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A fast vectorial network analyser for frequencies up to 4 GHz

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Abstract. A fast four channel network analyser is introduced to measure S-parameters in a frequency range from 10 MHz to 4 GHz. The measurement period of this kind of network analyser is short in comparison with a conventional one, which becomes possible by using highly linear analogue frequency ramps, realized with fractional divider phase locked loop circuits. The settling time for the discrete frequency points, which is typical for a conventional network analyser, is nearly eliminated by the use of frequency ramps.

To achieve the necessary bandwidth of 4 GHz in the baseband, a heterodyne principle is applied. A VCO-signal from 5 to 9 GHz is down converted to the baseband with a fixed frequency generator of 9 GHz. The structure of the following network analyser is similar to the structure of a conventional four channel network analyser and therefore all known calibration techniques are suited for the fast network analyser.

The intermediate frequency, which is necessary for the evaluation, carries the required information. However a special analysis technique is necessary. This technique is based on two Fourier transformations.

1 Introduction

Special applications of metrology often require fast measurements of complex-valued scattering parameters. Such applications are for example a high volume production in the industry, impulse measurements or systems in motion, where the S-parameters are changing very fast. A measurement system for these applications can be based on highly linear analogue frequency ramp systems. The advantage of these ramp systems is that the settling time of the discrete frequency points is nearly eliminated. This settling time is typical for a conventional network analyser. The requirements of linearity and stability for these ramps are very high. These requirements can be fulfilled by the use of fractional divider phase locked loop circuits. For a detailed description of these fractional divider phase locked loop circuits see (Musch and Schiek, 1997).

In conventional systems every frequency point is evaluated separately. Due to the fact that now discrete frequency points are not available the whole frequency sweep has to be recorded completely. Therefore a special analysis technique is necessary. This special analysis technique is based on two Fourier transformations and the analytical signal. Generally the fast network analyser must be able to perform calibrated measurements.

2 Structure of the fast network analyser

The structure of the fast network analyser is similar to a conventional four channel network analyser and can be divided into two parts. The first part comprises the signal generation, which is shown in Fig. 1, and the second part comprises the measuring part, which is shown in Fig. 2.

The fast network analyser is designed according to the heterodyne principle, which is based on two signals of slightly different frequencies. These two signals are generated by the VCO1 with the frequency f_1 and the VCO2 with the frequency f_2 in Fig. 1, whereby both oscillators are stabilised with a PLL system. These two ramp systems generate frequencies from 8.99 GHz to 5 GHz with a ramp duration of 20 ms. Between both frequency ramps a fixed frequency difference of 50 kHz exists. For a baseband system frequencies are required from 10 MHz up to 4 GHz. Therefore both frequency ramps are down converted with a fixed frequency of f_3=9 GHz to lower frequencies. All three PLL systems have the same clock of 12.5 MHz, which is the crystal oscillator XCO in Fig. 1. The required signals with a frequency difference of 50 kHz are amplified. In Fig. 2 the measurement signal f_m is led through a high-frequency switch to the two couplers. From there the signal is divided to the device under test (DUT) and to the mixers. At the mixers the signals are down-converted with the L.O.-signal f_r to the low

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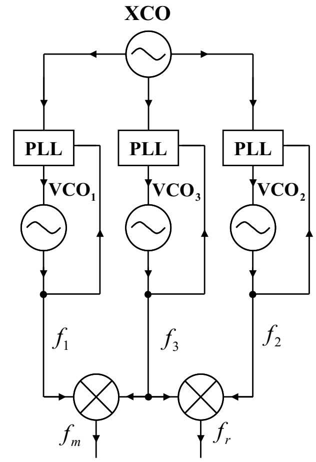

Fig. 1. Block diagram of the signal generation.

