Teaching Kit
R&S®FPC-Z10
User Guide

1179.3150.02-01

ROHDE&SCHWARZ

This document describes the following product:

- RS®FPC-Z10 Teaching kit (order number 1328.7338.02)

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1178.8436.02 | Version 01 | R&S®FPC-Z10

Throughout this user's guide, products from Rohde & Schwarz are indicated without the ® symbol, e.g. R&S®FPC-Z10 is indicated as R&SFPC-Z10.

PAD-T-M: 3576.8258.02/03.00/CI/1/EN

Contents

1 Installation and safety instructions

The R&S® FPC-Z10 Teaching kit (1328.7338.02) is a DUT to showcase different RF measurements in a lab enviroment.You can use it to demonstrate the functionality of a spectrum analyser, network analyser or signal generator. Examples for measurements are provide in this document.

1.1 Safety information and explanation of safety labels

The teaching kit is safe to use when you use it as intended and within its perfomance limits. The limits are described in the product documentation. (data sheet and manual)

However, electrically powered products still have some risks, such as electric shock, fire or a personal injury. Take the following measures for your safety.

- Use a power supply that is limited to 5 V DC at 1 A and complies to IEC/UL/EN 60950-1 or IEC/UL/EN 62368-1.
- Make sure that RF signal you apply have a maximum level 10 dBm and the voltage 0 V DC. Connected equipment must also comply to IEC/UL/EN 60950-1 or IEC/UL/EN 62368-1.
- The baseband input and output connectors support the maximum voltage +1 V / -1 V.

If any part of the product is damaged or broken, stop using the product immediately. Never open the casing of the product.

Potential risk.

Read the product documentation to avoid personal injury or prouduct damage.

Electrostatic discharge.

Elektrostatic discharge (ESD) can damage the electronic components of the device. Use the wrist strap and cord and connect yourself to the ground to prevent ESD. Alternatively, use a conductive floor mat and heel strap conbination.

Disposal of electronic equipment.

Dispose of electronic equipment separately instead of the unsorted municipal waste. Contact your Rohde & Schwarz customer service center for enviromentally responsible disposal of the product.

Operating area

Equipment is intended to be used in measurement areas. Equipment is considered to be a test instrument; see EN 1326-2-1, the clause 5.2.4.101, the note 1. Therefore, the usual operation may reduce the level of immunity in certain applications. The product is sensitive to electrical discharge (ESD) due to open modules. Therefore, the operating area should be protected against ESD to avoid the damage of electronic components. The product has to be operated in static-approved operating areas. Details are given in EN1326-2-1, the clause 5.2. The product has to be operated indoors only and has to be kept dry. The product has no case and is sensitive to moisture and humidity.

1.2 Service and maintaince

The R&S®FPC-Z10 is a replacement unit. In a case of malfunction or defects, the whole unit is replaced instead of being repaired.Contact your local Rohde & Schwarz service center for replacing of product. You can find the current addres of your representative at www.rohde-schwarz.com.

1.3 What is it

R&S®FPC-Z10 is a universal board with a transceiver, a DC/DC converter circuitry for electromagnetic interference (EMI) troubleshooting as well as an on-board calibration kit for the network analysis. The easy-to-operate teaching kit guides the students, step by step, through the exercises, and trains their practical skills to operate test and measurement instruments.

The R&S®FPC-Z10 teaching kit is ready for a teaching lab and can be combined with an R&S®FPC1500 spectrum analyzer. The R&S®FPC1500 spectrum analyzer can provide a good introduction to the regular work-flow and to tools of an RF engineer since the features of three most commonly used instruments are combined: a spectrum analyzer, a vector network analyzer and a signal generator. FPC1500 also features an easy virtual control and powerful all-in-one remote-control applications for PC and iOS/Android.

The free-of-charge remote control software R&S®InstrumentView for R&S®FPC includes the feature “Lab” that helps teachers conveniently manage, monitor, and assist students with their measurements from a central or remote location.

1.4 Who is it for

The RF teaching kit was developed to demonstrate RF measurements, acting as the device under test (DUT).

This RF teaching kit was designed:

a) For teachers and students at polytechnic colleges and universities
(in the 1., 2. and 3. year)

b) As the demo kit for dealers and partners to demonstrate capabilities of RF instruments

c) As the training kit for partners and customers

1.5 Board overview

1 - IQ modulator/demodulator

2 - Up-converter

3 - Calibrator

4 - Synthesizer and local oscillator

5 - PLL supply (DC/DC converter)

6 - SAM controller

7 - Down-converter

8 - Power supply

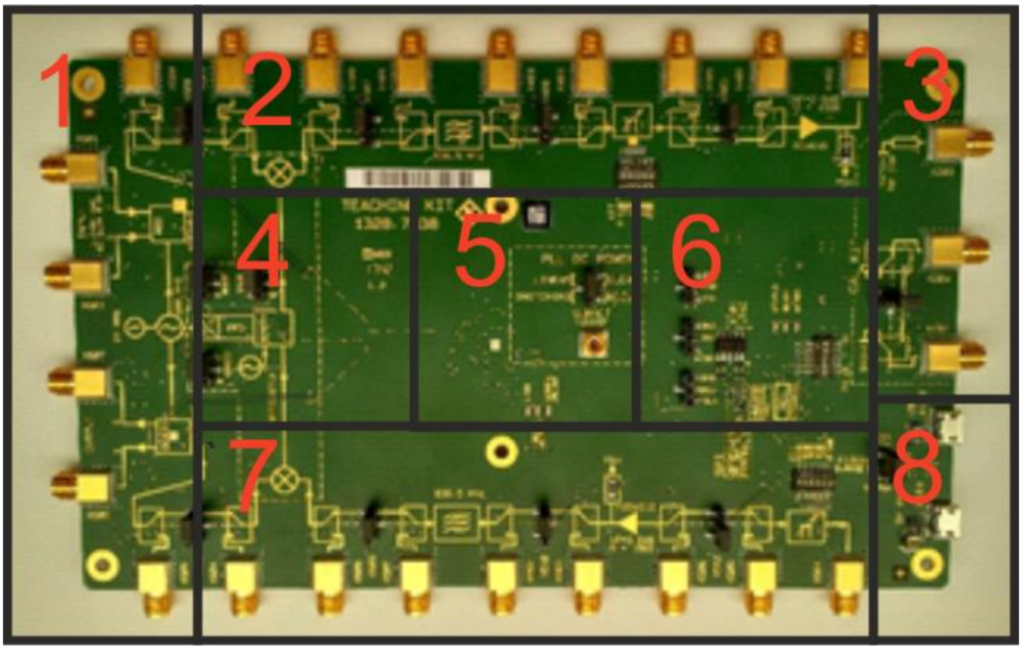

Figure 1.1 - View on the board

2 Board schematics

2.1 SAM controller

The Atmel SAM control unit (the block 6 in Figure 2.1) is the center of the control. The unit is aimed to control the synthesizer and the phase-locked loop (PLL). No matter the typical signals for programming the JTAG processor are available, the SPI interfaces (X100, X107, X108, X109) are not destined for users and are not recommended to be manipulated with.

The control processor also includes an S100 switch. PIN 1 of this switch has to be always in the ON position. For the IQ modulator operation, the PIN 7 has to be set to ON. Other PINs of this switch are not effective.

2.2 Power supply

The teaching kit has two USB interfaces to provide the power supply (see the block 8), a micro USB interface (X104) and a mini USB interface (X105). Both the interfaces can be used to supply the teaching kit with power.

A USB cable for the micro USB interface is included in the delivery. If a mini USB interface is going to be used for the power supply, a corresponding cable has to be used.

The USB interface to be used as the power supply can be selected by the X102 jumper.

Position 5VIN: Select the micro USB interface as the power supply.

Position 5VUSB: Select the mini USB interface as the power supply.

Interface power depends on an application.

- For measurements with an inactive baseband circuitry, the power is 5 V DC at 450 mA. To connect the teaching kit to another USB interface, the delivered USB cable can be used. As the power source, the USB interfaces on the R&S FPC can be exploited, for example.

- For measurements with an active baseband circuitry, the power is 5 V at 750 mA. The use of baseband connectors requires an external power supply with the power 5 V DC at 1 A. If an external power supply is required, the LED above the mini USB interface is on. This power supply is not the part of the delivered kit.

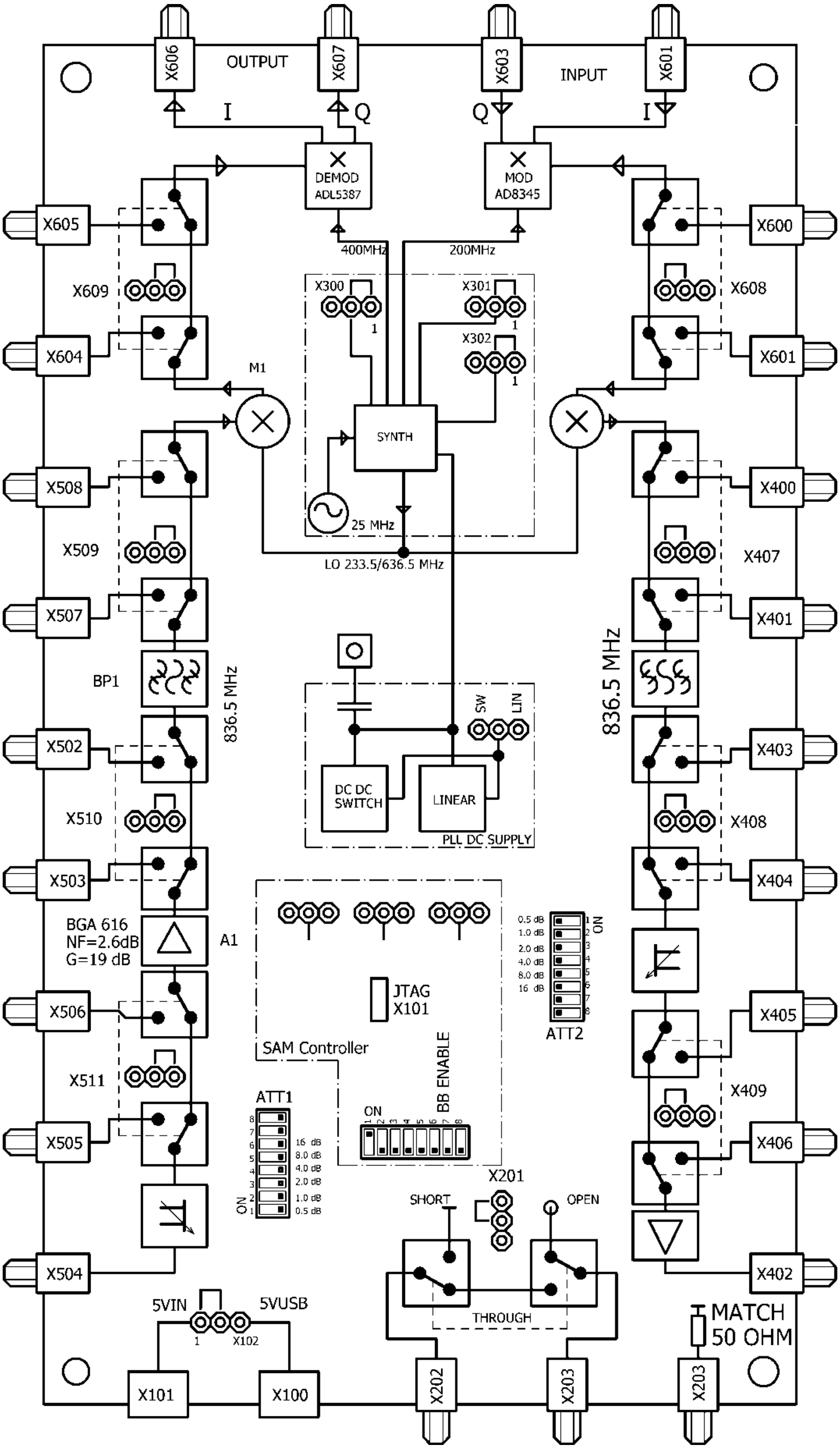

Figure 2.1 - Block diagram

2.3 Calibration kit

Calibration standards available on the R&S®FPC-Z10 allow the OSMT calibration. For the calibration, the structure with the X202, X203, X204 connectors and the X201 switch can be used (the block 3). When switching X201, the THROUGH mode (the position 1, lower) and the SHORT/OPEN mode (the position 2, upper) are toggled.

