Lab course RF Engineering
Lab 5: Planar Circuits

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1 Introduction

Planar transmission lines are used in the RF frequency range above some GHz and consist of flat conducting metal plains superimposed on a dielectric carrier substrate material.

Planar circuits consist of transmission line structures whose geometry was chosen ensuring the desired electrical behavior. These circuits are, e.g. steps in width or gaps of transmission lines in the most simplest cases, but usually more complex structures like filters, couplers and so on are subsumed in this circuit family. In the case of inserted lumped elements like resistances, capacitances, diodes, transistors, the circuits become hybrid circuits, so-called MIC(Microwave Integrated Circuit).

Planar transmission lines are together with the (hollow) waveguides—the most important RF transmission lines and have displaced the latter in the last decades in many application areas. The advantages are low weight, easy fabrication (almost the same procedure is used as in the well-known fabrication of PCBs, printed circuit boards), and connected with this is low production costs. Base materials used for these circuits are almost exclusively either synthetic materials like PTFE = Polytetrafluorethylene (Teflon), ceramics like aluminum oxide ceramics (Al₂O₃), or semi-isolating (high-ohmic) semiconductors like Si or GaAs. Using semiconducting materials it is possible to include active devices with appropriate semiconductor doping; the planar lines made in this technique will be evaporated on the substrate. These kinds of circuits are so-called MMICs (MMIC=Monolithic Microwave Integrated Circuit).

Drawbacks of planar circuits in comparison to appropriate circuits in waveguide technique are higher losses and therefore, lower quality factors of filters can be achieved. Additionally, the upper limit of transported power is much lower due to the very thin layers, the relatively high losses and the poor maximal electric strength due to the small distance between the transmission lines.

2 Planar Waveguides

There are lots of possibilities for building planar lines consisting of one or more dielectric layers and thin conducting strips; in Fig. 1(a) to (d) cross sections of the technical most important transmission line types are shown. The first drawing shows a microstrip line (MSL), up to now the most widespread type of planar line, used in MICs as well as in MMICs. It is a twofold bounded structure which means there is no cut-off frequency.

In Fig. 1(b) the cross section of a coplanar waveguide is shown. In contrast to the microstrip line, the ground plains are beside the hot wire. That allows the insertion of series and parallel elements as well on one layer side without via holes (holes through the substrate). This property is very attractive for the application in MMICs. Since the ground plane is divided into two parts, air bridges (e.g. by bonding) are used providing the same potential on both ground layers which avoids the odd quasi-TEM mode.
In Fig. 1(c) a **slot line** is shown. Generally, both layers beside the slot are on the same potential using air bridges which leads to a single bounded structure like the hollow waveguide and therefore, having a cut-off frequency as well. Without air bridges the fundamental mode is a quasi-TEM mode. The electromagnetic field is guided mainly in the slot; to improve the concentration of the fields in the slot a substrate material having a high relative dielectric constant is used ($\varepsilon_r \approx 10$).

Finally, Fig. 1(d) shows a **finline**, which may be considered as a shielded slot line. In contrast to the latter transmission line, the permittivity is in the range ($\varepsilon_r = 2 \ldots 3$). Again it is a single bounded structure having a cut-off frequency as well. The main application of such transmission lines is in the connection to the rectangular waveguide at frequencies starting at about $f \approx 30$ GHz. In other words: the finline is suited if there is a need for connecting waveguides and planar transmission lines.

About the item **mode**: An infinite number of solutions of the Maxwell’s equations exists for a transmission line. Theses solutions are called modes. At a given geometry, material parameters and frequency, there is only a finite number of modes which are propagating, i.e., are able to transport active power. On a waveguide consisting of $n$ separate metallizations there are $n-1$ modes of the TEM or quasi-TEM type, not having a cut-off frequency.

### 2.1 Microstrip Line

The electromagnetic field patterns on all planar transmission lines are partly in the air, and partly in the dielectric substrate. Due to the different electrical properties, the mode should be traveling with different phase velocities in each region. This is compensated for by the existence of additional longitudinal components of the electromagnetic field beside the transversal components. Waves having both electric and magnetic field components in the direction of the power flow are called **hybrid waves**. The fundamental mode on the microstrip line has—according to the explanation made above—all six field components, but in the technical
frequency range, the longitudinal components are negligible in comparison to the transversal ones. Thus, this mode is called **quasi-TEM mode**. A pure TEM(Transversal ElectroMagnetic) mode like the fundamental mode on a homogeneously filled coaxial cable has only transversal components.