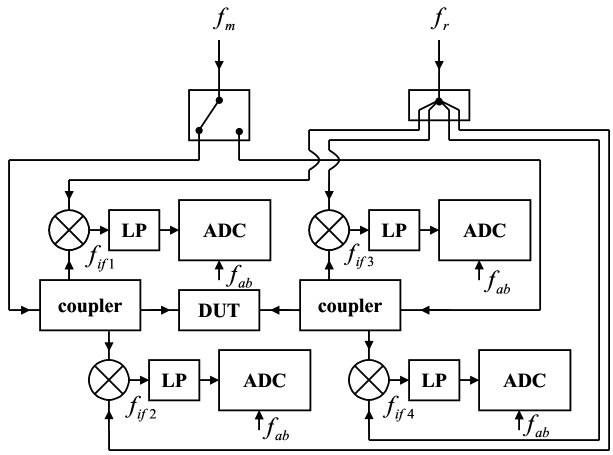

Fig. 2. Block diagram of the fast network analyser.

intermediate frequency f_{if}=50 kHz. Isolation amplifiers, not shown in the simplified diagram, are placed at both inputs of the mixers to prevent a cross-talk of the signals. The intermediate frequency f_{if} is filtered by a low pass filter and digitised with an analog digital converter. The recorded data are sent to a PC with an IEEE-interface. The data processing is performed with this PC.

It is very important for the fast network analyser that the recording of the frequency sweep is reproducible, because in the data processing it is necessary to establish relations between different data sets. For this reason the clock for the analog digital converter is the same crystal oscillator as for the three PLL systems in Fig. 1. Another important point in order to guarantee reproducibility is that the starting time of the frequency ramps and the starting time of the analog digital converters are always the same.

The measurement period of the fast network analyser is 40 ms. This time is caused by two switch settings, which are necessary for one measurement. Each frequency ramp takes 20 ms.

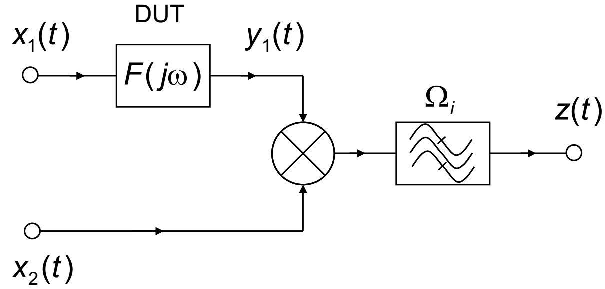

Fig. 3. Model of a measuring port.

The structure of the fast network analyser is similar to the structure of a conventional four channel network analyser. Thus all known calibration techniques like self-calibration can be used with the fast network analyser.

3 Special analysis technique

The data processing of the fast network analyser differs from the data processing of a conventional network analyser. For the fast network analyser a whole frequency sweep is recorded completely in contrast to the conventional one, where every single frequency point is evaluated separately. So a special analysis technique is necessary and this technique is based on the continuous sampling.

To explain this special technique a measurement channel is shown in Fig. 3 as a mathematical model.

For the data processing it is necessary to obtain the frequency response $F(j\omega)$ of the device under test. For this purpose the two signals $x_1(t)$ and $x_2(t)$ for a continuous sampling can be defined as

$$x_1(t) = \mathrm{Re}\{e^{j\Phi_1(t)}\} \tag{1}$$

$$x_2(t) = \mathrm{Re}\{e^{j\Phi_2(t)}\}. \tag{2}$$

The phases in these equations are

$$\Phi_1(t) = \Omega_0 t + \pi\sigma t^2 \tag{3}$$

$$\Phi_2(t) = (\Omega_0 - \Omega_i)t + \pi\sigma t^2. \tag{4}$$

The ramp steepness σ in these phase equations is defined as the ratio of bandwidth to rise time:

$$\sigma = \frac{\Delta F}{T_m}. \tag{5}$$

The signal $z(t)$ is required to get the frequency response $F(j\omega)$. Therefore the signal $y_1(t)$ must be obtained first, which is linked with the signal $x_1(t)$ via the convolution:

$$y_1(t) = \int_{-\infty}^{\infty} f(\tau)x_1(t-\tau)\mathrm{d}\tau$$

$$= \int_{-\infty}^{\infty} f(\tau)\mathrm{Re}\{\mathrm{e}^{j\Phi_1(t-\tau)}\}\mathrm{d}\tau$$
$$= \mathrm{Re}\left\{ \int_{-\infty}^{\infty} f(\tau)\mathrm{e}^{j\Phi_1(t-\tau)}\mathrm{d}\tau \right\}. \quad (6)$$