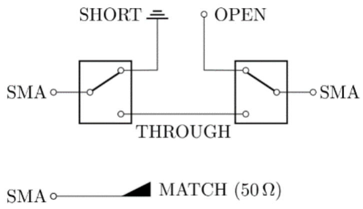

Figure 2.2 - Calibration kit

2.4 Status of the baseband part

If the application requires an I/Q modulator or demodulator (the block 1), the baseband signal path has to be activated by the DIP S100 switch. The PIN 7 controls the status of the baseband input. The BB ENABLE active state requires an external power supply as described in the section on the power supply. The position 1 has to be always in the ON state. Other positions of the DIP switch no. 8 are not supported. This user's guide does not deal with the IQ modulator more.

2.5 Frequency synthesis circuits

The frequency synthesis circuit (the block 4) is connected to a 25 MHz crystal oscillator. From this frequency, the 200 MHz and 400 MHz signals for IQ modulation and demodulation are derived by phase-locked loops (PLL). Here, the local oscillator is controlled by the X301 switch. After changing X301, the system of synthesis has to be reset by the X302 switch, or the USB power supply has to be switched off for a while and on.

The X301 switch (MODE)

- Position 1: f = 233.5 MHz
- Position 2: f = 636.5 MHz

The X302 switch (ENABLE)

- Position 1: the oscillator blocked; set after toggling to the position 2
- Position 2: operation; frequency determined by the X301 switch.

The signal of the local oscillator is delivered to mixers of the down-converter and the up-converter. The X300 switch is used for the frequency-modulation operation, and this user's guide does not deal with.

2.6 Power supply of phase-locked loop of frequency synthesis

For powering the system, two 3.3 V sources (a linear one and a pulse one) are available. The voltage sources can be toggled by the X106 switch. The LIN position is related to the linear analog stabilizer, and the SW position to the switch-mode power supply. The LIN position is the basic one. If the SW position is selected, an interfering voltage of the switch-mode power supply is superimposed on the supply voltage of PLL circuits causing the phase noise - the undesirable parasitic frequency modulation of the signal of the local oscillator. Simultaneously, this voltage is delivered to the SMP X105 coaxial connector and the DC component is filtered out by a capacitor.

2.7 Signal path of transmitter and receiver

Both paths (2,7) consist of identical components, differing only in the location of the amplifier and its orientation. Each block can be operated independently, or the blocks can be connected to a continuous path using switching jumpers.

2.7.1 Attenuator

Attenuators are passive, lossy two-port networks those exhibit a non-zero, well-defined attenuation $L \neq 0$ dB for the transmission between the input port and the output port. Attenuators are manufactured as networks with a fixed attenuation or a variable one. Attenuators usually consist of the T topology or the π topology of the network of resistors or PIN diodes.

Attenuators are used to control the power level of the signal. Thanks to the attenuators, extra strong signals can be measured, and the inter-modulation distortion can be suppressed. Tunable attenuators are the key element of loops of the automatic level control (ALC). In the measuring technology, attenuators are used for impedance matching.

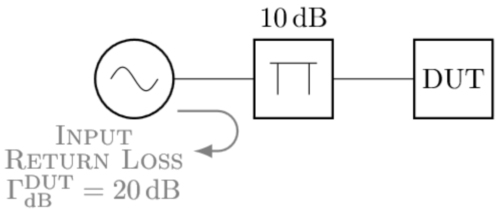

Fig 2.3 - Attenuator as a matching element

The signal from the generator passes the attenuator with the attenuation $L = 10$ dB. A part of the wave is reflected from the measured device under test (DUT). The reflected wave passes the attenuator with the attenuation $L = 10$ dB again. From the viewpoint of the generator, the DUT is very well matched.

The kit contains adjustable attenuators accessible via S401 and S500 DIP switches. Attenuations set by switch positions are summed. Adjustable attenuations are ranging from 0 dB to 31.5 dB. Attenuators are located between X404 and X405 (the transmitter) and X504 and X505 (the receiver).

2.7.2 Bandpass filter

Each branch contains one identical bandpass filter, tuned to the frequency 835.5 MHz with a bandwidth of 20 MHz. The filters are located between X401 and X403, or between X502 and X507.

2.7.3 Broadband amplifier

Two broadband amplifiers, equipped with BGA 616, are located between X402 and X406, and between X506 and X503. Amplifiers amplify in the band 80 to 2 500 MHz, approximately. The gain of the amplifier should be G = 19 dB at 850 MHz, approximately, and the noise figure NF = 2.6 dB.

2.7.4 Balanced mixer

Two balanced mixers M1 and M2, equipped with RFF207, are located between X400 and X601, and between X508 and X604. The output X604 or the input X 600 can be interconnected with the input/output of the IQ modulator/demodulator using jumpers. This user’s guide does not deal with the IQ modulator.

3 Scalar normalization

3.1 Theoretical introduction

In several radio-engineering measurements, knowledge of the transmission module or the position of spectrum maxims is sufficient; phase information is not important. For such measurements, spectrum analyzers are used.

Since RF cables, connectors, and transmission lines towards the DUT have non-zero attenuation and are dispersive, the effect of these errors has to be taken into account and eliminated before starting the DUT measurement.

3.2 Measurement

Output of the generator should be connected to the X202 SMA input and the input of the spectrum analyzer should be connected to the X204 SMA input.

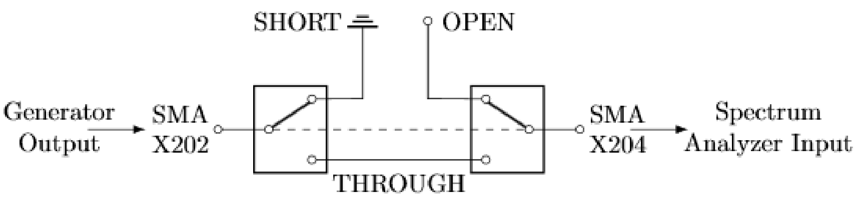

Figure 3.1: Wiring for scalar calibration

f (MHz)	Signal Generator Power Level $P_{\text{gen}}^{\text{nom}}$ (dBm)	Measured Power $P_{\text{gen}}^{\text{meas}}$ (dBm)	Path Loss $L_{\text{path}} = P_{\text{gen}}^{\text{nom}} - P_{\text{gen}}^{\text{meas}}$ (dB)
100	-20		
500	-20		
836.5	-20		
1673	-20		
2509.5	-20		

Table 3.1: Measured powers

3.3 Conclusion

The attenuation consists of partial contributions of the attenuation of the coaxial cable, connectors and semiconductor switches used in the kit. The total attenuation of the measured path has to be compensated. Since the phase information is missing, the measurement is called the scalar one.

4 Calibration of vector analyzer

4.1 Theoretical introduction

The Vector Network Analyzer (VNA) is the most comprehensive and versatile measuring instrument used in radio engineering. In order to describe high frequency and microwave circuits comprehensively, scattering parameters are used (s-parameters).

In fundamentals of circuit theory, the systems of impedance parameters, admittance parameters and hybrid parameters of linear n ports are introduced. Their definition and measurement requires well-defined short circuits (short ports) and open ends (open ports). Such requirements are difficult to be met at high frequencies and in microwaves. For example, SMD packages can contain a number of parasitic leaks and latent immittances which degrade the nominal value of the element and can cause resonance effects. Therefore, the system of scattering parameters was introduced.

The system of scattering parameters requires the termination by the characteristic impedance Z_0, usually $Z_0 = 50\ \Omega$. This termination brings several advantages. The real impedance Z_0 can be relatively easily implement, and moreover, the termination is well defined in a large bandwidth. In addition, active loads support the stability of active elements being better suited to operating conditions than short circuits and open ends.

Scattering parameters $s_{ij} \in \mathbb{C}$ can be defined for two-port networks. The two-port network can be measured by a system with a characteristic impedance Z_0.

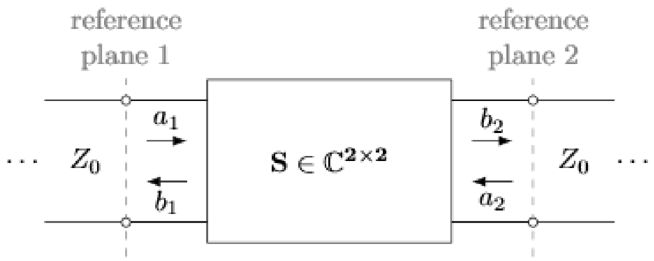

Figure 4.1: Definition of scattering parameters of a two-port network

Properties of two-port networks are measured in reference planes 1 and 2. In reference planes, normalized incident and reflected voltage waves are defined

$$a_i = \frac{V^+}{\sqrt{Z_0}}$$

$$b_i = \frac{V^-}{\sqrt{Z_0}}$$

$i = 1{,}2$. The incident power can be expressed as

$$P_i^+ = \frac{1}{2}|a_i|^2$$

And the reflected power can be given by

$$P_i^- = \frac{1}{2}|b_i|^2$$

The two-port network can be described by the system of scattering parameters $s_{ij} \in \mathbb{C}$

$$\begin{pmatrix} b_1 \\ b_2 \end{pmatrix} = \begin{pmatrix} s_{11} & s_{12} \\ s_{21} & s_{22} \end{pmatrix} \begin{pmatrix} a_1 \\ a_2 \end{pmatrix}$$

The system of s parameters is consistent with other systems. The systems can be mutually transformed.

4.2 VNA calibration

The possibility to eliminate all systematic errors is a significant advantage of VNA measurements. The error expresses the fact that s parameters in the data plane are different than true s parameters in the reference plane of DUT. The errors are caused by reflections at the inputs and the outputs of the measuring system, by the finite isolation of directional couplers inside the VNA, by imperfections of RF transitions, etc.

Since s parameters inherently describe the DUT as a linear system, errors can be depicted by a linear transformation from the reference plane to the data plane, and vice versa. This can be understood as an insertion of fictitious error two-port networks. If error two-port networks are known, the measurement can be corrected by transforming the data plane into the reference plane.

Error two-port networks are determined during the calibration. Instead of DUT, calibration standards with known, well-defined s parameters are connected.

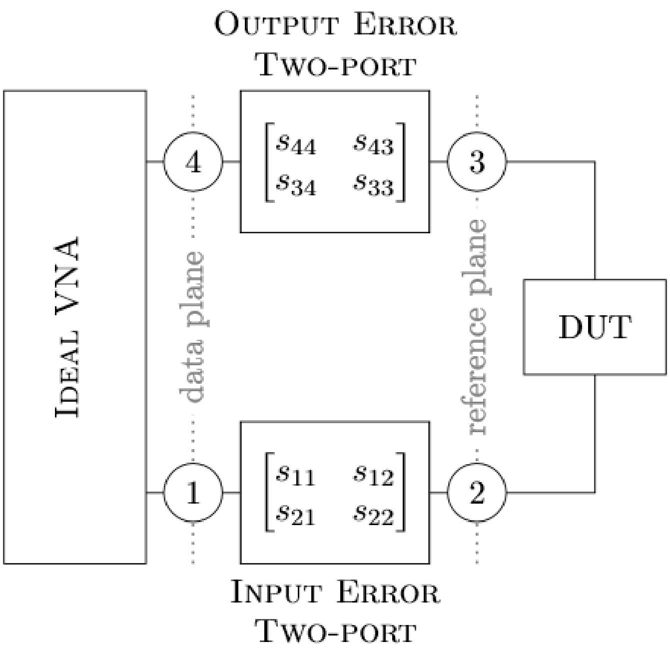

Figure 4.2: Error two-port network

The role of the calibration standard can be played by:

- Short circuit (SHORT) $s_{11} = \mathrm{e}^{-\mathrm{j}180°}$
- Open end (OPEN) $s_{11} = \mathrm{e}^{\mathrm{j}0°}$
- Matched load (MATCH) $s_{11} = 0$
- Through line (THROUGH) $s_{11} = 0,\ s_{21} = \mathrm{e}^{-\mathrm{j}\varphi°}$

Using these known calibration standards with known s parameters and measured s parameters in the data plane, a system of linear equations can be compiled, and error two-port networks are obtained as a solution of this system.