![Quasi TEM fundamental mode](image1)

**Fig. 2**: Transversal electric field pattern of the fundamental mode and the first higher order mode on a microstrip line.

In Fig. 2(a) the transversal electric field of the fundamental mode is shown. There is an infinite number of higher order modes which are propagating with increasing frequency, but in the frequency range of technical interest, all modes except the fundamental one are **evanescent modes**, i.e. under cut-off and are only able to store energy, but not to propagate. Figure 2(b) shows the transversal electric field of the first higher order mode, having an electric wall as symmetry condition.

### 2.2 Coplanar Waveguide

As already explained in Section 2 on page 1, both outer ground layers are for technical reasons on the same potential. Therefore, this waveguide is a twofold bounded transmission line and thus, there is no cut-off frequency of the fundamental mode. Figure 3 shows the transversal electric field pattern of the only quasi-TEM mode on this structure.

![Coplanar waveguide](image2)

**Fig. 3**: Transversal electric field pattern of the fundamental wave in the cross section of a coplanar waveguide.

### 3 Characteristic Parameters of Planar Waveguides

The local distribution of the electromagnetic fields in planar waveguides is more or less frequency dependent. In principle, the field energy is stored more and more in the dielectric layer with increasing frequency. Due to this, the **propagation constant** $\gamma$ as well as the **characteristic impedance** $Z_0$ change with frequency. The dependence of the propagation constant or the **phase velocity** is called **dispersion**, whereas this expression has other meanings, too. Generally, the characteristic impedance is of interest of only propagating modes. In the case
of weak losses, the imaginary part may be neglected, therefore \( Z_0 \) is approximately assumed as real. In contrast, \( \gamma \) is purely imaginary only for propagating modes on lossless transmission lines. It is real for evanescent modes, and in the case of losses complex for all modes.

For both circuit synthesis and analysis, knowledge of the mentioned characteristic parameters is vital. They have to be measured or calculated in the first place. These results are either stored in **look-up tables** or—more frequently—a curve fitting procedure leads to an equation for \( \varepsilon_{\text{eff}} \) as function of width and thickness of the conducting layers, substrate permittivity, height and so on.

An exact analysis needs the solution of the complete set of Maxwell’s equations. The numerical effort in computing only straight lines is quite high even by applying idealized assumptions concerning ohmic and dielectric losses, surface roughness etc. On the other hand, there is a number of simple models which leads in many cases to results which are precise enough for practical problems.

### 3.1 Propagation Constant \( \gamma \)

**Propagation constant** \( \gamma \), wave number \( k_z \) and the so-called **effective dielectric constant** \( \varepsilon_{\text{eff}} \) are different values for the same characterization of a wave. The effective dielectric constant is a more fictive parameter; a wave traveling on a planar line having the substrate permittivity \( \varepsilon_s \) has the same propagation properties as another wave traveling in a TEM waveguide loaded with the permittivity \( \varepsilon_{\text{eff}} \).

The propagation of an electromagnetic wave may be expressed by the following time and space dependency (in propagating direction \( z \)):

\[
\vec{E}, \vec{H} \propto e^{j\omega t - \gamma z}.
\]  

(1)

Attenuation and phase constants are the real and imaginary part of the propagation constant:

\[
\gamma = \alpha + j\beta,
\]

(2)

the imaginary part corresponds to the phase characteristics, the real part to the attenuation of the propagating mode. These losses may be ohmic losses of the dielectric layer or of the conducting strip, but radiation is also possible.

The wave number \( k_z \) is:

\[
k_z = -j\gamma = \beta - j\alpha,
\]

(3)

and the effective dielectric constant \( \varepsilon_{\text{eff}} \):

\[
\varepsilon_{\text{eff}} = \left( \frac{k_z}{k_0} \right)^2
\]

(4)

with the wave number of the free space:

\[
k_0 = \omega \sqrt{\mu_0 \varepsilon_0}.
\]

(5)
The descriptive meaning of $\varepsilon_{\text{eff}}$ shown above is however valid for lossless media and by considering of the fundamental mode, only. In this case $\varepsilon_{\text{eff}}$ has to be real. Assuming this pre-requisite and the well-known relation:

$$ c = f \lambda, $$

inserted into (4) leads to the interrelation:

$$ \varepsilon_{\text{eff}} = \left( \frac{\lambda_0}{\lambda} \right)^2, $$

with the free space wavelength $\lambda_0$.

