ‘CAL’ output of the spectrum analyser (SPA).
- Display the spectrum of RF signals present in the classroom (using a short wire as an antenna).
- Measure the spectrum of an output signal from a signal generator (CW mode, no modulation) and look for second and third harmonics. How can you discriminate against SPA input mixer-related harmonics? (Do not exceed 0 dBm generator output power.)
- Measure the frequency response of an amplifier (scalar network analyser mode); if no tracking generator is available, measure 10 frequency points (100 MHz – 1 GHz) manually.
- Measure the 1 dB compression point (i.e. small signal gain reduction by 1 dB due to the beginning of saturation effects of the amplifier under test) at three different frequencies (low, mid-band, high).
- Measure the second-order intercept point (non-linear products at sum and difference frequency of the two input signals; both input signals (= tones) should have equal amplitude; select suitable frequencies for input signals in order to be able to display the sum and difference frequencies). Measure at three different amplitude levels; watch out for second and third order generator harmonics; you may use low-pass filters at the input.
- Measure the Third-Order Intercept (TOI) point (use two frequencies about 50 MHz apart). The IM3 products appear separated by the frequency difference from each tone. Use the automatic function (meas/user) button and TOI on (if available).

3.2 Noise and noise-figure measurements

- Make sure you are familiar with the most important functions of the SPA (frequency setting, Resolution Bandwidth, RBW, Video Bandwidth, VBW, amplitude scale). Note that the spectrum analyser should be used in the sample mode and not the usual peak detector mode.
- With resolution BW = 1 MHz, Start = 10 MHz, Stop = 1000 MHz, Video BW = 100 Hz, input attenuator = 0 dB display the baseline and read the power. How many dB is it above the thermal noise floor (thermal noise at 290 K = -174 dBm/Hz)?
- For the same settings, now connect the solid-state noise source (to be powered with +28 V DC via rear BNC connector) to the SPA input. The Excess Noise Ratio (ENR) of this device is close to 16 dB or a factor of 40 in spectral power density greater than the thermal noise of a common 50 Ω load. Use the table below to correct for absolute power reading. Note that for absolute power measurements with the spectrum analyser close to the noise floor the reading is too high by the amount indicated in the right column. The analyser should be set for this measurement in ‘sample’ mode and not in ‘peak hold’ mode, which may be a default setting.
- Load the noise measurement option software to the SPA (if available). Otherwise skip the next five points.
- Calibrate the SPA with the preamplifier. Record the measured noise figure of the system (SPA + preamplifier) from the reading on the CRT after calibration.
- Measure the gain and noise figure of some amplifiers. Convert noise figure into noise measure.
- Measure the gain and noise figure of some attenuators.
- Measure the noise figure of two amplifiers in cascade by the method already described. Measure the noise figure of an attenuator in the same way.

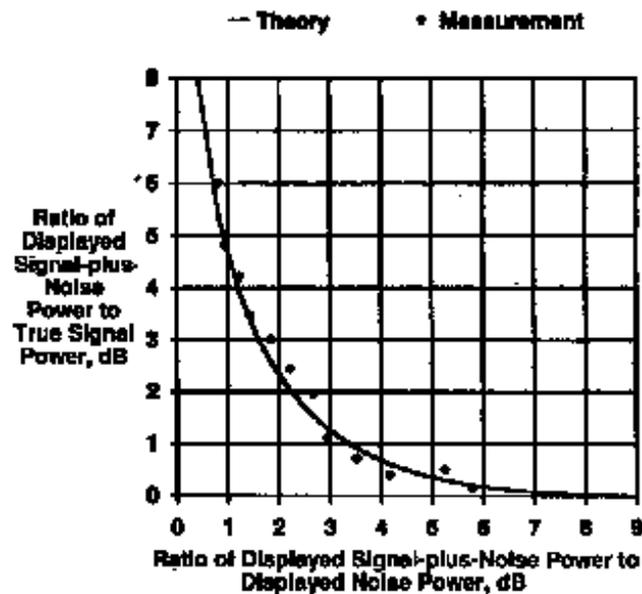


Fig. 6: Corrections for absolute noise power measurements on a spectrum analyser close to the instrument noise floor (from: A. Moulthrop, M. Muha, 'Accurate measurements of signals close to the noise floor on a spectrum analyser', IEEE Transactions MTT, Vol. 39, No. 11, 1991, pp. 1882–1884).

- Connect to the input of the preamplifier with a coaxial cable of 1–2 m length terminated by a short or open. Discuss the results observed (frequency range 10 MHz – 1 GHz). Also use a 50 Ω load or a triple stub tuner. Try to tune the 50 Ω input termination into an optimum noise source match using the triple stub tuner.
- If the noise measurement option is not available, connect a preamplifier to the SPA (input attenuator = 0 dB) and record the two traces 'noise source on' and 'noise source off'. Calculate from those traces the noise figure of the DUT. Convert noise figure into noise measure.