4.3 Measurements

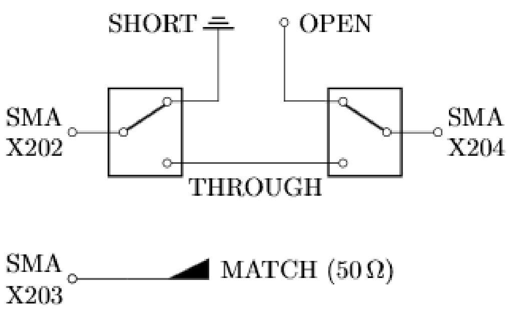

Figure 4.3: Calibration kit

1) The R&S®FPC1500 analyzer is switched to the “Vector Network Analyzer” mode.

2) The Smith Chart is selected as the format of reflection measurements s_{11}. The “Full 1-Port” calibration is performed (details are given in the FPC1500 user’s guide).

3) The Generator Output is successively connected to the corresponding calibration standards on SMA connectors of the kit; the connectors are marked X202, X203 and X204. Make sure that the position of the jumper 6 does not connect X202 and X204.

4) The calibration is verified by other calibration standards which were not used in the calibration process. Verification standards should be of the comparable electrical length.

A second kit, if available, can be used for the verification.

- SHORT standard verification:

The true value of reflectiont s_{11} is recorded in the Smith chart and compared with the nominal position of the ideal calibration standard SHORT. Inductive or capacitive character of the reflection is indicated.

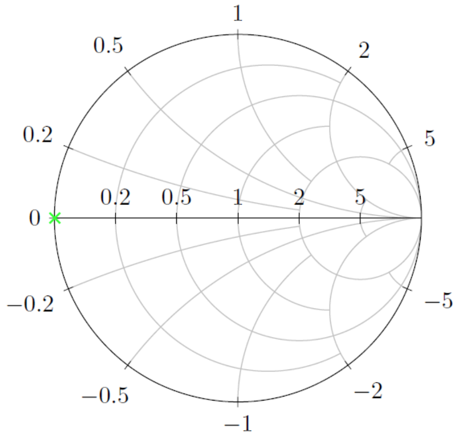

Figure 4.4: Smith chart (SHORT)

- OPEN standard verification:

The true value of reflection s_{11} is recorded in the Smith chart and compared with the nominal position of the ideal calibration standard OPEN. Inductive or capacitive character of the reflection is indicated.

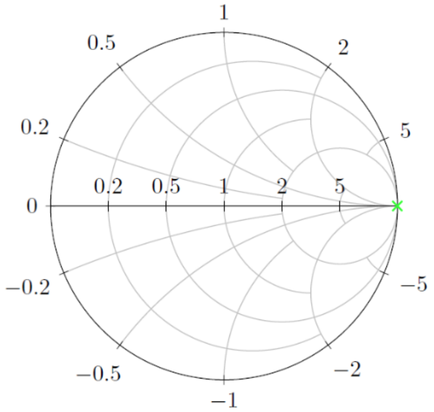

Figure 4.5: Smith chart (OPEN)

- MATCH standard verification:

The true value of reflection s_{11} is recorded in the Smith chart and compared with the nominal position of the ideal calibration standard MATCH.

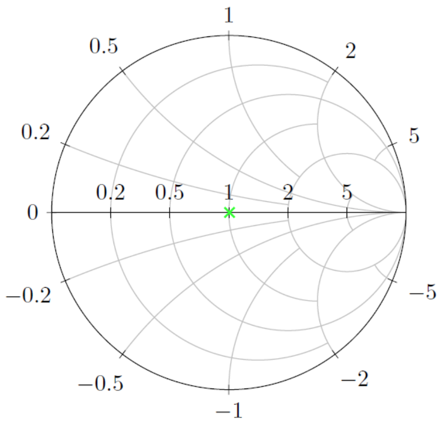

Figure 4.6: Smith chart (MATCH)

5 Bandpass filter

5.1 Theoretical introduction

Frequency selective n ports or filters are the fundamental element of most radio engineering equipment. The R&S® FPC-Z10 kit is equipped with a pair of bandpass filters. These filters are passive two-port networks. Various frequency switches and diplexers are examples of general n port filters.

The filter is requested to suppress selected parts of the spectrum of the passing signal in a defined way. Since the filter is assumed to be linear, the input-output relation can be described by a convolution integral in the time domain

$$s_2(t) = \int_{-\infty}^{\infty} h(t - \tau) s_1(\tau)\,\mathrm{d}\tau$$

Here, $h(t)$ is the pulse response of the filter, $s_1(t)$ is the input signal and $s_2(t)$ is the output signal. Image of the operator $H(\omega) = F[h(t)](\omega)$ of the pulse response is symmetric in the Hermitian sense with respect to the Fourier transform:

$S_2(\omega) = H(\omega) S_1(\omega)$

Here, $S_1(\omega)$ and $S_2(\omega)$ are operator images of signals $s_1(t)$ and $s_2(t)$. The function $H(\omega)$ is the transfer function of the filter.

Analysis and synthesis of transfer functions of frequency selective systems and synthesis of circuit structures implementing transfer properties of such functions $H(\omega)$ is a sophisticated area of the conventional theory of linear circuits.

Module of the transfer function $|H(\omega)|$ can show iso-extremal ripples due to various reasons. These ripples can be defined as follows

$$\varepsilon = \max(|H(\omega)|) - \min(|H(\omega)|)\ \forall \omega \in B_{\mathrm{filtr}}$$

Here, the quantity B_{filtr} denotes the operation bandwidth of a corresponding filter (band-stop, high-pass, low-pass, band-pass).

Selectivity denotes the slope of module characteristics at margins of bands. In general, maximally flat filters with the Butterworth approximation in the passband have low selectivity, and highly selective elliptical filters with transmission zeros at finite frequencies show high ripples.

In case of filters, phase characteristics $\arg H(\omega)$ and the group delay are often the matter of interest.

$$\tau(\omega) = -\frac{\mathrm{d} \arg H(\omega)}{\mathrm{d}\omega}$$

The group delay describes dispersive properties of the filter at break frequencies, especially.

5.2 Measurement

1) The R&S®FPC1500 analyzer is switched to the "Vector Network Analyzer" mode. Then, we set:

- Center frequency: 836.5 MHz
- Reference level: 0 dB
- RBW: 10 kHz
- SPAN: 100 MHz

Then, the normalization of s_{21} and the calibration of s_{11} are performed.

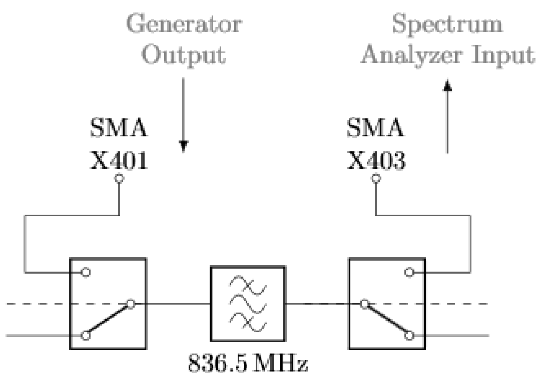

Figure 5.1: Wiring for bandpass filter measurements

2) The generator output is connected to the X401 SMA input and the RF input is connected to the X403 SMA input

The measured bandpass filter is a reciprocal circuit and exchanging the ports does not have any effect. Let us verify that!

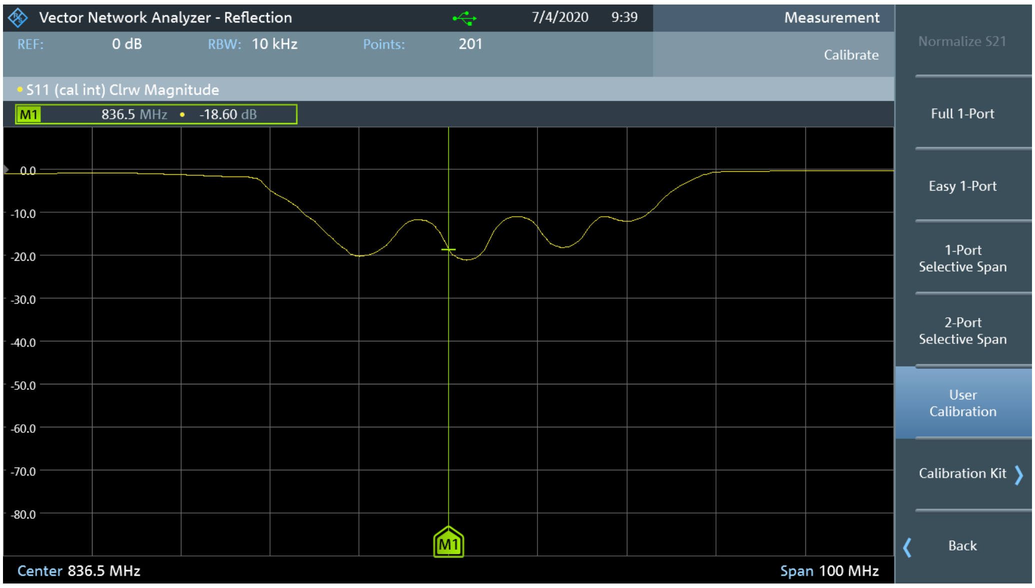

Figure 5.2: Response of the reflection module

3) Response of the reflection module $|s_{11}|$ should be measured

4) Response of the transmission module should be measured $|s_{21}|$

- The bandwidth of the bandpass filter $B_{-3\mathrm{dB}}$ and break frequencies should be determined (the transmission module $|s_{21}|$ drops by 3 dB compared to the module at $f = 836.5$ MHz)
- The value of the insertion loss IL at $f = 836.5$ MHz should be determined
- The value ripples ε within the band $B_{-3\mathrm{dB}}$ should be determined

Figure 5.3: Response of the operational transmission module

6 Attenuator

6.1 Theoretical introduction

Attenuators are passive, lossy two-port networks those exhibit a non-zero, well-defined attenuation $L \neq 0$ dB for the transmission between the input port and the output port. Attenuators are manufactured as networks with a fixed attenuation or a variable one. Attenuators usually consist of the T topology or the π topology of the network of resistors or PIN diodes.

Attenuators are used to control the power level of the signal. Thanks to the attenuators, extra strong signals can be measured, and the inter-modulation distortion can be suppressed. Tunable attenuators are the key element of loops of the automatic level control (ALC). In the measuring technology, attenuators are used for impedance matching.

The signal from the generator passes the attenuator with the attenuation $L = 10$ dB. A part of the wave is reflected from the measured device under test (DUT). The reflected wave passes the attenuator with the attenuation $L = 10$ dB again. From the viewpoint of the generator, the DUT is very well matched.

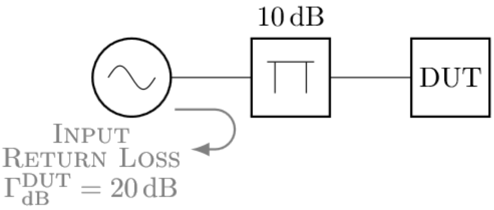

Figure 6.1: Attenuator as a matching element

The kit contains adjustable attenuators accessible via DIP switches. Attenuations set by switch positions are summed. Adjustable attenuations are ranging from 0 dB to 31.5 dB.