$\sqrt{\varepsilon_{\text{eff}}}$ is an extent for the changing of the wavelength with respect to the value of the free space. Formally, $\varepsilon_{\text{eff}}$ may be complex, e.g. in the case of lossy media.

The propagation constant of the fundamental mode on a lossless TEM transmission line (e.g. the coaxial line) is:

$$ \gamma = j \omega \sqrt{\mu \varepsilon}, $$

$$ \mu = \mu_0 \mu_r, $$

$$ \varepsilon = \varepsilon_0 \varepsilon_r. $$

$\mu_r$ and $\varepsilon_r$ are the material constants of the layer in between the metals.

The application of the conformal mapping by H.A. Wheeler was one of the first approaches concerning the calculation of the propagation constant of the microstrip line. With the aid of this mathematical method static, cylindrical waveguides may be analyzed (derivatives of the electromagnetic field components in $z$ direction have to be zero!).

One of the already mentioned methods is very good suited for on line CAD programs: in the first step, the characteristic parameters of the most important planar transmission lines were determined very precisely using full-wave numerical methods based on Maxwell’s equations. In the second step, the dependencies of these characteristic parameters on the physical dimensions, material parameters and frequencies were fitted by analytical functions. These functions may be evaluated in a computer very fast.

Figure 4(a) shows a typical dispersion curve of a microstrip line (characteristic impedance 50$\Omega$) on a substrate having the dielectric constant $\varepsilon_r=11$.

Starting at the static value, the dispersion curve increases monotonically up to infinitely high frequencies, but values are always in the limited range:

$$ \frac{\varepsilon_r + 1}{2} < \varepsilon_{\text{eff}} < \varepsilon_r. $$

The coplanar waveguide shows a minor dispersion characteristic. In the same figure, curve (b) is shown for such a transmission line, built on the same substrate material and having the same characteristic impedance. The value of the static effective permittivity at very low frequencies is about $\frac{\varepsilon_r + 1}{2}$ (this is valid for metallization thickness $t=0$ only, the values are due to this thickness smaller), since half of the field is in the infinite air region, the other half is in the infinite dielectric layer.
3 Characteristic Parameters of Planar Waveguides

Fig. 4: Typical dispersion curve of a $50\,\Omega$ microstrip line (a) and a $50\,\Omega$ coplanar waveguide (b) on a substrate having the permittivity $\varepsilon_r=11$ and metallization thickness $t=17\,\mu m$, MSL: center conductor width 0.42 mm, substrate height $h=0.5\,mm$, CPW: center conductor width 0.2 mm, gap width 0.1 mm.

3.2 Characteristic Impedance

The characteristic impedance is an important parameter of all RF waveguides and may be determined very easily in the case of TEM transmission lines. For the lossless coaxial transmission line we get:

$$Z_0 = \sqrt{\frac{L'}{C'}} = \frac{1}{2\pi} \ln \left( \frac{r_2}{r_1} \right) \sqrt{\frac{\mu}{\varepsilon}},$$

where $r_2$ and $r_1$ are the outer and inner radius of the coaxial lines, resp. A TEM transmission line means that the fundamental mode is a TEM wave. All higher order modes start propagating at higher frequencies and have at most five electromagnetic components, one longitudinal component (in direction of propagation) is always zero. The specification of a characteristic impedance of higher order modes is possible in a more formal manner, but without technical concern.

There are no analytical relations of the characteristic impedances in non TEM transmission lines. But there is the quite reasonable method by using integral values characterizing the mode in a whole as it is well-known in the network theory. Voltage, current, and power may be derived from Maxwell’s equations in integral form. To this end, we get the impedance:

$$Z_0 = \frac{U}{I} = \frac{U^2}{P} = \frac{P}{I^2},$$  \hspace{1cm} (8)

with the voltage: $U = \int \vec{E} d\vec{s}$, \hspace{1cm} (9)

the current: $I = \int \vec{H} d\vec{s}$, \hspace{1cm} (10)
and the power traveling in direction of propagation ($z$-component of the Poynting vector):

$$P = \iint_{(A)} (\vec{E} \times \vec{H}^*) \cdot dA.$$  \hspace{1cm} (11)