Here $f(t)$ is the pulse response of $F(j\omega)$:

$$F(j\omega) = \int_{-\infty}^{\infty} f(t)\mathrm{e}^{-j\omega t}\mathrm{d}t. \quad (7)$$

Since $f(t)$ is the pulse response of a real system, the real part operator of $x_1(t)$ can be placed in front of the integral. For the exponential function in Eq. (6) we get several factors:

$$\mathrm{e}^{j\Phi_1(t-\tau)} = \underbrace{\mathrm{e}^{j\Omega_0 t} \cdot \mathrm{e}^{j\pi\sigma t^2}} \cdot \underbrace{\mathrm{e}^{-j\Omega_0 \tau} \cdot \mathrm{e}^{-j2\pi\sigma t\tau}} \cdot \mathrm{e}^{j\pi\sigma\tau^2}$$
$$= \mathrm{e}^{j\Phi_1(t)}\mathrm{e}^{-j\Omega(t)\tau}\mathrm{e}^{j\pi\sigma\tau^2}. \quad (8)$$

Inserting this relation into equation 6 yields:

$$y_1(t) = \mathrm{Re}\{\mathrm{e}^{j\Phi_1(t)} \int_{-\infty}^{\infty} f(\tau)\mathrm{e}^{j\pi\sigma\tau^2}\mathrm{e}^{-j\Omega(t)\tau}\mathrm{d}\tau\}. \quad (9)$$

Now a simplification is necessary. Due to the fact that only short pulse responses are allowed, τ is nearly zero. Therefore the expression $\mathrm{e}^{j\pi\sigma\tau^2}$ is approximately 1. Therefore this expression can be neglected and a new equation for $y_1(t)$ is obtained:

$$y_1(t) = \mathrm{Re}\{\mathrm{e}^{j\Phi_1(t)} \int_{-\infty}^{\infty} f(\tau)\mathrm{e}^{-j\Omega(t)\tau}\mathrm{d}\tau\}$$
$$= \mathrm{Re}\{F(j\Omega(t))\mathrm{e}^{j\Phi_1(t)}\}. \quad (10)$$

The required signal $z(t)$ is received by mixing or multiplying the signals $y_1(t)$ and $x_2(t)$:

$$\begin{aligned} z(t) &= y_1(t) \cdot x_2(t) \\ &= \mathrm{Re}\{F(j\Omega(t))\mathrm{e}^{j\Phi_1(t)}\} \cdot \mathrm{Re}\{\mathrm{e}^{j\Phi_2(t)}\} \\ &= \mathrm{Re}\{F(j\Omega(t))[\cos(\Phi_1(t)) \\ &\quad + j\sin(\Phi_1(t))]\} \cdot \mathrm{Re}\{\cos(\Phi_2(t)) + j\sin(\Phi_2(t))\} \\ &= \mathrm{Re}\{F(j\Omega(t))[\cos(\Phi_1(t)) \\ &\quad + j\sin(\Phi_1(t))]\cos(\Phi_2(t))\} \\ &= \mathrm{Re}\{F(j\Omega(t))[\cos(\Phi_1(t))\cos(\Phi_2(t)) \\ &\quad + j\sin(\Phi_1(t))\cos(\Phi_2(t))]\} \\ &= \frac{1}{2}\mathrm{Re}\{F(j\Omega(t))[(\cos(\Omega_i t) + \cos(2\Omega_0 t - \Omega_i t + 2\pi\sigma t^2)) \\ &\quad + j(\sin(\Omega_i t) + \sin(2\Omega_0 t - \Omega_i t + 2\pi\sigma t^2))]\}. \end{aligned} \quad (11)$$