True attenuations may differ from the adjusted nominal attenuations, and this additional attenuation has to be included into the measurement.

6.2 Measurements

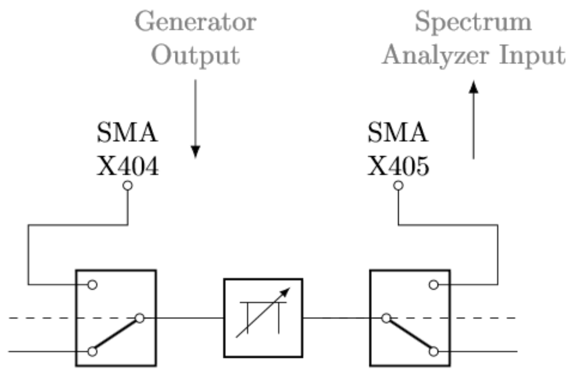

Figure 6.2 Wiring for measuring attenuator

1) The R&S®FPC1500 analyzer is switched to the “Vector Network Analyzer” mode. Then, we set:

- Start frequency: 2 MHz
- Stop frequency: 3 GHz
- Reference level: 0 dB
- RBW: 10 kHz
- The normalization of s_{21} should be performed

2) The generator output is connected to the X404 SMA input and the RF input is connected to the X405 SMA input

3) Attenuation of the attenuator can be changed by positions of the DIP switch 7 according to the table below.

Table 6.1: DIP switch settings

Attenuation (dB)	DIP switch settings					
	1 0.5 dB	2 1 dB	3 2 dB	4 4 dB	5 8 dB	6 16 dB
0	ON	ON	ON	ON	ON	ON
10	ON	ON	OFF	ON	OFF	ON
20	ON	ON	ON	OFF	ON	OFF
30	ON	ON	OFF	OFF	OFF	OFF

Required attenuations should be set and the true attenuation for three frequencies should be measured according to the table below. The influence of the increasing frequency should be observed, and the own attenuation of the attenuator should be determined at the nominal attenuation of 0 dB at $f = 836.5$ MHz.

Attenuation (dB)	Measured (dB) 836.5 MHz	Measured (dB) 1.673 GHz	Measured (dB) 2.5095 GHz
0			
10			
20			
30			

Table 6.2: Measured attenuations

Figure 6.3: Example of attenuation measurement

7 RF power amplifier

7.1 Theoretical introduction

The amplifier is an active two-port network that amplifies the level of input signals. According to the bandwidth of amplified frequencies, amplifiers can be divided to narrow-band ones with the fractional bandwidth FBW

$$FBW = \frac{f_0}{B} \leq 0.1 f_0$$

where f_0 is the geometric average of the break frequencies of the amplified band B, and to broad-band ones with $FBW > 0.1 f_0$. The bandwidth of today's MMIC implementations can reach even several decades.

According to the level of amplification, amplifiers can be classified as low-power ones (for highly sensitive front ends of radio receivers) and high-power ones (back ends of radio transmitters). In the task, the high-power amplifier is measured.

From the viewpoint of operation, the amplifier is usually characterized by the following parameters:

- Efficiency η given by the class of the amplifier
- The input matching s_{11} and the output matching s_{22}
- The transducer gain G_t defined for general impedances of the excitation source and the load, $G_\mathrm{t} = 20 \log_{10}|s_{21}|$ (dB)
- Dynamic range DR, operating bandwidth B
- Sensitivity (the lowest level of the input power the amplifier is able to amplify)
- The IP1 intercept point (the point of 1 dB gain compression)
- The IP3 point (the point characterizing the inter-modulation distortion for the excitation of the amplifier by a biharmonic signal)
- Harmonic distortion characterized by the factor of harmonic distortion THD

  $$\mathrm{THD}\ (\%) = 100 \sqrt{\frac{U_2^2 + U_3^2 + \cdots + U_n^2}{U_1^2}}$$

  Here, U_1^2 is the square of the rms value of the voltage of the fundamental harmonic of the excitation signal and U_i^2, $i = 2, \dots, n$, n is the number of considered harmonic products
- Noise figure NF

7.1.1 Intercept point (IP1)

Active elements of the amplifiers (BJT, MOS, MES, HEMT transistors, etc.) are linear (quasilinear) only for a given range of excitation powers. Due to the saturation of the output power to a constant value P_out the signal is limited. This nonlinearity causes the production of harmonic frequencies nf_1, $n \in \mathbb{Z}$ (inter-modulation products of the 1st order) of the fundamental excitation harmonic signal f_1. In case of the excitation by a biharmonic signal with frequencies f_1 and f_2, $f_1 \neq f_2$, the nonlinearity produces the signal with products $\pm m f_1 \pm n f_2, \forall m, n \in \mathbb{Z}$ at the output. Such products are denoted as inter-modulation products of the $m + n$ order. Practically, the product of the 3rd order is important since being close to useful signals. If a power combiner and a second generator are available, then the IP3 point of a given amplifier can be measured for the FPC-Z10 kit.

The degree of nonlinearity for the harmonic excitation is quantified by the IP1 intercept point. The intercept point can be determined as follows: at the selected frequency, the generator excites the amplifier by the input power P_{in} and the spectrum analyzer is used to measure the amplified power P_{out}. The intercept point IP1 is determined then by linear interpolation of measured points (see figure 8.2).

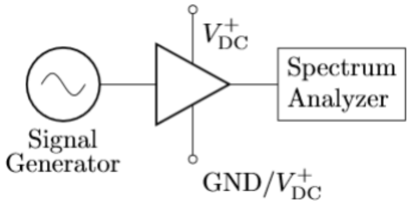

Figure 7.2: Wiring for measuring the intercept point

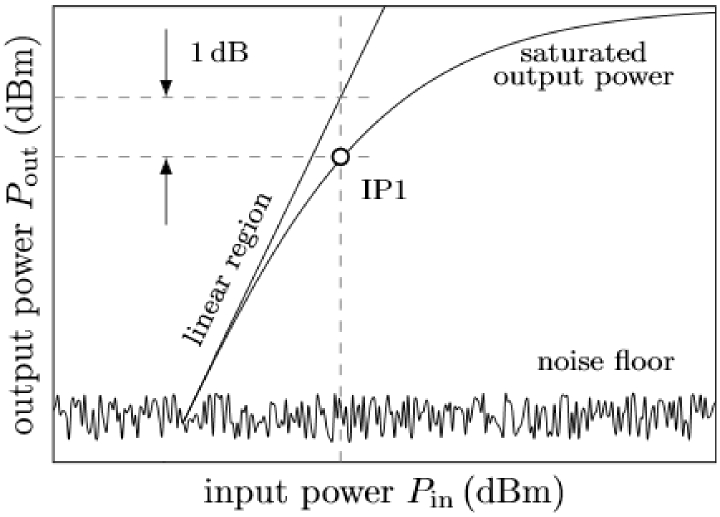

Figure 7.1: On the definition of the intercept point

In following sections, some of the above parameters will be measured for the amplifier available in the kit.

7.2 Measurement of transmission properties

1) The R&S®FPC1500 analyzer is switched to the “Vector Network Analyzer” mode. Then, we set:

- Start frequency: 2 MHz

- Stop frequency: 3 GHz
- Reference level: 20 dB
- RBW: 10 kHz
- The normalization of s_{21} should be performed

2) The generator output is connected to the X406 SMA input and the RF input is connected to the X402 SMA input

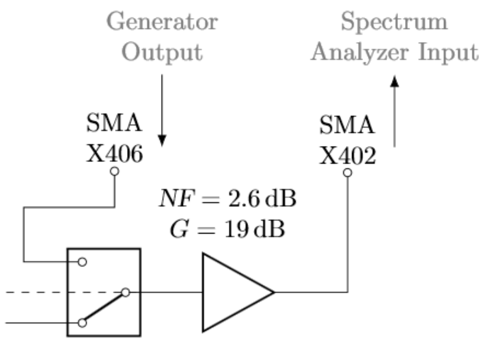

Figure 7.3: Wiring for measuring the linearity of the amplifier

3) The gain $G = 20\log_{10}|s_{21}|$ (dB) is measured. The maximum and minimum value of the gain and the bandwidth of the amplifier should be determined, where the gain $G \geq 10$ dB.

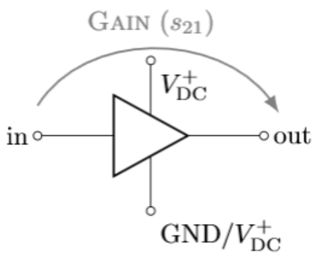

Figure 7.4: Simplified definition of the gain

4) The reference level should be changed to 0 dB and the OSM calibration of the reflection “Full 1-Port” should be performed.

5) The input reflection of the amplifier $|s_{11}|$ should be measured and the minimum should be found.

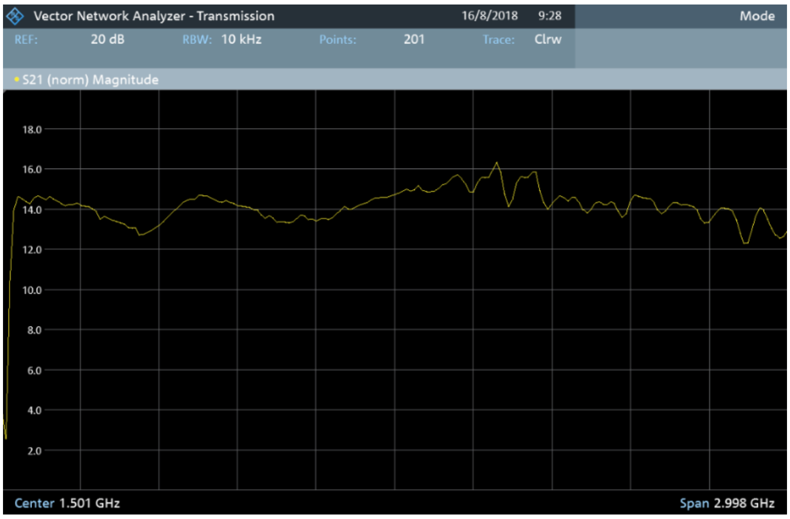

Figure 7.7: Response of the module of the transmission

Figure 7.6: Response of the module of the reflection

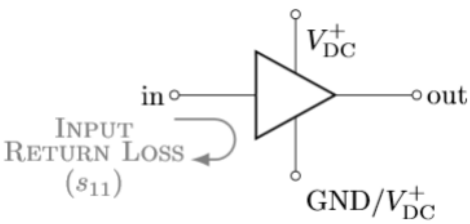

Figure 7.5: Definition of reflection

The amplifier is not a reciprocal circuit. The backward transmission s_{12} defines the isolation parameter. Let us verify that.