All relations of (8) are not equally well suited for all planar transmission lines. Equ. (9) and (10), e.g. depend on the surface or line integrals. In practice the determination of the characteristic impedance on a microstrip line is more reasonable from current and power whereas on a slotline the impedance should be calculated from voltage in the slot and power. The current is the closed-loop integral on the conducting strip line, the power is the surface integral over the infinite cross section of the transmission line, the voltage is the line integral across the shortest distance between the two conducting layers. The characteristic impedance of the microstrip line is determined by the current and the power, that of the coplanar waveguide by the voltage and the power. It may be summarized that this definition is very useful to characterize the transition from the transmission line to therein inserted lumped elements.

## 4 Planar Components

Generally, it is not possible to derive the electric behavior of a whole circuit by analyzing individual parts of it. Due to additional stray fields at step in width of the center conductor (as an discontinuity example), the overall transfer function may differ totally as predicted by the transmission line theory for the connection of the individual circuit parts. Only solving Maxwell’s equations simultaneously by using numerical methods leads to precise results. The drawback of these analysis methods is the large (or huge) amount of computation time, unbearable by applying this procedure to interactive CAD programs. This either leads to the computation of complex elements in advance or to deriving models which need much less computation time as the so-called full-wave analysis using Maxwell’s equations as described before.

A well-suited model for computing a lot of microstrip line circuits is the so-called Magnetic-wall model. The microstrip line with the quasi-TEM mode can be considered as a kind of a TEM mode carrying rectangular waveguide bordered by electric walls at the horizontal boundaries and by magnetic walls at the vertical boundaries, refer Fig. 5 and Fig. 2(a). The field pattern in Fig 2(a) has to be changed only slightly to fit into the waveguide in Fig. 5: the stray field is considered by a bit greater width of the waveguide ($w_{\text{eff}} \gtrsim w$) and by a homogenous filling of the waveguide with a relative effective dielectric constant with a value between the values of air and the substrate material of the microstrip line ($1 < \varepsilon_{\text{eff}} < \varepsilon_r$). The two vertical magnetic walls are almost in parallel to the original electric field and do not disturb the field pattern, and the two horizontal electric walls stem from the ground and the microstrip. The word waveguide suggest a cut-off frequency but the new waveguide has two conductors as the original microstrip line.

Using this model, it is much easier to find solutions for the transmission line parameters than by using the full-wave methods described before. Beside the lateral stray field there are additional distortions of the electromagnetic field at discontinuities, i.e. abrupt changing of the cross section. The influence of this change of the straight line has to be calculated.
using numerical methods and may be considered in the model as additional changes of the
effective parameters. In the case the circuit contains bifurcations of the strip, steps in width
or similar discontinuities, the model results in a simple waveguide circuit which may be solved
numerically in reasonable computation time.

In the following section, three basic components in microstrip line technique will be presented
in a more precise manner: the step in width (impedance step), the open-ended transmission
line and a microstrip resonator.

### 4.1 Step in Width on a Microstrip Line

Figure 6(a) shows a step in width of the metal strip of a microstrip line. The symmetric
step is characterized by the substrate parameters \((\varepsilon_r, h)\), the geometrical dimensions of both
strips (widths \(w_1, w_2\), metallization thickness \(t\)), and the distances between the step and the
reference planes \(l_1\) and \(l_2\).

At low frequencies, this structure may be considered as the connection of two transmission
lines having different characteristic impedances. The reflection coefficient at the transition
with respect to port 1 is (well-known from the transmission-line theory):

\[ r_1 = \frac{Z_2 - Z_1}{Z_2 + Z_1}. \]

With increasing frequency the stray fields store more and more energy and the step in width
can be described by an equivalent circuit consisting of an additional series inductance due to
the stringing of the current and a parallel capacitance due to the stray fields at the additional
metal edge at the interface, shown in Fig. 6(c). By applying the Magnetic-wall model, the
structure becomes a waveguide circuit having magnetic side walls and is characterized by:

\[ w_{\text{eff},1} > w_1, \quad w_{\text{eff},2} > w_2, \quad h, \]
\[ l_{\text{eff},1} < l_1, \quad l_{\text{eff},2} > l_2, \quad 1 < \varepsilon_{\text{eff},1,2} < \varepsilon_r, \]

and of course, it is:

\[ l_1 + l_2 = l_{\text{eff},1} + l_{\text{eff},2}, \]

because the reference planes should not be moved.
With increasing frequency the scattering parameters become more and more inaccurate due to the model. The reasons are the higher order modes on the microstrip line which have already close-by but still under their cut-off frequency—an observable influence on the transmission behavior of the fundamental mode. Due to the Magnetic-wall model, these modes may not be considered, because a higher order mode has another field distribution as the fundamental one which would lead to other effective physical parameters.

