CALIBRATION OF SHORT RANGE FMCW-RADARS WITH NETWORK ANALYZER CALIBRATION TECHNIQUES

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Abstract

FMCW-radars can be used as sensors in the short range, in order to locate and determine reflections of targets by amplitude and phase. Such sensors have to be accurate, fast and inexpensive. The paper describes, how these goals can be reached by applying a single channel system and establishing complex measurement capability. This is done by using the Hilbert transform and by removing the time domain system error using techniques similar to those applied for vector network analyzers. Details of the system implementation are given and results are compared to network analyzer measurements.

1 Introduction

FMCW-radars can be employed in the short range for sensor applications. Using this type of radar as a sensor it is often necessary to determine phase ϕ and amplitude r of unknown short range reflections of a device under test (D.U.T.) very fast. For industrial applications the system also must have a simple design to ensure low manufacturing costs. Conventional FMCW-radars do not show these features inherently.

A typical setup for a short range FMCW-radar is sketched in fig. 1. In the literature this type of radar is often called a reflectometer and the description is well known (e. g. [Som72]). The system can be seen as a homodyne network analyzer. The output signal of this analyzer is complex and sampling of two signals is necessary. Due to the different path length of the reference and the measurement path the swept frequency generates low frequency output signals. If the radar is calibrated, it is possible to obtain the complex reflection coefficients r_i as a function of their locations. The locations are coded in the low frequency components f_i of the output signal $y(t)$

$$y(t) = \sum_{i=1}^{M} |r_i| \cdot e^{j(2\pi \cdot f_i \cdot t + \phi_i)} \quad (1)$$

$$\text{with} \quad f_i = \frac{B}{T_S} \cdot (t_i - t_r) \quad .$$

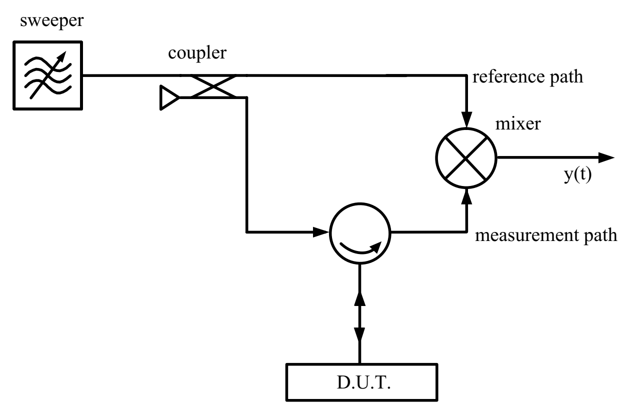

Fig. 1: FMCW-radar for short range applications

The FMCW-radar used in this work has a bandwidth B of $4.4\,G$Hz in the X-band and a sweep time T_S of $16\,m$s. Using this system two main problems occur. First, the receiving mixer is normally a quadrature receiver which can cause problems especially when applied over a broad bandwidth. The second problem is to calibrate the reflectometer with known standards in order to remove the system error.

The quadrature receiver error can be removed in theory [Neu88] but this is a complicated task. As a better solution the Hilbert transform is applied to calculate the complex data $y(t)$ from the real output signal of only one sampled mixer output. This simplifies the system and makes it fast and cheap. The calibration of the system error is performed with techniques known from network analyzer techniques.

2 Hilbert transform theory

The signal $y(t)$ is separated into a real $r(t)$ and an imaginary part $i(t)$ (for simplicity only one reflection with frequency $f_i = f_R = \omega_r/2\pi$ is considered)

$$y(t) = r(t) + j \cdot i(t) \quad . \quad (2)$$

In frequency domain the imaginary part $I(j\omega)$ can be calculated from the real part $R(j\omega)$ by multiplication with the

Hilbert transform operator $H_H(j\omega)$

$$\begin{aligned} I(j\omega) &= \hat{R}(j\omega) = H_H(j\omega) \cdot R(j\omega) & (3) \\ \text{with } H_H(j\omega) &= -j \cdot sgn(\omega) & (4) \end{aligned}$$

Transforming this result back into time domain, the multiplication in frequency domain is transformed into a convolution (denoted by the symbol $*$),

$$i(t) = \hat{r}(t) = r(t) * h_h(t) \quad . \tag{5}$$

A problem in using the Hilbert transform is the error $\Delta i(t)$ between the original and the calculated imaginary part

$$\Delta i(t) = i(t) - \hat{r}(t) \quad , \tag{6}$$

which must be minimized. This error occurs due to the limited sampling time T_S, which corresponds to a rectangular time window ($rect(t)$). Due to this window, the signal has spectral components not only at the frequencies $\pm\omega_r$, but also at higher and lower frequencies (Fourier transform $rect(t)$ corresponds to $\text{si}(j\omega) = \sin(j\omega)/j\omega$). This causes problems in the Hilbert transform operator $H_H(j\omega)$, if the components near zero frequency have significant amplitudes.