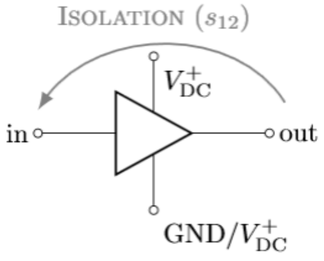

Figure 7.9: Definition of isolation (backward transmission)

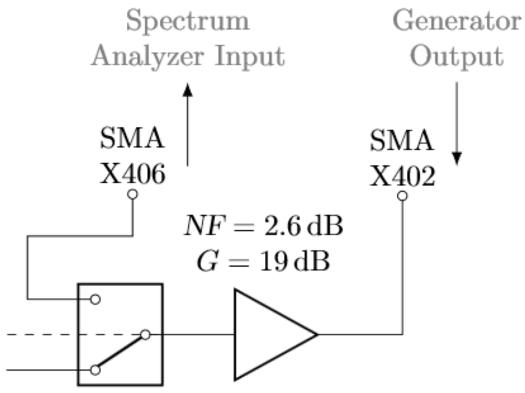

Figure 7.10: Wiring for measuring the backward transmission

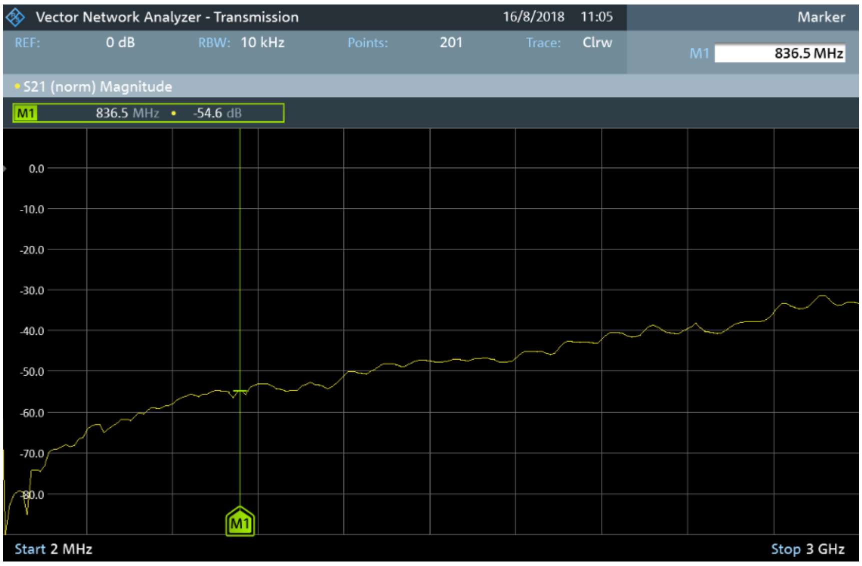

Figure 7.8: Response of the module of the backward transmission

7.3 Linearity measurement of the amplifier, IP1

1) The R&S®FPC1500 analyzer is switched to the “Spectrum” mode. Then, we set:

- Center frequency: 836.5 MHz
- SPAN: 20 MHz
- RBW: 10 kHz
- VBW: 10 kHz
- ATT: 30 dB

And for the generator:

- Source: CW
- Frequency: 836.5 MHz
- Level: -20 dBm

The wiring should be returned to the initial state.

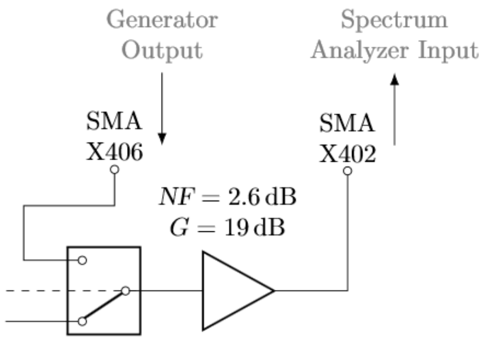

Figure 7.11: Wiring for measuring the intercept point

The tables below should be used to make records of measured powers P_{out}. for $f = 836.5$ MHz and prescribed excitation powers P_{in}.The values should be corrected for the attenuation of a cable and RF connectors (L_{p}) in the sense of the task 1. The value of the gain G should be calculated.

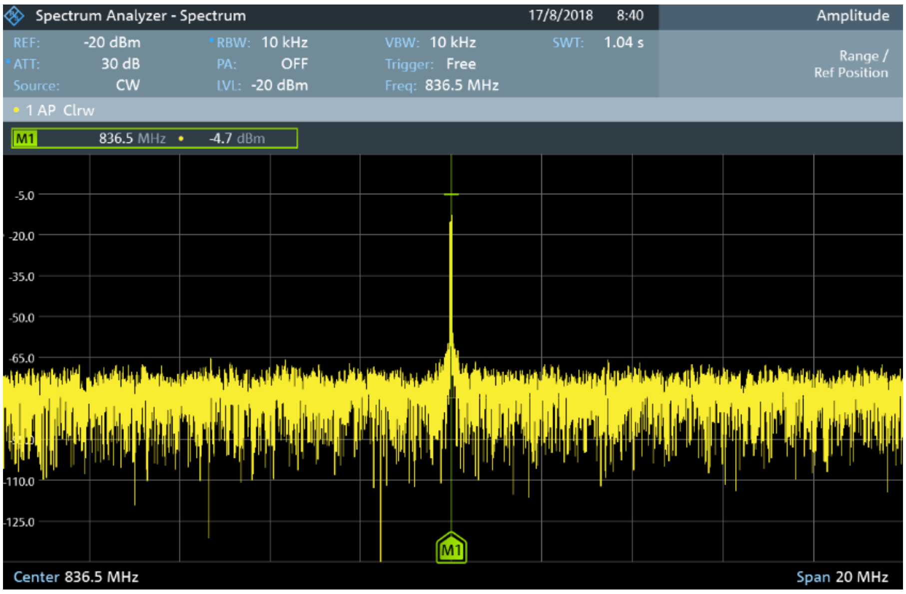

Figure 7.12: Example of the response of the output power

Values G should be plotted in the chart and the IP1 intercept point should be determined.

Table 7.1: Measured powers and gain of the amplifier

P_{in} (dBm)	−20	−19	−18	−17	−16	−15	−14	−13	−12	−11
P_{out} (dBm)										
L_{p} (dB)										
$G = P_{\mathrm{out}} - P_{\mathrm{in}} + L_{\mathrm{p}}$ (dB)										

P_{in} (dBm)	−10	−9	−8	−7	−6	−5	−4	−3	−2	−1	0
P_{out} (dBm)											
L_{p} (dB)											
$G = P_{\mathrm{out}} - P_{\mathrm{in}} + L_{\mathrm{p}}$ (dB)											

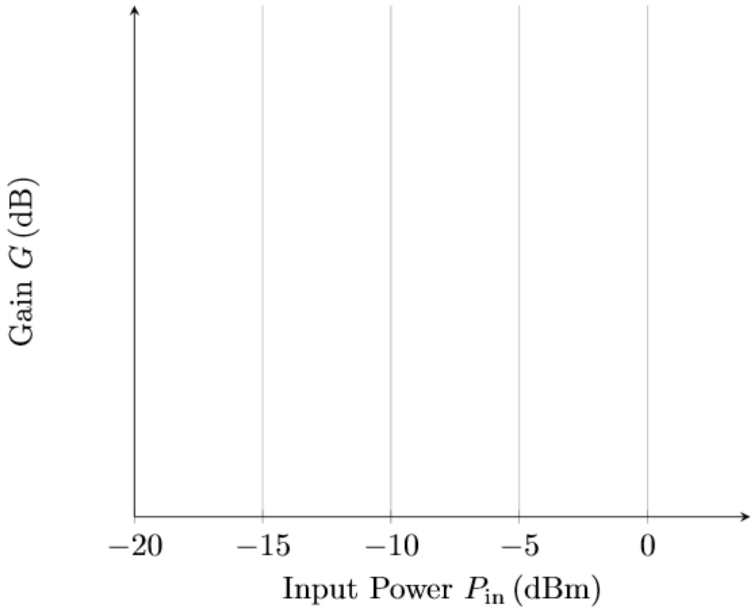

Figure 7.13: Dependence of the gain on the input power

7.4 Measurement of harmonic distortion factor

Note: First-order products produced by the generator are neglected in this measurement.

1) The R&S®FPC1500 analyzer is switched to the “Spectrum” mode. Then, we set:

- SPAN: Full span
- RBW: 100 kHz
- VBW: 1 MHz
- ATT: 30 dB

And for the generator:

- Source: CW
- Frequency: 836.5 MHz
- Level: -5 dBm

The wiring is the same as described in the previous section.

2) The distance of harmonic components has to be determined. The correction has to be applied !

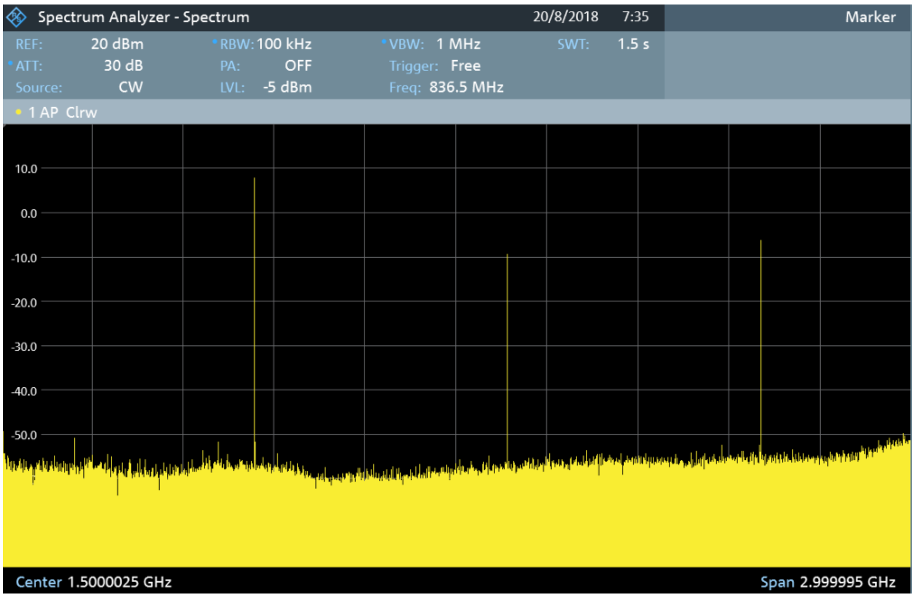

Figure 7.14: Products of the harmonic distortion

3) On the R&S®FPC1500, the “MEAS→Measurement Mode→Harmonic Distortion” has to be turned on. In the “Harmonics” item, the required number of harmonics has to be set to 3. The “Harmonic Distortion” mode directly indicates the value of the total harmonic distortion (THD) factor.

Figure 7.15: Example of measuring the harmonic distortion factor on the R&S® FPC-1500

8 Mixer measurement

8.1 Theoretical introduction

The mixer is a nonlinear three-port network generating signals with new frequencies due to nonlinearities. On the RF port, the signal with frequency f_{RF} is applied. On the LO port, the signal with frequency f_{LO} is applied. On the IF port, the signal with intermodulation products appears

$$\pm n f_{\text{RF}} \pm m f_{\text{LO}}, \quad n, m \in \mathbb{Z}$$

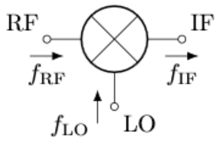

Figure 8.1: Description of ports of the mixer

The principle of the super-heterodyne receiver consists in connecting a high-quality bandpass filter (an intermediate frequency filter) to the IF port. A high-frequency modulated signal is applied to the RF port and a tunable local oscillator is connected to the LO port.

From all the products at the IF output, the bandpass filter selects one of the signals with the frequency $m, n = 1$

$$f_{\text{IF}} = f_{\text{RF}} \pm f_{\text{LO}}$$

In case of the sign $+$ an up-converter is obtained. In case of the sign $-$ a down-converter is implemented.

The equation clearly shows that for the signal f_{RF} mixed into the IF filter on f_{IF} there is a signal $f_{\text{RF}}^{\text{image}}$ shifted for f_{IF} from f_{LO} (on the opposite side than f_{RF}). This so-called mirror signal is mixed into the intermediate frequency as well.

The mirror reception is usually undesired and has to be effectively suppressed by a proper concept of a mixer circuit (Image Reject Mixer) or by double or triple mixing. First, the signal is mixed to a high intermediate frequency f_{IF}. That way, the position of the mirror signal $f_{\text{RF}}^{\text{image}}$ far from the frequency of the local oscillator f_{LO} is ensured. And second, the mirror signal is easily filtered out by a non-selective filter. The signal processed that way proceeds for further mixing according to the frequency plan of the receiver.

Mixers can be active ones or passive ones.

For passive mixers, conversion losses can be determined:

$$L_k = P_{RF} - P_{\text{IF}} \text{ (dB)}$$

Here, P_{RF} (dBm) is the signal power at the RF port with frequency f_{RF} and P_{IF} (dBm) is the signal power at the IF port with frequency f_{IF}.

The isolation, describing the leakage of the RF signal with f_{RF} to the IF port, is the next parameter.

