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Earthing in HF and Microwave Circuits ... A Case for Puff

In any circuit with active components, we are confronted with the problem that certain lines or parts of circuits do not belong to the purely AC voltage circuit. Using appropriately adjusted earthing points, we can ensure that these components fulfil only DC tasks. And these earthing points are the subject of this article. In practise, we see every possible structure on printed circuit boards, and ask ourselves what advantages this structure is intended to convey. Which is precisely where this article is intended to be of use.

1.

Definition of Term Correct Earthing

We should first agree on the definition below for the term correct earthing.

A circuit point or a connection line can be said to be correctly earthed if, within the frequency range under consideration, the impedance (Z) between this point and earth is lower than 5 Ohms.

Let us now put the miscellaneous standard measures under the microscope, taking this point of view, originating from developers practise.

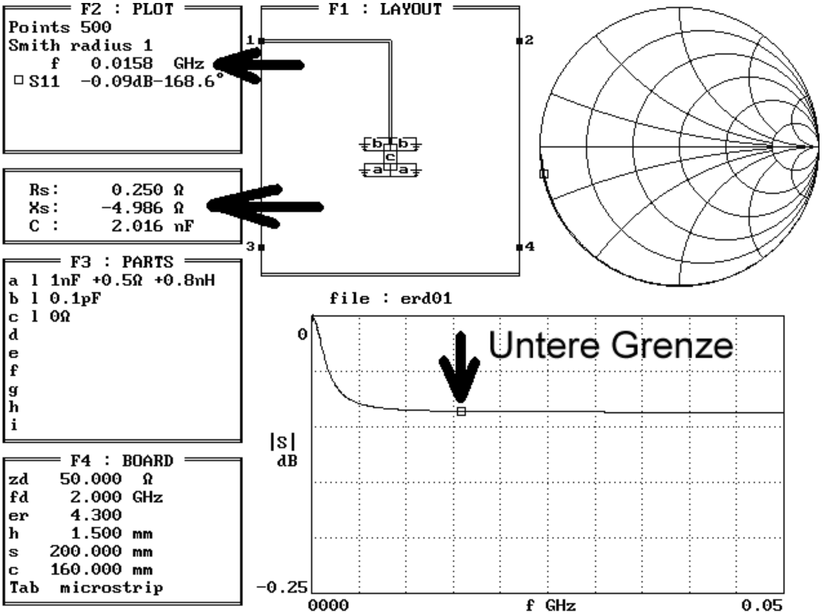
2.

The Earthing Capacitor

This measure is certainly the oldest in electronics. In theory this method is first class, since it is indeed well-known that the AC resistance of a capacitor decreases as the frequency increases, and thus the connection to earth can only become even better.

2.1. Aluminium electrolytic capacitors

In practice, though, things look rather different. At very low frequencies, polarised aluminium electrolytic capacitors are used, but these display a series resistance which rises with the frequency and a reactive component of the impedance which also rises, due to inherent inductance. Moreover, it is not officially mentioned that the quality of electrolytic capacitors has deteriorated. The mechanical dimensions have been continuously reduced, but any radio or television technician will be happy to confirm that, for example, modern electrolytic



```
F2 : PLOT
Points 500
Smith radius 1
    f   0.0158  GHz
□S11  -0.09dB-168.6°

  Rs:      0.250 Ω
  Xs:     -4.986 Ω
  C :      2.016 nF

F3 : PARTS
a l 1nF +0.5Ω +0.8nH
b l 0.1pF
c l 0Ω
d
e
f
g
h
i

F4 : BOARD
zd    50.000  Ω
fd     2.000 GHz
er     4.300
h      1.500 mm
s    200.000 mm
c    160.000 mm
Tab  microstrip
```

Fig. 1: Determination of lower limit for use of 2 parallel-wired 1-nF SMD capacitors as earthing (Untere Grenze = Lower limit)

capacitors are the cause of the problem in 50% of repairs of TV sets. They no longer break down, but usually slowly dry out, because the seals become porous. The unfortunate result of this is that, in addition to the capacity drop, which usually would not be so awful, there is very rapidly a sharp rise in the series resistance and thats bad!