In this paper several methods are applied to minimize this error, which improve the Hilbert transform and can be implemented easily.

- By using a long delay line in the measurement path, the spectral component ω_r can be increased and the error decreased. Negative components of the $\text{si}(\omega - \omega_r)$ spectrum are minimized by this approach.

- Due to the fact that the main deviations occur at the edges of the used frequency band, an improvement can be obtained by increasing the measurement bandwidth slightly. Although the components of the radar system may not work well in this larger frequency range, it is possible to decrease the error. After transformation the original bandwidth can be used for further calculations.

- A Kaiser-window prior to the Hilbert transform and inverse filtering afterwards is applied [Lip88]. The Kaiser-window is defined as

  $$h_W(t) = \frac{I_0\left(\beta \cdot \sqrt{1 - [\frac{t - T_S/2}{T_S/2}]^2}\right)}{I_0(\beta)} \tag{7}$$

  in the time region $0 \le t \le T_S$ and $h_W(t) = 0$ elsewhere. Using a parameter of $\beta = 15$, optimal solutions are obtained.

Combining the above steps, it is possible to improve the accuracy significantly. An error still occurs at the band edges but is extremely low as can be shown by system simulations (fig. 2).

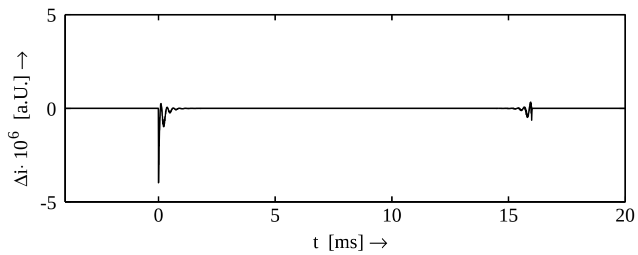

Fig. 2: Error $\Delta i(t)$, combination of higher modulation frequency, wider bandwidth and Kaiser-window with $\beta = 15$

3 Calibration of system errors

Let us assume a D.U.T. is connected directly to the calibration plane. Following from eq. 1 the output signal in this case must be

$$y(t) = |r| \cdot e^{j \cdot \phi} \quad . \tag{8}$$

Several errors may interfere this signal. Normally the directivity of the circulator used to separate the reflected from the transmitted wave is low. Therefore an error term in the output signal occurs (index D due to directivity), which can be described as

$$y_D(t) = |r_D| \cdot e^{j(2\pi \cdot f_D \cdot t + \phi_D)} \quad . \tag{9}$$

Different path lengths and attenuations (index Δ due to difference) between the measurement and the reference path result in

$$y_\Delta(t) = |r_\Delta| \cdot e^{j(2\pi \cdot f_\Delta \cdot t + \phi_\Delta)} \quad . \tag{10}$$

Due to mismatch between the D.U.T. and the radar system some of the backscattered signal is reflected back to the D.U.T. (index R due to reflection)

$$y_R(t) = |r_R| \cdot e^{j \cdot \phi_R} \quad . \tag{11}$$

This causes multiple reflections between the D.U.T. and the measurement system. The measured signal (index M) disturbed by the three errors described above may be written as:

$$\begin{aligned} y_M(t) &= y_D(t) + y_\Delta(t) \cdot y(t) \cdot \Big(1 + \cdots \\ &\quad + y(t) \cdot y_R(t) + [y(t) \cdot y_R(t)]^2 + \cdots \\ &\quad + [y(t) \cdot y_R(t)]^3 + \cdots \Big) & (12) \\ y_M(t) &= y_D(t) + \frac{y_\Delta(t) \cdot y(t)}{1 - y(t) \cdot y_R(t)} & (13) \\ &\quad \text{for } |y(t) \cdot y_R(t)| < 1 \end{aligned}$$

$y(t)$ is derived from the measured signal $y_M(t)$. Eq. 13 when rearranged becomes:

$$y(t) = \frac{y_M(t) - y_D(t)}{y_\Delta(t) - y_R(t) \cdot \big(y_M(t) - y_D(t)\big)} \tag{14}$$

If the error terms are known it is possible to solve this equation. A basic idea of this paper is to compare this description of a time dependent signal with the known error model for vector network analyzer calibration (e. g. [Bry93]).

A vector network analyzer is calibrated by using an error-term flowgraph for the frequency signal (fig. 3). The reflection r, which is a function of frequency, can be calculated using the well known formula

$$r = \frac{r_M - R_{11}}{R_{21}R_{12} - R_{22}(r_M - R_{11})} \quad , \tag{15}$$

using the measurement of three known standards (open, short and matched load).