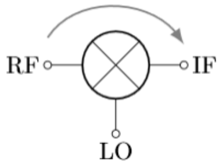

Figure 8.2: Conversion losses

8.2 Measurement

Note: When the signal propagates through the separate two-port networks of the kit, new undesired inter-modulation products (the spurious output) are generated. These products are ignored in the measurement.

1) The R&S®FPC1500 analyzer is switched to the “Spectrum” mode. Then, we set:

- Center frequency: 636.5 MHz
- SPAN: 10 MHz
- Reference level: -20 dBm
- RBW: 300 kHz
- ATT: 0 dB

The output of the RF generator should be left in the OFF state for now.

2) The following wiring should be connected

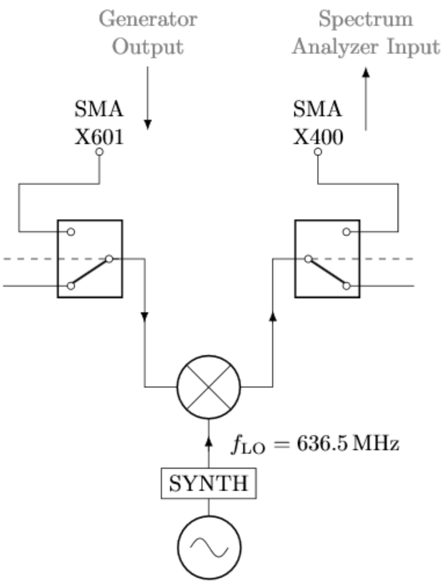

Figure 8.3: Schematics of wiring

3) Positions of X301 and X302 jumpers should be in the position 1. At the output of the X400 SMA connector, a signal similar to the response shown in the figure below should appear. Otherwise, the USB power supply of the kit should be disconnected and reconnected.

Figure 8.4: The power propagating from LO to the IF port

The measured power is the power of the local oscillator at $f_{\mathrm{LO}} = 636.5\ \mathrm{MHz}$.

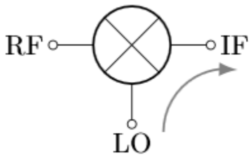

Figure 8.5: Isolation

4) The RF generator should be set as follows:

- Source: Tracking generator
- TG offset: 0 Hz
- Level: 0 dBm

A on the spectrum analyzer

- Start frequency: 2 MHz
- Stop frequency: 2363.5 MHz
- Reference Level: 0 dBm
- RBW: 300 kHz
- VBW: 300 kHz
- ATT: 30 dB

5) The isolation of RF-IF ports for $f_{\mathrm{RF}} = 836.3\ \mathrm{MHz}$ should be determined

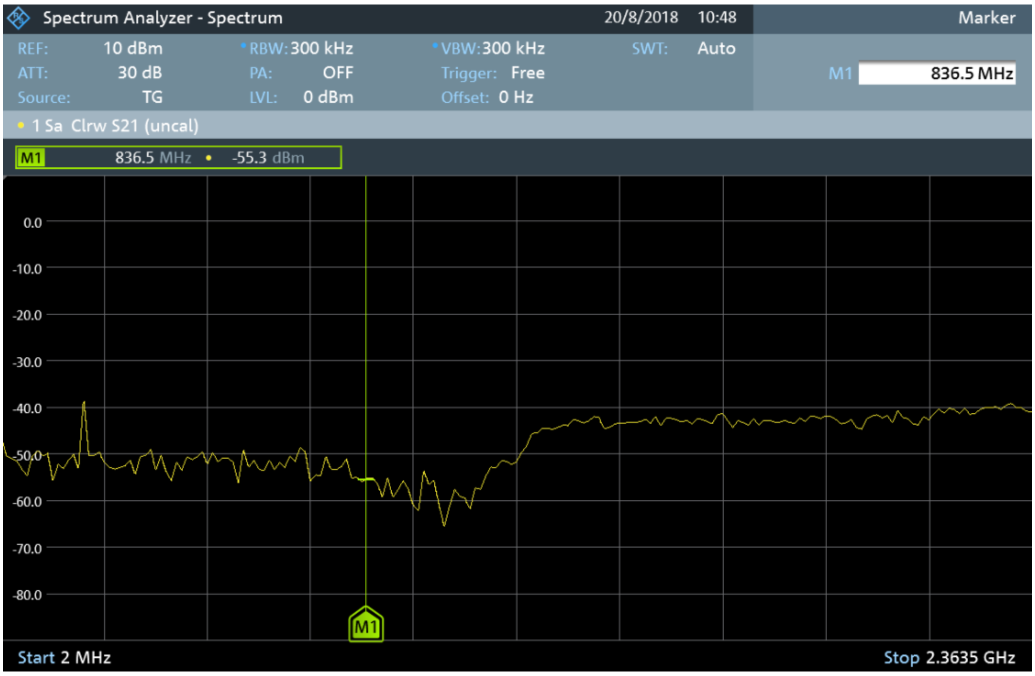

Figure 8.6: Example of measuring the response of isolation

6) For the tracking generator:

- TG offset: 636.5 MHz

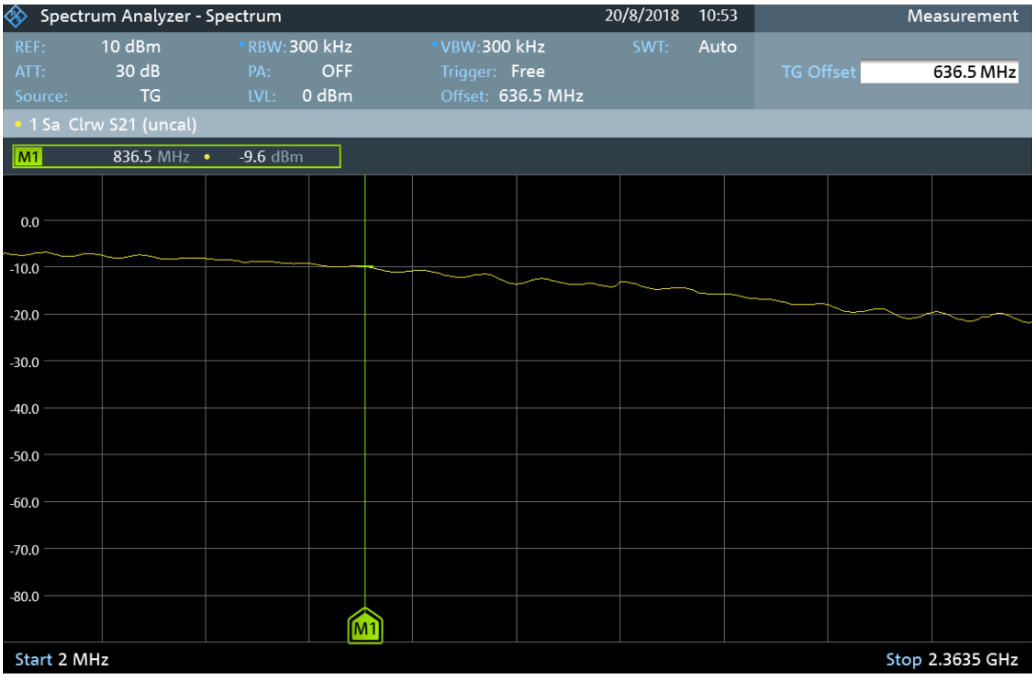

Figure 8.7: Example of measuring the response of conversion losses

That way, conversion losses L_k can be determined

Let us assume $f_{RF} = 1473$ MHz. The conversion losses can be determined for the corresponding intermediate frequency f_{IF} if the frequency of the local oscillator is $f_{LO} = 636.5$ MHz.

Note: Correction of the attenuation of cables and connectors should not be considered.

7) The FPC1500 analyzer is switched to the “Vector Network Analyzer” mode. Then, we set:

- Start frequency: 2 MHz
- Stop frequency: 3GHz
- Reference level: 0 dB
- RBW: 10 kHz

The “Full 1-Port” calibration should be performed, or the Recall function should be used to load settings from previous tasks.

8) Module the reflection $|s_{11}|$ should be measured, the minimum of the reflection should be found, and the value should be read for:

- $f_1 = 200$ MHz
- $f_2 = 1.473$ GHz

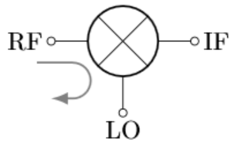

Figure 8.9: Input reflection

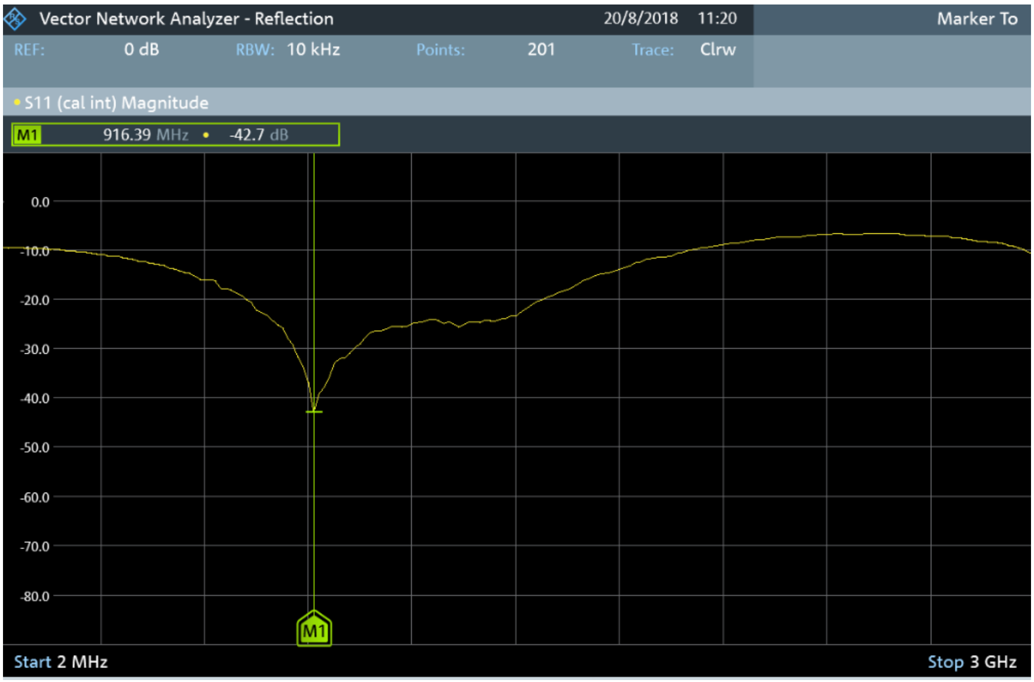

Figure 8.8: Example of measuring the module of the reflection

9 RF front end transceiver

9.1 Theoretical introduction

Nearly all the radio communication is based on the principle of the super-heterodyne reception. The super-heterodyne receiver (hereinafter referred to as a superhet) is a suitably arranged cascade of RF functional blocks measured in previous tasks.

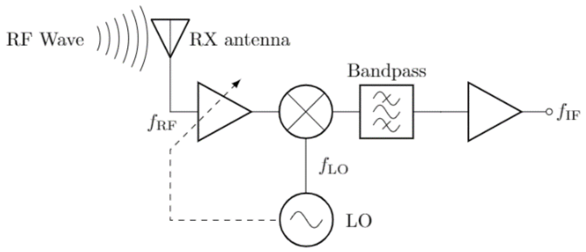

Figure 9.1: A simplified schematics of superhet

By re-tuning the frequency of the LO local oscillator, the received frequency f_{RF} is selected, mixed and forwarded into the interference filter as the selected interference f_{IF}. From the filter, the signal flows to the blocks of the signal detection and demodulation.

The multiple mixing (up to the triple mixing) is aimed to eliminate problems with the undesired mirror frequency f_{IM} by the mixing towards the high intermediate frequency. That way, the mirror frequency is shifted far away from the received frequency and can be easily filtered out by the low-pass filter, which is not requested to have as strong selectivity as in the case of a single mixing. Exploitation of different concepts, for example an IRM receiver, is another option.