After this signal has been filtered at Ω_i and the sum terms for higher frequencies are cancelled, the following equation is obtained:

$$\begin{aligned} z(t) &= \frac{1}{2}\mathrm{Re}\{F(j\Omega(t))[\cos(\Omega_i t) + j\sin(\Omega_i t)]\} \\ &= \frac{1}{2}\mathrm{Re}\{F(j\Omega(t))\mathrm{e}^{j\Omega_i t}\} \\ &= \frac{1}{2}\mathrm{Re}\{F(j\Omega(t))\mathrm{e}^{j\Phi_i(t)}\} \end{aligned}$$

with

$$\Phi_i(t) = \Omega_i t. \quad (12)$$

In order to obtain the frequency response of the device under test, the analytical signal $z_+(t)$ is generated. The following expression holds:

$$z_+(t) = \frac{1}{2}F(j\Omega(t))\mathrm{e}^{j\Phi_i(t)}. \quad (13)$$

This expression contains the wanted frequency response of the device under test modified with the exponential function $\mathrm{e}^{j\Phi_i(t)}$. In order to get the transfer function $F(j\Omega(t))$ this exponential function can be eliminated by building ratios of different analytical signals, because this exponential function is the same function for all analytical signals and therefore will cancel. These ratios between different analytical signals are needed anyway to calculate the S-parameters. An example for the transmission coefficient S_{21} is the ratio between the wave, which is transmitted through the device under test, and the wave which is incident to the device under test. Therefore this evaluation method is an easy way to calculate S-parameters.

4 Measurements

A comparison between a conventional and the fast network analyser was done to demonstrate the functionality of the fast network analyser. For the comparison a commercial network analyser (ZVR) was used. First, a bandpass filter at 2.85 GHz, designed in microstrip technique was used as a device under test. For this measurement the conventional network analyser has a record time of 200 ms with 501 points and full i.f. bandwidth. In contrast the fast network analyser has a record time of 40 ms with 2500 points. For both network analysers a TLR-calibration (Through, Line, Reflect), which belongs to the self-calibration techniques, was performed with coaxial standards. The magnitude and the phase of the transmission coefficient S_{21} is shown in Figs. 4 and 5 for a frequency range from 1 GHz to 4 GHz.

The results of the measurements demonstrate the functionality of the fast network analyser. The values for the magnitude and for the phase are nearly identical. The only difference appears in the noise performance of the network analyser. The values of the fast network analyser are somewhat noisier than the values of the conventional one. This is shown in Fig. 4. The reason is a lower dynamic range of the fast network analyser in comparison to the conventional one.

The Figs. 6 and 7 show the magnitude and the phase of the reflection S_{11} of the bandpass filter measured with the fast network analyser and with the ZVR respectively.

The values for the magnitudes and for the phases of both network analysers are again nearly identical.

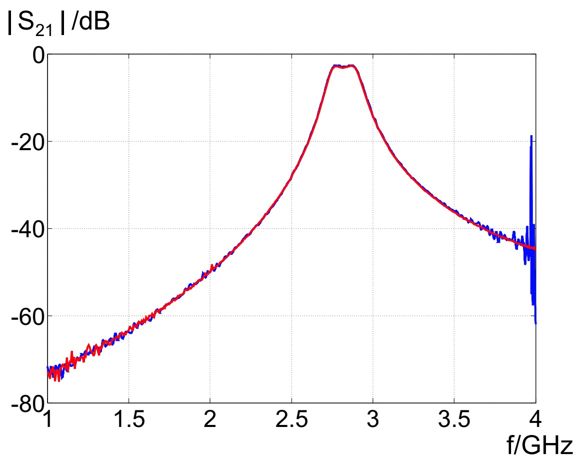

Fig. 4. Magnitude of S_{21} of a 2.85 GHz bandpass filter, fast network analyser (blue), commercial network analyser (ZVR)(red).

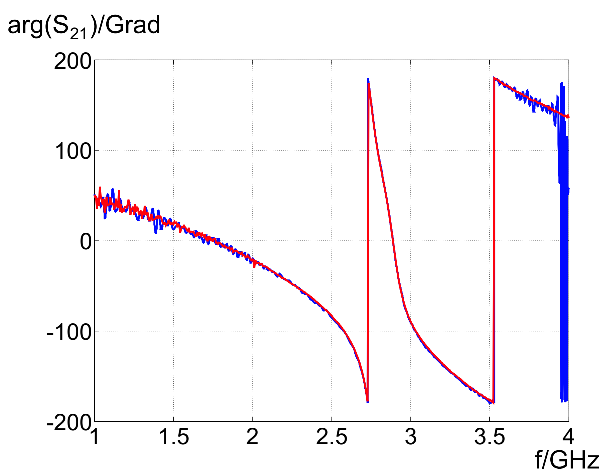

Fig. 5. Phase of S_{21} of a 2.85 GHz bandpass filter, fast network analyser (blue), commercial network analyser (ZVR)(red).