Examining multi-layer structures or gaps in the strip, the model fails again; precise numerical methods requiring appropriate computing time have to be used to get correct results, either for each individual circuit or by creating other models, e.g. models consisting of only lumped elements.

### 4.2 Open-Ended Transmission Line

In Fig. 7, an open-ended microstrip line with its cross section and electric field pattern is shown, which has to be characterized. The stray field at the end of the transmission line may
be considered either by an additional capacitance as shown in Fig. 7(b) or the field outside of the transmission line may be replaced by an additional line length $\Delta l$ of the same transmission line and an ideal magnetic wall (reflection coefficient $r=1$) in the end as shown in Fig. 7(c). The equivalent circuits “open end with a short transmission line” and “open end with a capacitance” behaves the same as a short view into the Smith chart shows.

Fig. 7: Consideration of the stray field of an open-ended transmission line.

At high frequencies the open-ended transmission line acts as an antenna and radiates a portion of the incident RF power. Therefore, the reflection coefficient is smaller than one.

4.3 Planar Transmission Line Resonator

Fig. 8: Top view and cross section of a resonator in microstrip line technique.

The determination of the effective dielectric constant based on measurements will be done in this lab by measuring of the resonance frequencies of resonators. The resonator consists of a short transmission line coupled by two gaps to the feeding transmission lines, according to Fig. 8. The gap between the feeding transmission lines and the resonator may be considered as an inverter. The whole structure is a bandpass filter having a very high quality factor. Here the external quality factor is supposed, i.e. the quality factor with the connections (by gaps) to the transmission lines. The weaker the coupling the higher the external quality factor. The first resonance of this structure occurs at that frequency, where the field pattern of half a
wavelength on this transmission line “fits” into the physical structure. Due to the stray field at both sides of the resonator (coupling gaps), the physical length of the transmission line is larger than half a wavelength:

\[ l = \frac{\lambda}{2} - 2\Delta l. \]

In the case two resonators were measured, their lengths differ approximately by \( \frac{\lambda}{2} \) at the lowest resonance frequency; it follows:

\[ l_1 = \frac{\lambda_1}{2} - 2\Delta l, \]
\[ l_2 = \lambda_2 - 2\Delta l. \]  (12)

The influence of the two open ends, i.e. the virtual enlargement of the transmission line \( \Delta l \) is in both cases almost identical, since the end effect has only a weak frequency dependence and may be canceled by subtracting the two equations. From (6) and (7), the effective permittivity becomes:

\[ \varepsilon_{\text{eff}} = \left( \frac{c_0}{l_1 - l_2} \left( \frac{1}{2f_1} - \frac{1}{f_2} \right) \right)^2, \]  (13)

if the resonance frequencies \( f_1 \) and \( f_2 \) are close enough to neglect the dispersion between these two frequency points.

5 Planar Circuits

In this section, a hybrid ring coupler (also known as quadrature coupler) distributing the incident power to two output ports should be characterized as an example of a planar circuit. Figure 9 shows the top view of such a planar hybrid ring coupler with the power distribution \( P \), the electric lengths, and the characteristic impedances of the connecting and feeding lines.

![Fig. 9: Schematic plot of a hybrid ring coupler.](image_url)
The characteristic impedances and the electric lengths were chosen to ensure the proper behavior of the circuit. Using the transmission-line theory, the mode of operation may be explained in the following manner:

Assuming port 4 is short-circuited, transformed over $\frac{\lambda}{4}$ long lines results in open ends at ports 1 and 3. Thus port 3 is loaded by $Z_0$, only, transformation of $Z_0$ over a line having the characteristic impedance $Z_0$ results in $Z_0$, port 2 therefore is loaded with $Z_0$. Transformation of this impedance over a $\frac{\lambda}{4}$ long line (characteristic impedance $Z_0/\sqrt{2}$) between port 2 and 1 results in the input impedance $Z_0$ of port 1 (matched condition).