In the task, the entire transmission chain (including the radio channel transmission) is created and fundamental principles of such a simple communication are demonstrated.

9.2 Measurement

Note: The measurement requires a pair of antennas designed for the RF operation up to 3 GHz.

1) The FPC1500 analyzer is switched to the “Spectrum” mode. Then, we set:

- Center Frequency: 836.5 MHz
- Reference level: -10 dBm
- SPAN: 200 MHz
- RBW: 100 MHz
- VBW: 100 MHz
- ATT: 0 dB

And at the generator output:

- CW
- Frequency: 200 MHz
- Level: 100 dBm

2) The kit should be connected according to the below-given schematics.

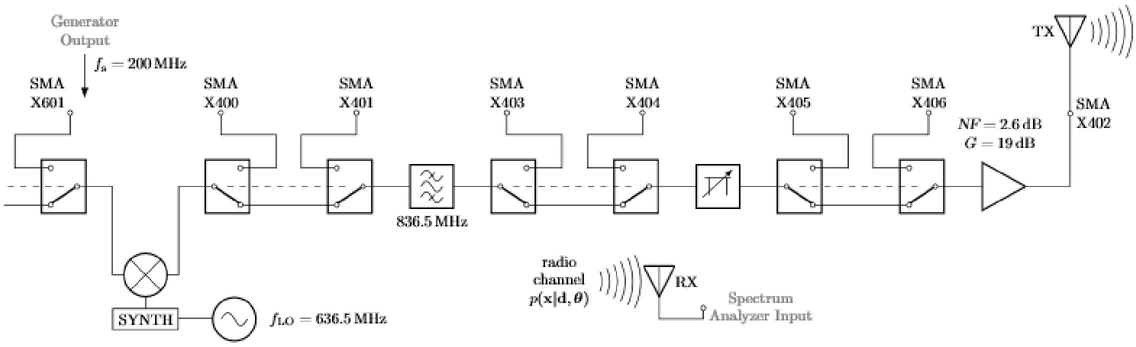

Figure 9.2: Schematics of wiring the up-converter

The X407, X408, X409 jumpers should be verified to allow the signal flow from the mixer to the amplifier. The X301 and X302 jumpers of the synthesizer status selection should be in the position 1. The jumper X608 is in the position 1.

The Tx transmit antenna should be placed far away from the Rx receive antenna. The position of antennas against obstacles should be experimented.

3) The power value of the received signal P_{Rx} should be read on the spectrum analyzer.

Figure 9.3: Example of measuring the received power

The lower power value can be explained by free space losses (FSL) and mismatched antennas.

4) The antennas should be placed to be without the direct line of sight, or the transmit power P_{Tx} should be changed at the generator output.

5) The CW frequency of the RF generator should be changed according to the table, the frequency of the converted signal f_{Tx} should be calculated and the received power P_{Rx} should be measured. The dependency P_{Rx} on f_{Tx} should be plotted into a chart and compared with the dependency of the module of the transmission characteristics of the bandpass filter measured in the task @7.

f_s (MHz)	f_{Tx} (MHz)	P_{Rx} (dBm)
170		
180		
190		
200		
210		
220		
230		

6) On the FPC-1500 analyzer, we should change:
SPAN: Full span

And at the generator output:

- Frequency: 200 MHz
- Level: 0 dBm
- RBW: 30 kHz

Figure 9.4: Traffic in the ISM band

And the presence of other signals should be examined. In order to eliminate spectral lines caused by the signal itself, the generator output should be turned off. At the frequency $f = 2.45\ \mathrm{GHz}$ in the band $B = 100\ \mathrm{MHz}$, a significant traffic can be expected with a high probability. The above-specified ISM band is used for WLAN networks, Bluetooth communication, etc.

9.3 Down-converter measurement

1) On the FPC1500 analyzer, we should adjust:

- Center frequency: 200 MHz
- Reference level: -20 dBm
- SPAN: 200 MHz
- RBW: 1 MHz
- VBW: 1 MHz
- ATT: 0 dB

And at the generator output:

- CW
- Frequency: 836.5 MHz
- Level: 0 dBm

2) The kit should be connected according to the below-given schematics.

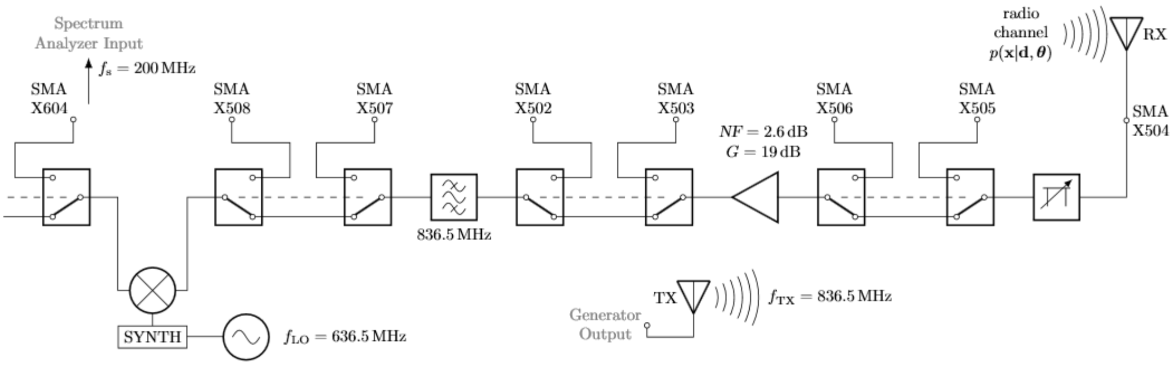

Figure 9.5: Schematics of wiring the down-converter

3) The level of the product of mixing should be read. As in the previous task with the up-conversion, the width of the passband of the bandpass filter of the receiver should be re-tuned.

Figure 9.6: Example of measuring the received power

10 Spectrum analyzer

10.1 Theoretical introduction

If the RF input of the spectrum analyzer is terminated with a non-reflecting termination, the achievable spectrum power density of the input signal of the spectrum analyzer is

$$N_0(f) = kT_0 \cong -174 \text{ dBm/Hz},$$

where $T_0 = 290$ K. Such a signal represents a stationary and ergodic random process - an additive white Gaussian noise (AWGN). The signal power enters the resolution filter of the spectrum analyzer (an intermediate frequency filter, at the second to third intermediate frequency usually).

$$P_{\text{in}} = |H_{\text{max}}|^2 \int_0^\infty \left|\frac{H(f)}{H_{\text{max}}}\right|^2 S(f)\,\text{d}f = S(f)\,|H_{\text{max}}|^2\text{ENB}$$

Here, ENB is the equivalent noise bandwidth.

This bandwidth is defined as the bandwidth of an ideal filter with rectangular module characteristics, which can pass the same AWGN power as the intermediate frequency filter (or the resolution filter) with the characteristics $H(f)$. For the resolution filter, we can obtain ENB = kRBW, $k > 1$. After the calibration, we can assume for the module $|H_{\text{max}}|^2 = 1$.

By averaging the noise in a logarithmic scale, we cause the measurement error ≈ 2.5 dB. Then, the spectrum power density can be determined as

$$S(f) = P_{\text{in}} - 10\log\left(\frac{k\text{RBW}}{1 \text{ Hz}}\right) + 2.5$$

The SA sweep speed is slower for a narrower resolution filter. Spectrum power density $S(f)$ is typically higher than the theoretical value for the non-reflective load.

For SA, we can specify the noise figure at the selected frequency NF(f) [dB] as

$$\text{NF}(f) = S(f) - L_{\text{ATT}} - N_0(f)$$

Here, L_{ATT} is the attenuation of the SA input attenuator and $N_0(f)$ is the theoretically achievable spectrum density of the non-reflective load.

Assuming the input attenuator $L_{\text{ATT}} = 0$ dB, the SA noise temperature is significantly affected by conversion losses of mixing circuits.

The input attenuator primarily serves to prevent the over-excitation of the SA input and the generation of undesired products (spurious outputs) caused by nonlinearities of mixing circuits.

The SA noise threshold can be shifted using an LNA preamplifier. If the preamplifier is of a sufficient gain (15 to 20 dB), then the equivalent noise temperature of the LNA+SA cascade approaches the equivalent noise temperature of the LNA itself.

During sweeping, the input harmonic signal is converted to the proximity of the intermediate frequency by the multiplication. The SA screen shows the convolution of spectra of the resolution filter (the intermediate frequency filter) and the harmonic signal (the Dirac pulse for the infinitely long observation in the ideal case).

Hence, the displayed spectrum is the module power characteristics of the IF filter.

The RBW filter therefore determines the resolution in the spectrum.

10.2 Measurement

1) On the FPC1500 analyzer, we should adjust:

- Center frequency: 836.5 MHz
- Reference level: -20 dBm
- SPAN: 1 MHz
- RBW: 30 kHz
- VBW: 30 kHz
- ATT: 0 dB

A non-reflective termination should be connected to the RF input of the analyzer.

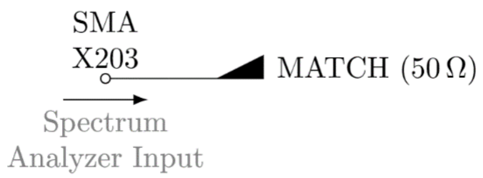

Figure 10.1: Wiring of the non-reflective termination

2) The width of the RBW resolution filter should be changed according to the table. The power P_{in} and the spectrum power density $S(f)$ should be read.

RBW	10 Hz	100 Hz	1 kHz	10 kHz	100 kHz	1 MHz
P_{in} (dBm)						
$S(f)\vert_{f=836.5\,\text{MHz}}$ (dBm/Hz)						

Obviously, the value of the averaged power P_{in} varies with RBW according to the given relations. Spectrum power density $S(f)$ stays the same. In order to change the noise threshold ΔN (the margin at least 10 dB above the noise threshold) when changing from RBW_1 on RBW_2 , one can obtain

$$\Delta N = 10\log(\text{RBW}_1/\text{RBW}_2)$$

3) The measured value of the spectrum power density $S(f)$ is typically higher than the theoretical value for the non-reflective load. From this, the noise figure of the spectrum analyzer can be determined.

The noise figure $\text{NF}(f)$ of input circuits of the spectrum analyzer is usually quite high (20 to 30 dB). This value might be surprising compared to the front-end inputs of conventional RF communication devices equipped with LNA blocks.

4) The value of the SA attenuator should be changed, and the background noise level should be monitored.

Each $10\ \mathrm{dB}$ of the attenuation shifts the SA noise threshold for $10\ \mathrm{dB}$ higher. When measuring very weak signals, as narrow as possible RBW has to be used. For the input attenuator, $L_{\mathrm{ATT}} = 0\ \mathrm{dB}$ has to be ensured. The VBW filter following the SA detector does not reduce the noise threshold. Nevertheless, the VBW is recommended to be selected for 20 dB narrower than RBW when measuring the noise.

The input attenuator primarily serves to prevent the over-excitation of the SA input and the generation of undesired products (spurious outputs) caused by nonlinearities of mixing circuits.

The SA noise threshold can be shifted using an LNA preamplifier. If the preamplifier is of a sufficient gain (15 to 20 dB), then the equivalent noise temperature of the LNA+SA cascade approaches the equivalent noise temperature of the LNA itself.

5) The non-reflective termination should be disconnected. The generator output and the input of the spectrum analyzer should be directly connected.

6) On the FPC1500 analyzer, we should adjust:

- Center frequency: 836.5 MHz
- Reference level: 0 dBm
- SPAN: 20 MHz
- ATT: 20 dB

And at the generator output:

- CW
- Frequency: 836.5 MHz
- Level: 0 dBm

7) The RBW should be changed according to the table below, and the 3 dB bandwidth of the measured harmonic signal should be determined.