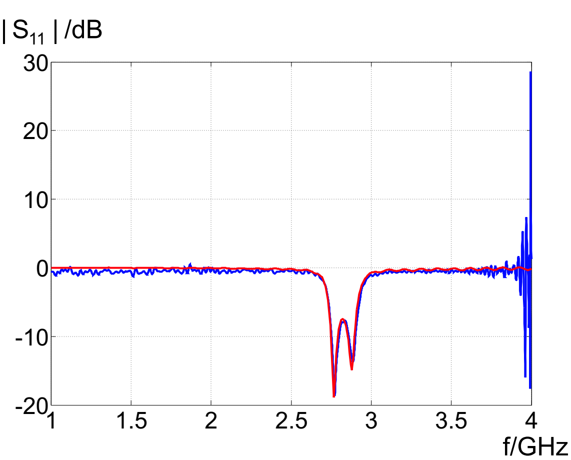

Fig. 6. Magnitude of S_{11} of a 2.85 GHz bandpass filter, fast network analyser (blue), commercial network analyser (ZVR)(red).

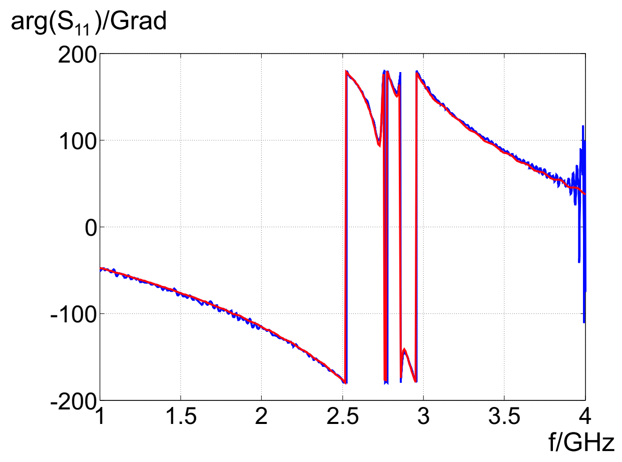

Fig. 7. Phase of S_{11} of a 2.85 GHz bandpass filter, fast network analyser (blue), commercial network analyser (ZVR)(red).

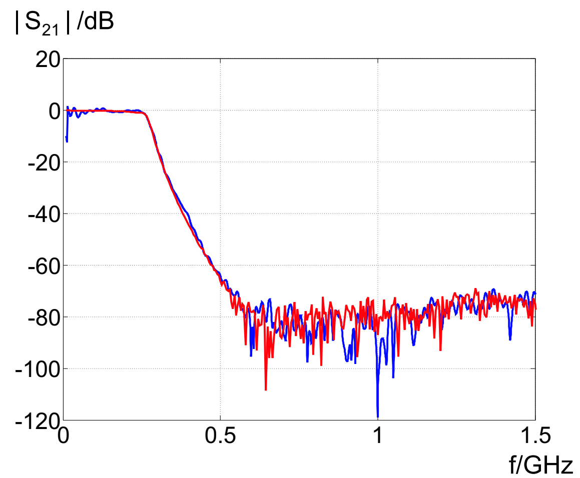

Fig. 8. Magnitude of S_{21} of a 300 MHz lowpass filter, fast network analyser (blue), commercial network analyser (ZVR)(red).

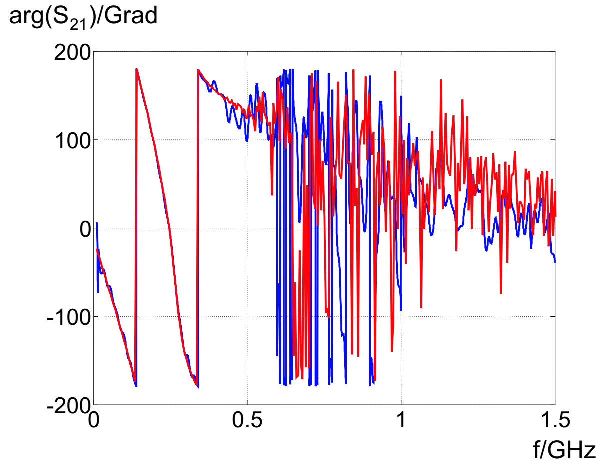

Fig. 9. Phase of S_{21} of a 300 MHz lowpass filter, fast network analyser (blue), commercial network analyser (ZVR)(red).