Figure 10 shows the equivalent circuit of this operation state.

![Equivalent circuit of the hybrid ring coupler in the case port 4 is short circuited.](image)

Fig. 10: Equivalent circuit of the hybrid ring coupler in the case port 4 is short circuited.

If port 1 is fed, the voltages at ports 2 and 3 as function of the input voltages are:

\[
U_2 = -j\frac{U_1}{\sqrt{2}},
\]

\[
U_3 = -\frac{U_1}{\sqrt{2}},
\]

and the short-circuit current (not sketched) effected by both signal paths 1-4 and 1-2-3-4 at port 4:

\[
I_{14} = -j\frac{U_1}{Z_0},
\]

\[
I_{34} = j\frac{U_1}{Z_0}.
\]

Equations (14) to (17) signify that

1) half of the incident power leaves the four port at ports 2 and 3, ($P \approx U^2$),
2) the values at these ports are shifted by $90^\circ$ and $180^\circ$ with respect to port 1,
3) port 4 is decoupled.

In a whole the scattering matrix looks like that:

\[
S = \frac{-1}{\sqrt{2}} \begin{pmatrix}
0 & j & 1 & 0 \\
 j & 0 & 0 & 1 \\
1 & 0 & 0 & j \\
0 & 1 & j & 0
\end{pmatrix} .
\]
6 Measuring Set-up

Figure 11: Measuring set-up.

Figure 11 shows the measurement set-up. A sweep generator for the frequency range from 100 kHz up to 6 GHz is used as RF source. The output power of the generator may be adjusted.

The directional coupler in the measurement set-up is used to redirect the power reflected by the DUT to the diodes. The RF signal travels to the mounting in which the planar circuits is fixed. The detector diodes rectify the RF signal, the output will be amplified in a logarithmic amplifier.

An introduction to the scattering parameter measurements is given in the script of lab Scalar scattering parameter measurements (coaxial).
7 Questions and Problems on the Lab

Problem 1: Is the characteristic impedance of the quasi-TEM mode on a microstrip line frequency-dependent, in the case $\varepsilon_r \neq f(f\text{requency})$ is valid? Give reason for your answer.

Problem 2: What is the reason for the insertion of air bridges in coplanar waveguides ensuring the same potential of both ground layers?

Problem 3: Make a sketch of all electromagnetic field component of the fundamental mode in the Magnetic-wall model.

Problem 4: Using (13) on page 11 the effective permittivity of a transmission line can be determined using the first resonance on the resonator, i.e., there is a $\lambda/2$-resonance at $f_1 \approx f_2$.

Deduce (13) for multiples of this frequencies.

Problem 5: How many fundamental modes without cut-off frequency exist on the four different planar waveguides shown in Fig. 1? Make sketches of the tangential electric fields of all these modes.

Problem 6: Deduce (14) to (17) on page 12.

Problem 7: Which qualitative influence has the length of the coupling gap $s$ between the line resonator and the feeding microstrip line in Fig. 8 on page 10 on the resonance frequency and the quality factor of the resonator?

Problem 8: What is the the meaning of mode dispersion, waveguide dispersion, material dispersion, and chromatic dispersion?
8 Measuring Tasks

Set-up of the RF signal source generator *Agilent N5181A*:
1) Connect the SMA plug of the data logger with the input 'TRIG IN' of the source.
2) Adjust the following parameters of the source:
   - SWEEP - Sweep - Freq On
   - SWEEP - Configure Step Sweep
     - Freq Start: 100kHz
     - Freq Stop: 6GHz
     - # Points: 1000
     - More - Step Dwell: 100us (Minimum)
   - SWEEP - More - Sweep Trigger - Ext - Trigger In Polarity → Pos
   - SWEEP - More - Point Trigger - Ext - Trigger In Polarity → Pos

Set-up the power:
- AMPTD - according to the task
- RF On

3) Hints:
   - If during the measurements the 'Current Point' is not the first point (1/1000), perform:
     - SWEEP - Single Sweep - Sweep Repeat Cont

Usage of the software used for the calibration and measurement data transfer and display
1) Log in a user 'praktikum', your tutor knows the password.
2) Change into the directory for all data. Examples:
   - cd 1_Praktikum/ehf_V04_Skalare-Spara/
   - cd 1_Praktikum/ehf_V05_Planare_Schaltungen/
3) Start the graphical user interface (GUI)
   DataloggerFrontend
4) Press the following button if you want to change the virtual sheet:
   New Sheet
5) For the storage of a new calibration curve:
   Calibration
6) For the storage of a new measurement curve (differs from the calibration curve only in color and line type):
   Measurement
7) Display of all curves and printing using ’gv’:
   Plot
8) Please label the curves immediately after printing.