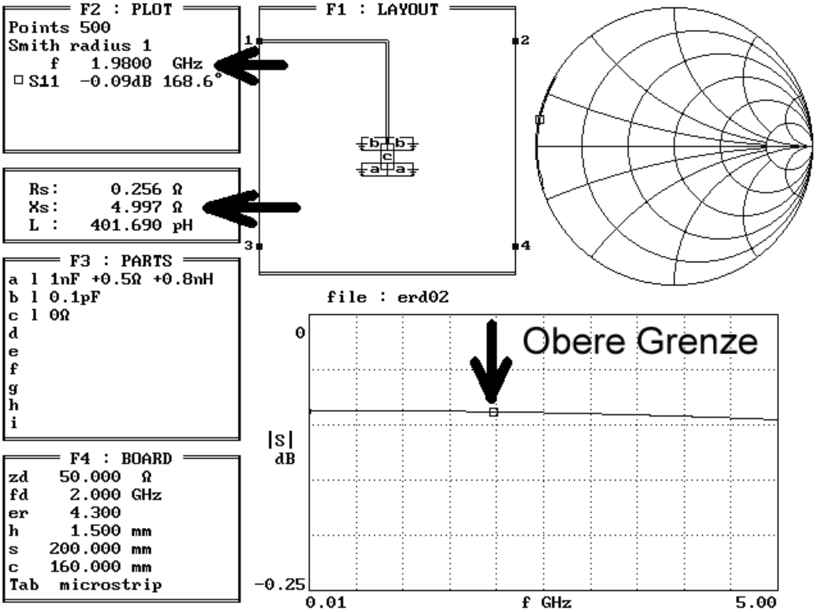
2.2. Tantalum electrolytic capacitors

Tantalum electrolytic capacitors behave considerably better here. Sintered from powder, they basically represent semiconductors (i.e. diodes) which are operated in the filter attenuation band. Their advantages lie in their small dimensions, their low series resistance (approximately 1-2 Ohms) and their low inherent inductance of a few nH. This gives excellent coupling capacitors and / or wide-band earthings. In particular, in parallel circuits with several capacitors, they cover a range from low frequency to far above 100 MHz.

Let us not conceal their negative characteristics. Like any semi-conductor, they react extremely sensitively to overvoltage and / or pole reversal. In the most favourable case, disruptive short-circuits then occur. In the most unfavourable case, however, there can be scarcely predictable reductions, great or small, in the insulation resistance, which are to some extent dependent on temperature.

One factor here is often misunderstood. It will certainly say somewhere in the data sheet that, in the interests of service life, standard tantalum electrolytic capacitors should not be subjected to a switching current pulse exceeding approximately 0.3 A.

The reason for this is very simple. If this value is exceeded, it can quickly be-



```
F2 : PLOT
Points 500
Smith radius 1
    f    1.9800   GHz
□ S11   -0.09dB 168.6°

  Rs:      0.256 Ω
  Xs:      4.997 Ω
  L :    401.690 pH

F3 : PARTS
a  1  1nF +0.5Ω +0.8nH
b  1  0.1pF
c  1  0Ω
d
e
f
g
h
i

F4 : BOARD
zd    50.000  Ω
fd     2.000 GHz
er     4.300
h      1.500 mm
s    200.000 mm
c    160.000 mm
Tab   microstrip

file : erd02
```

Fig. 2: Upper limit for earthing from two parallel-wired 1-nF SMD capacitors (Obere Grenze = Upper limit)

come so hot at individual points within the sintered material (in which, indeed small grains are in contact with each other) that a melt-on occurs, leading to a short circuit.

It makes no difference whether the current limitation required is brought about through the power supply itself or using a pre-resistance.

2.3. Ceramic capacitors

From 100 MHz upwards, the only capacitors still used are essentially ceramic models. Their quality is adequate for all requirements up to 2 GHz, even with the cheapest standard formats.

However, one should, as far as possible, attempt to use only SMD solutions, for only their low inherent inductance (< 1 nH) eliminates the tiresome inherent resonances of the wired-up copies. Here it is particularly advantageously to have parallel circuits of at least two capacitors, which means that the Ohmic resistance and the inherent inductance are smoothly halved.

We can use Puff to show how such a layout behaves at various frequencies. To do this, four components are parallel-wired in a simulation:

a) Two SMD capacitors, each with 1 nF + 0.5 Ohms series resistance + 0.8 nH inherent inductance, and

b) Two capacitors, each with 0.1 pF, each simulating the size 0805 SMD pad.

It is recommended that analyses should be distributed into an upper and a lower range for such very broad-band layouts

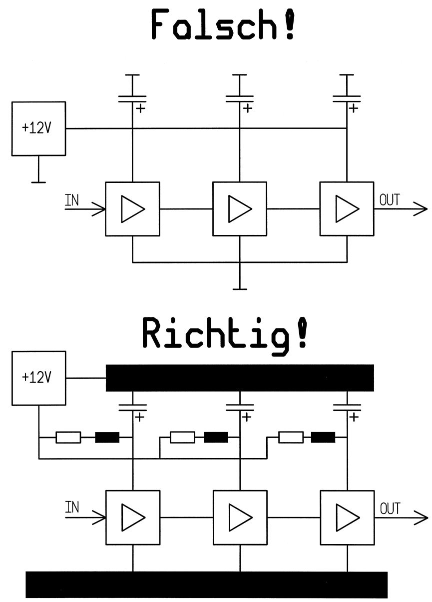


Fig. 3: Wrong and right layout for a multi-stage broad-band amplifier (Falsch = Wrong, Richtig = Right)

and for the Puff representation, restricted to 500 dots. Fig. 1 therefore shows the behaviour from 0 to 50 MHz, in the form of the reflection factor S11, for this parallel circuit, which will supply the values for the impedance curve of the layout.