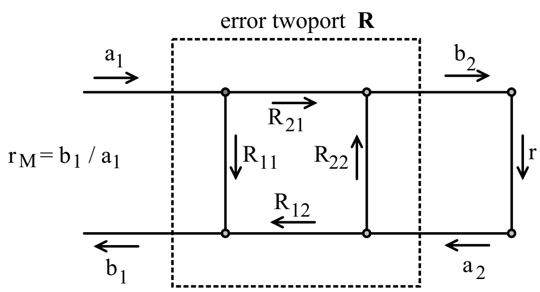

Fig. 3: Error-term flowgraph for one port vector network analyzers; a and b are incident and reflected waves

A comparison of eq. 15 and eq. 14 delivers the relationships

$$\begin{aligned} y_D(t) &\;\hat{=}\; R_{11} \\ y_\Delta(t) &\;\hat{=}\; R_{21} \cdot R_{12} \\ y_R(t) &\;\hat{=}\; R_{22} \quad . \end{aligned}$$

It must therefore be possible to calibrate the reflectometer by employing the same three calibration standards used in vector network analyzer calibration. It is merely required to transform the description of the standards from a frequency dependent into a time dependent one. This can be done easily if one takes into account that the start frequency of the used frequency band corresponds with the start time and the stop frequency with the stop time. In our case a time of $t = 0$ ms corresponds to the frequency $f = 8\,G$Hz, and a time of $t = T_S = 16$ ms corresponds to the frequency $f = 12.4\,G$Hz. To ensure an accurate relationship, these correspondences have to be calibrated. Employing waveguide resonators with resonant frequencies of $f_{res} = 8\,G$Hz and $f_{res} = 12.4\,G$Hz, it is possible to calibrate the start and stop frequencies.

4 Results

To show the accuracy of the system error calibration under application of the Hilbert transform, measurements are

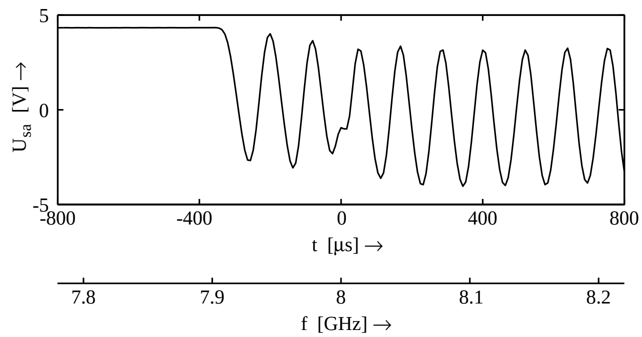

Fig. 4: Waveguide resonator with resonant frequency of $f_{res} = 8\,G$Hz, sampled signal U_{sa} as a function of time and frequency , detailed view

made on a third waveguide resonator (resonant frequency $f_{res} = 10.2\,G$Hz) with a short range FMCW-radar and for comparison with a vector network analyzer (index HP due to the used HP 8510C). Both system are calibrated with the same calibration standards[1]. Excellent results are obtained

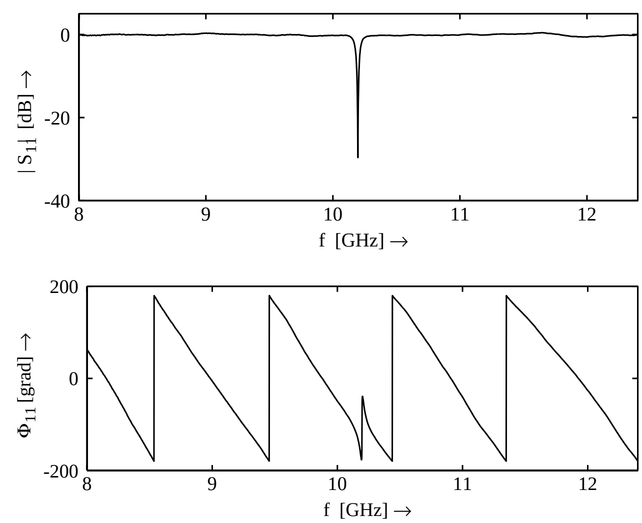

Fig. 5: Calibrated amplitude and phase S_{11} of a resonator $f_{res} = 10.2\,G$Hz

for the phase and amplitude of the reflection coefficient S_{11} of this example (fig. 5). The error of the reflection coefficient S_{11} has a low amplitude $\Delta S_{11} = S_{11} - S_{11}^{\mathrm{HP}} \leq -25$ dB and phase $\Delta\Phi_{11} = \Phi_{11} - \Phi_{11}^{\mathrm{HP}} \leq \pm 3°$ (fig. 6). It is possible to show that these errors are due to imperfections of the fast sweeping radar and not the calibration.