Hints:
- All files (data files, postscript files and so on) are stored in the same directory where
  the program was invoked from. These files have unique numbers referring to the virtual
  sheets, and it is always possible to rename or to delete some of them.
- If the GUI was terminated and started again the file number of the following measured
  data are increased automatically.

Important notes:
Almost all components of the measurement set-up are accomplished by SMA connecting tech-
ique. In connecting the components or transmission lines as well as in detaching, strictly,
only the compression nut have to be turned. The first fixing of a microstrip line or coplanar
line has to be done on the supervision of the tutor, only. Fix the connecting coaxial lines by
using a torque wrench.

The substrate material is Rogers RT/Duroid 6010 with thickness 0.635 mm and 1.27 mm.
Disregarding these rules leads inevitably to a fast abrasion or damaging of the sensitive com-
ponents of the measurement set-up and DUTs.

The following measurements will be performed in the frequency range 100 kHz–6 GHz and
nearly at maximum available RF power (turn the internal power potentiometer at position
5 dBm, the switchable attenuator at position 0 dB).

Task 1: Determination of the dispersion

The dispersion of a microstrip line and a coplanar waveguide (frequency dependence of
the effective dielectric constant) should be determined by using the resonance method
(transmission measurements, refer section 4.3 on page 10).

Practical performance:
Take the calibration curves (transmission calibration) but you have to distinguish which
generator you are using:

**Agilent N5181A** (bis 6 GHz):
from 5 dBm ≅ 0 dB in 5 dB steps until −35 dBm ≅ −40 dB.

**Agilent N5183A** (bis 32 GHz):
from 20 dBm ≅ 0 dB in 5 dB steps until −20 dBm ≅ −40 dB.

Make use of a 20 dB attenuator at the output port of the generator to avoid the destruc-
tion of the detector diode.

Actually, this calibration is not necessary but you get information about the power levels.
Determine the resonance frequencies of two different resonance structures per transmis-
sion line type (microstrip line and coplanar waveguide). Make use of different sheets for
the two transmission line types. Print each diagram, label the calibration curves and
read the resonance frequencies. Determine the values of the effective permittivity from
those resonance frequencies at four frequency points.

The physical lengths of the microstrip resonators are 37 mm and 76 mm, those of the coplanar waveguides are 52 mm and 110 mm.

**Task 2: Determination of characteristic impedances**

Determine the characteristic impedances of three low-resistance transmission lines and discuss the possibly occurring problems. On the one side with the low-resistance transmission lines, a $50 \, \Omega$ transmission line is attached abruptly, on the other side, there is an inhomogeneous (in case of at least the length of the wavelength a so-called *tapered*) transmission line. The taper is a gradual transition between two geometries in such a way, that no disturbing reflection on the interface coaxial measuring system – microstrip line may occur.

Practical performance:
Perform a reflection calibration, the DUT is a short circuit. Take the calibration curves for $5 \, \text{dBm}$ to $-20 \, \text{dBm}$ in $5 \, \text{dB}$ steps. Afterwards, take the measurements of the three impedance steps and determine from these results in each case the impedances of the wider transmission lines. The not connected port has to be matched using a $50 \, \Omega$ load.

**Task 3: Measurement of a hybrid ring coupler**

Determine the amplitudes of all scattering coefficients which would be different also in case of an ideal coupler. Ensure matching of all ports which are not connected to the measurement system. Determine the operating frequency of the coupler.

What kind of losses are possible?

Practical performance:
Perform a transmission calibration at $0 \, \text{dB}$ up to $-30 \, \text{dB}$ in $3 \, \text{dB}$ steps. Draw all of the three transmission parameters in the same diagram.

After this perform a reflection calibration at $0 \, \text{dB}$ up to $-35 \, \text{dB}$ in $5 \, \text{dB}$ steps and measure one reflection coefficient.

For a better evaluation of the losses, measure again the transmission coefficients of the coupled ports by using a transmission calibration at $5 \, \text{dB}$ in $1 \, \text{dB}$ steps.