3.2.1 Some useful equations for noise-figure evaluation

There is frequently confusion over how to handle the dB (deci-Bel). The dB is used to describe a power ratio and thus is a dimensionless unit. As the power dissipated in a resistor is proportional to the square of the voltage or the square of the current, one may also take the ratio of these quantities into account. The dB is also used to describe absolute signal levels, but then there must be an additional letter to indicate which reference one refers to, e.g. +10 dBm (= 10 mW) is a power level of 10 dB above 1 mW (+20 dBm = 100 mW).

$$\alpha[dB] = 10 \cdot \log\left(\frac{P_1}{P_2}\right) = 20 \cdot \log\left(\frac{V_1}{V_2}\right)$$

$$10^{\frac{\alpha[dB]}{10}} = \frac{P_1}{P_2} \qquad 10^{\frac{\alpha[dB]}{20}} = \frac{V_1}{V_2}$$

The terms 'noise figure' and 'noise factor' are used to describe the noise properties of amplifiers. The term F is defined as signal-to-noise (power) ratio at the input of the DUT versus signal-to-noise power ratio at the output. F is always >1 for linear networks, i.e. the signal-to-noise ratio at the output of a two-port or four-pole is always more or less degraded. In other words, the DUT (which may be also an amplifier with a gain smaller than unity, i.e. an attenuator) always adds some of its own noise to the signal.

F [dB] is called the noise figure,

F [linear units of power ratio] is sometimes called noise factor,

F [dB] = 10 log F [linear units].

$$F[\text{linearunit}] = \frac{\text{ENR}[\text{linearunit}]}{Y[\text{linearunit}] - 1} = \frac{T_{\text{ex}}}{T_0 \cdot (Y - 1)} \quad \text{with } T_{\text{ex}} = T_H - T_0 .$$

ENR is the excess noise ratio delivered by the noise diode and tells us how much ‘warmer’ than room temperature the noise diode appears. For an ENR of 16 dB this amounts to roughly a factor of 40 in power, or $40 \times 300 = 12\,000$ K.

The quantity Y is the ratio of noise power densities measured on the SPA between the settings: ‘noise source on’ and ‘noise source off’.

As shown in the equations below, the gain of the DUT can also be found from the two readings on the SPA. Thus one can simultaneously measure gain and noise figure.

The technique for noise-figure measurement described above is commonly used for noise-figure evaluation of an amplifier in the RF and microwave range for frequencies higher than about 10 MHz. For lower frequencies the characterization of noise properties is normally done in terms of specifying the (input) noise voltage and the (input) noise current of some amplifier, by placing a short between the input terminals or leaving the input open, respectively. Obviously one can define an optimum generator impedance (noise match) for which the combined effect of voltage and current noise is at minimum. This noise voltage and current may also be converted into an equivalent noise temperature. The noise temperature of an electronic amplifier at room temperature may be surprisingly low (depending on the technology applied) in the range 1 kHz to about 10 MHz: values below 1 K have been reported. This is not a contradiction to basic thermodynamic concepts, since a device like an amplifier (or also a forward-biased diode) connected to a power supply is no longer in thermodynamical equilibrium and may show equivalent noise temperatures well below its physical temperature. For the frequency range 100 MHz to about 10 GHz noise figures of 1 dB (= 70 K) and better are available over an octave bandwidth for non-cooled amplifiers (e.g. the low noise module of 12 GHz satellite receivers has a noise figure of around 1 dB and the antenna, looking into cold (3 K) space, has a noise temperature of about 30 K due to the losses of the atmosphere).

$$Y = \frac{\text{measured DUT output power (density) with noise source = hot}}{\text{measured DUT output power (density) with noise source = cold}}$$

$$\text{ENR [linear unit]} = \frac{(T_H - T_0)}{T_0}$$

$$\text{ENR [dB]} = 10 \cdot \log \left(\frac{(T_H - T_0)}{T_0} \right)$$

$$G_{(DUT)} [\text{lin}] = \frac{N(\text{SPA} + \text{DUT}, \text{Diode on}) [\text{lin}] - N(\text{SPA} + \text{DUT}, \text{Diode off}) [\text{lin}]}{N(\text{SPA}, \text{Diode on}) [\text{lin}] - N(\text{SPA}, \text{Diode off}) [\text{lin}]}$$

N = noise power measured on the SPA for, e.g., 1 MHz resolution bandwidth

$$F_{\text{total}} [\text{linear units}] = F_1 [\text{linear units}] + \frac{F_2 [\text{linear units}] - 1}{G_1 [\text{linear units}]} + \dots$$

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There is a large amount of very useful (and free!) information available on the Web, such as application notes by instrument manufacturers (Anritsu, Agilent (Hewlett-Packard), Rhode & Schwarz, Marconi, IFR, Tektronix, and many others). These application notes are usually very easy to read and give a lot of practical hints and examples after a brief theoretical introduction.

Apart from these application notes a few books are listed below that may also be helpful to gain a deeper insight to all questions of RF spectrum, network and noise measurements. Since they are not quoted in the text, no reference numbers are added (references for the figures are cited in full in the figure caption).

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