To ensure that the error is not due to the Hilbert transform, we perform a second measurement with the HP 8510C. Although this is a heterodyne analyzer, the two output si-

[1] manufacturer: Rosenberger Hochfrequenztechnik 3.5mm Calibration Kit 50 Ω, Typ 03CK10A-150, Ser. No. H2501

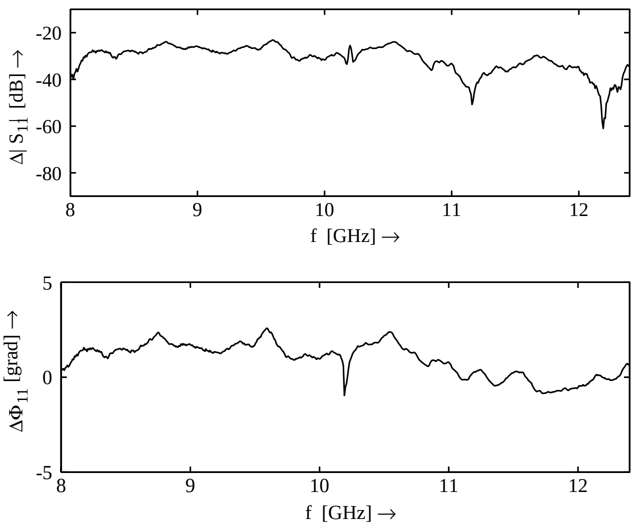

Fig. 6: Error of amplitude and phase S_{11} compared to HP 8510C measurement, resonator at $f_{res} = 10.2\,G$Hz

gnals have the same behaviour as the signals of a homodyne system. We use this analyser because the accuracy of the quadrature output signals is known and very high. Again the waveguide resonator with the resonant frequency $f_{res} = 10.2\,G$Hz is measured and analyzed in the normal way by using the quadrature signals and alternatively by applying the Hilbert transform with only one sampled signal (index H). Both versions are calibrated afterwards. The error $\Delta S_{11}^{H} = S_{11}^{H} - S_{11}$ is shown in fig. 7. It is lower than $\Delta|S_{11}^{H}| = -40\,$dB over the whole frequency band. The phase error, which is not shown here, is better than $\Delta\Phi_{11}^{H} = \Phi_{11}^{H} - \Phi_{11} = \pm 1°$.

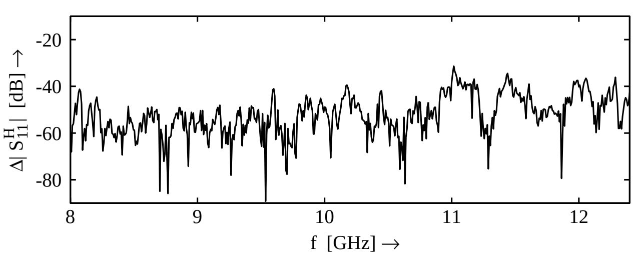

Fig. 7: Error in amplitude between S_{11}^{H} and normal amplitude, resonator $f_{res} = 10.2\,G$Hz

The dynamic range of the FMCW-radar is demonstrated by making measurements on various shorted attenuators. This measurement setup simulates absorbing layers in free space, which are terminated by short circuits. Five different attenuations from 0 dB up to 40 dB are measured with the FMCW-radar and the vector network analyzer. Again excellent agreement is obtained.

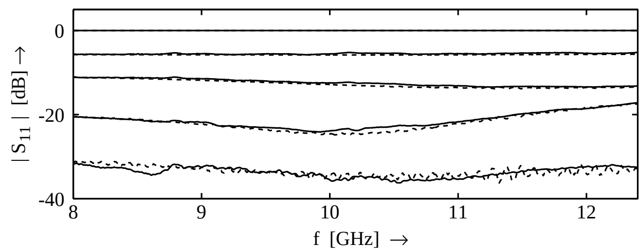

Fig. 8: Shorted attenuators, attenuation from the top to the bottom: 0 dB, 3 dB, 6 dB, 10 dB and 20 dB, S_{11} (———), HP 8510C S_{11}^{HP} (– – – –)

5 Conclusions

This paper has presented an accurate method to calibrate short range FMCW-radars using simple, well known calibration schemes from vector network analyzer techniques. The known calibration in the frequency domain can easily be applied on the time dependent output signal of reflectometers. It has been shown that the Hilbert transform offers the possibility of speeding up and simplifying the system using a one channel design to obtain vector network measurements.

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