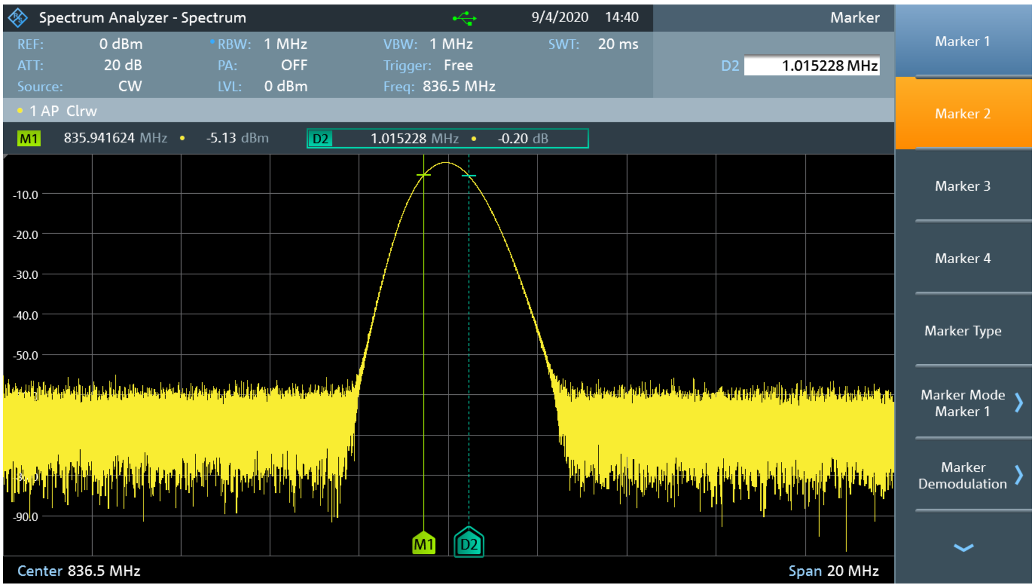

Figure 10.2: Example of measured response

Table 10.1: Measured bandwidths

RBW	10 Hz	100 Hz	1 kHz	10 kHz	100 kHz	1 MHz
$B_{-3\,\mathrm{dB}}$ (kHz)						

8) The module characteristics of the bandpass filter from the task @7 should be compared with the module characteristics of the RBW filter.

In the task focused on measuring the bandpass filter, the module characteristics of the filter shows iso-extremal ripples in the passband. The power module characteristics of the RBW filter of the SA shows just a single maximum.

A filter with the rectangular module characteristics is not causal, and therefore is not possible to be implemented. Filters with the quasi-elliptical approximation or the elliptical one show high selectivity, but the impulse response is long and the transient response (the response to the excitation by the unit pulse signal) has significant and long-decreasing overshoots.

This behavior would extend the sweeping length. Therefore, the filters with the Gaussian transmission characteristics are used as RBW filters. Very narrow filters can be implemented as digital ones.

11 Noise figure of the passive two-port network

11.1 Theoretical introduction

The task is related to the previous task. The exercise introduces a simplified procedure for measuring the noise figure of a two-port network. The role of the measured two-port network (DUT) is played by the bandpass filter that was already measured in the previous task. In order to measure the noise figure, a preamplifier is used. The accuracy of the method is not high (± 0.5 dB) but is sufficient for many applications.

The frequency filter or the attenuator from previous tasks are examples of linear passive two-port networks. The two-port network is linear if the input power does not produce other frequency products at the output (see previous tasks). The two-port network is passive if the output power is not higher than the input power.

11.1.1 Noise figure

Radio-electronic devices generally contain several sources of noise. The noise number F is introduced to effectively describe noise parameters of the system.

$$F = \frac{\frac{S_1}{N_1}}{\frac{S_2}{N_2}}$$

Here, S_1 is the signal power at the input of the two-port network, N_1 is the noise power of the non-reflective termination at the input, i.e. $N_1 = kT_0B$, S_2 is the power at the output of the two-port network and N_2 is the total noise power at the output. The equivalent power definition follows:

$$F = \frac{N_2}{GN_1} = 1 + \frac{N_a}{GN_1}$$

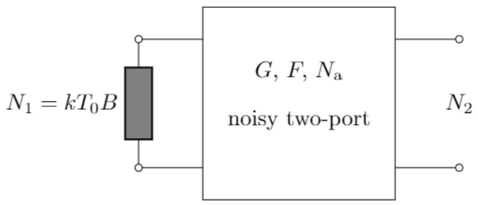

Figure 11.1: To define the noise figure

Here, GN_1 is the amplified noise of the non-reflective termination at the input port, G is the achievable gain of the measured two-port network and N_a is the basic noise of the measured two-port.

In order to determine F, the output power N_2 and the achievable gain G have to be measured at the same time. Two states of the noise power are usually used for this purpose: the basic $N_{1C} = kT_0B$ and the significantly higher noise power N_{1H}. This is the principle of the Hot-Cold method of measuring F. In this task, we use only $N_{1C} = kT_0B$ and an auxiliary generator that excites the measured DUT through the attenuator with the attenuation $L = 10\,dB$. The role of the measured two-port network is played by the bandpass filter measured in previous tasks.

Optionally, another linear two-port network can be measured by bridging the given bandpass filter.

11.2 Measurement

The generator should be set as follows:

- CW: $f = 836{,}5$ MHz
- $P = -30$ dBm.

An attenuator with attenuation $L = 10$ dB has to be connected to the generator output. To implement, the output is connected via the second attenuator in the down-converter. All other instructions assume this connection, and the output behind the attenuator is considered to be the generator output. This fact will no longer be notified.

The SA should be set as follows:

- RBW = 10 Hz,
- VBW = 300 kHz,
- ATT = 0 dB,
- SPAN = 20 kHz

- The attenuator (7) should be ensured to be inactive and to show an additive attenuation only.
- The X409 and X408 jumpers should be in the position 1.
- The generator output should be connected to the X404 SMA input.
- The SA RF input should be connected to the X405 SMA input.

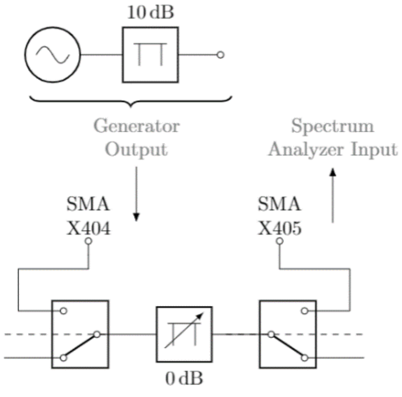

Figure 11.2: Wiring for measuring the output power of the generator

- The value of the power P_{in} should be read
- Position of the X409 jumper should be changed to the position 2

- The SA RF input should be disconnected from the X405 connector and should be connected to the X402 input
- The value of the power P_{PA} should be read

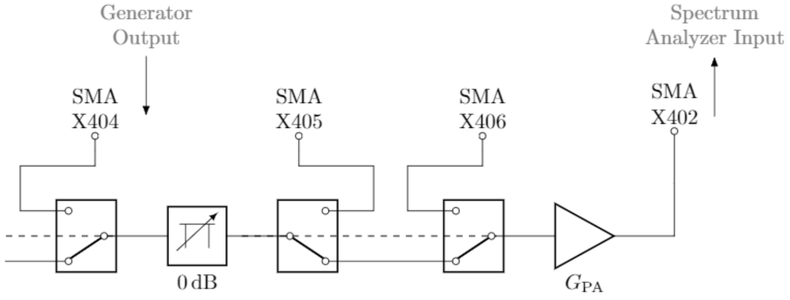

Figure 1111.3: Wiring for measuring

- The value of the gain should be determined

$$G_{\mathrm{PA_{dB}}} = P_{\mathrm{PA}} - P_{\mathrm{in}} \ \ (\mathrm{dB})$$

$$G_{\mathrm{PA}} = 10^{\frac{G_{\mathrm{PA_{dB}}}}{10}} \ \ (-)$$

- The output of the generator should be turned off, and the power $N_{\mathrm{PA_{dB}}}$ (dBm) should be read (the REF LEVEL should be changed accordingly)

$$N_{\mathrm{PA}} = 10^{-3} 10^{\frac{N_{\mathrm{PA_{dB}}}}{10}} \ \ (\mathrm{W})$$

- The generator output should be disconnected from the X404 input and should be connected to the X401 SMA input

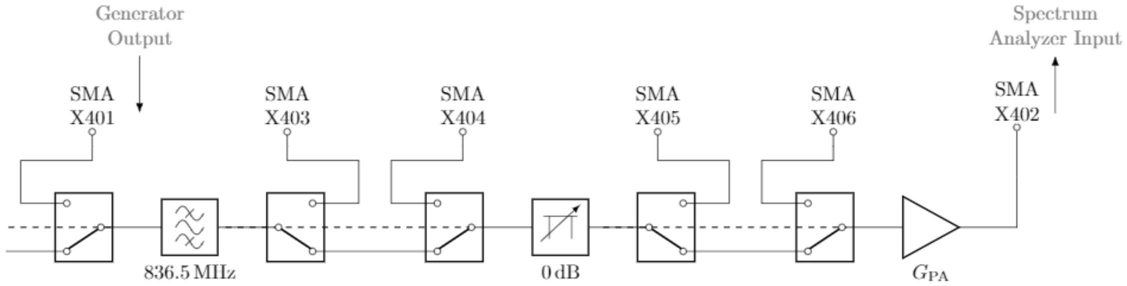

Figure 11.4: Wiring for measuring DUT

- Position of the X408 jumper should be changed to the position 2 and the X407 jumper should be ensured to be in the position 1

- The generator output should be turned on, and the DUT gain should be determined

$$G_{\mathrm{DUT_{dB}}} = P_{\mathrm{DUT}} - P_{\mathrm{PA}} \ \ (\mathrm{dB})$$

$$G_{\mathrm{DUT}} = 10^{\frac{G_{\mathrm{DUT_{dB}}}}{10}} \ \ (-)$$

- The generator output should be turned off, and the noise power $N_{\mathrm{DUT_{dB}}}$ (dBm) should be read (the REF LEVEL should be changed accordingly)

$$N_{\mathrm{DUT}} = 10^{-3} 10^{\frac{N_{\mathrm{DUT_{dB}}}}{10}} \ \ (\mathrm{W})$$

11.2.1 Evaluation

- The noise figure of the “preamplifier-SA” cascade should be determined

$$F_{\mathrm{PZ}} = \frac{N_{\mathrm{PA}}}{kT_0 B G_{\mathrm{PA}}}$$

- The noise figure of the “DUT-preamplifier-SA” cascade F_{K} should be determined

$$F_{\mathrm{K}} = \frac{N_{\mathrm{DUT}}}{kT_0 B G_{\mathrm{PA}} G_{\mathrm{DUT}}}$$

- The noise figure of the DUT should be determined F_{DUT}

$$F_{\mathrm{DUT}} = F_{\mathrm{K}} - \frac{F_{\mathrm{PZ}} - 1}{G_{\mathrm{DUT}}}$$

$$F_{\mathrm{DUT_{dB}}} = 10 \log F_{\mathrm{DUT}}$$

The $\boldsymbol{F_{\mathrm{DUT_{dB}}}}$ should be compared with the value of the insertion loss of the bandpass filter measured in the previous task.

The measurement is inaccurate but can be used to verify that the noise figure F of a passive element (an attenuator, a reactive passive filter, etc.) equals to the attenuation L in accordance with the power definition of the noise figure

$$F = \frac{N_2}{GN_1} = \frac{kT_0 B}{\frac{kT_0 B}{L}} = L$$

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12 References

12.1 Recommended books

- Pozar, David M., *Microwave Engineering*, 4th ed., Wiley 2011, ISBN 9780470631553
- Chen, Wai-Kai, *Passive, Active, and Digital Filters*, 2nd ed., CRC Press 2009, ISBN 9781420058857
- Maas, Stephen A., *Nonlinear Microwave and RF Circuits*, Artech House 2003, ISBN 9781580534840.