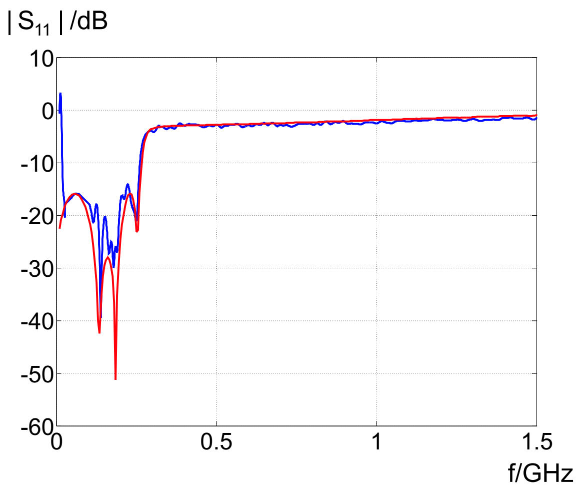

Fig. 10. Magnitude of S_{11} of a 300 MHz lowpass filter, fast network analyser (blue), commercial network analyser (ZVR)(red).

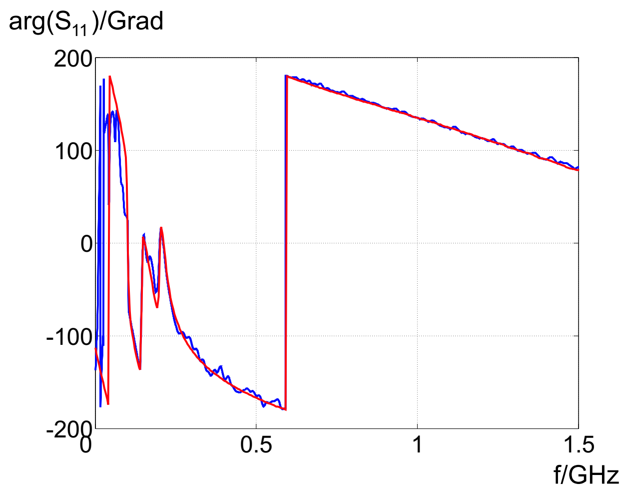

Fig. 11. Phase of S_{11} of a 300 MHz lowpass filter, fast network analyser (blue), commercial network analyser (ZVR)(red).

As a second device under test a lowpass filter with a cut-off frequency of 300 MHz was chosen, in order to show the functionality of the fast network analyser for lower frequencies. For the comparison the same conventional network analyser, the ZVR, was used. For this measurement the conventional network analyser has a record time of 200 ms with 501 points. The fast network analyser has a record time of 40 ms with 2500 points by comparison. Both network analysers were calibrated with coaxial standards, i.e. TOM (Through, Open, Match). In the Figs. 8 and 9 the magnitude and the phase of the transmission coefficient S_{21} for a frequency range from 10 MHz to 1.5 GHz are displayed.

The curves in both figures show the good performance of the fast network analyser as in the figures before. Again the values for the magnitude and for the phase are nearly identical.

In the Figs. 10 and 11 the magnitude and the phase of the reflection coefficient S_{11} are shown, respectively. The measurement values for the magnitude as well as for the phase are nearly identical.

Thus the functionality of the fast network analyser as regards the structure of the analyser, the special analysis technique and the possibility to apply conventional calibration techniques is demonstrated.

5 Conclusion

A fast network analyser has been introduced for frequencies from 10 MHz up to 4 GHz. The system is realised with two highly linear analogue frequency ramps which are based on fractional divider phase locked loop circuits. This analyser has the same structure as a conventional four channel network analyser. Thus all known calibration techniques can be used. The analysis technique differs from the conventional one, because the whole data record of a frequency sweep is evaluated at once and not frequency point by frequency point. For this reason the analytical signal must be determined to acquire the S-parameters.

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