For this purpose, use Page up or Page Down to go to a specific frequency, place the cursor in field F2 on S 11 and key in the equals sign, =.

The active and reactive components of the input resistance immediately appear in the dialogue window, and the associated Substitute dummy component is displayed as an inductance or capacity value. Now we look for the frequency at which the capacitive reactive component of this layout undershoots 5 Ohms. In Fig. 1, this happens at 15.8 MHz.

In a second pass, we take the range from 10 MHz to 5 GHz amd there we find, at precisely 2 GHz, 5 Ohms for the (now inductive) reactive component (Fig. 2). It specifies the upper limit of the usable frequency band.

Anyone who wants to know why the actual value zero was not taken as the lower frequency limit for Puff can repeat the simulation using this value. The blemish arising in the representation in the Smith chart, due to the fact that the step widths at very low frequencies are now far too great, is avoided using the setting proposed.

The parallel-wired layout of ceramic capacitors is thus effective and correspondingly popular. To set the bottom frequency limit still lower, we simply replace the ceramic-SMD capacitors with a parallel circuit of several tantalum electrolytic capacitors, in which the loss resistances and inherent inductances are approximately twice or three times as great.

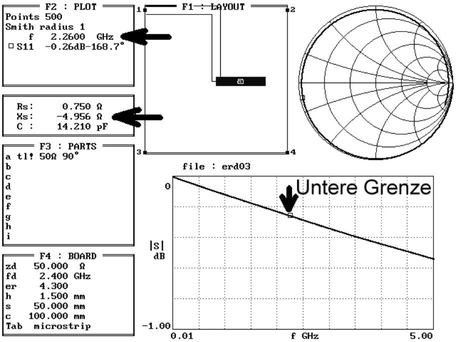
2.4. Notes on sources of errors

Unfortunately, we can find some circuit technology earthing errors in power supply systems not only in DIY projects, but even in industrially produced circuits.

Fig. 3 shows a wrong and a right format for a multi-stage broad-band amplifier as an example:

a) There are no protective resistors for the tantalum electrolytic capacitors.

b) The power supply lines of the individual stages must have star connections to the power supply and must be de-coupled from one another. With the help of a choke coil and / or an Ohmic resistor, we can thus then create an LC or RC low pass in each feeder with the tantalum electrolytic capacitor. Only in

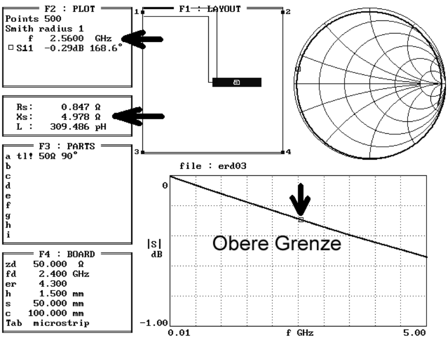


```
F2 : PLOT
Points 500
Smith radius 1
    f   2.2600  GHz
□ S11  -0.26dB-168.7°

  Rs:      0.750 Ω
  Xs:     -4.956 Ω
  C :     14.210 pF

F3 : PARTS
a tl! 50Ω 90°
b
c
d
e
f
g
h
i

F4 : BOARD
zd   50.000  Ω
fd    2.400 GHz
er    4.300
h     1.500 mm
s    50.000 mm
c   100.000 mm
Tab  microstrip
```



```
F2 : PLOT
Points 500
Smith radius 1
    f   2.5600  GHz
□ S11  -0.29dB 168.6°

  Rs:      0.847 Ω
  Xs:      4.978 Ω
  L :    309.486 pH

F3 : PARTS
a tl! 50Ω 90°
b
c
d
e
f
g
h
i

F4 : BOARD
zd   50.000  Ω
fd    2.400 GHz
er    4.300
h     1.500 mm
s    50.000 mm
c   100.000 mm
Tab  microstrip
```

Fig. 4: Determination of lower limiting frequency using a 50-Ohm microstrip line on no-load as earthing (Untere Grenze = Lower limit)

Fig. 5: Upper limit using 50-Ohm microstrip line on no-load as earthing (Obere Grenze = Upper limit)

this way can we reliably prevent part of the high level of the last stage from going back to the input section and causing the layout to oscillate.

The HF choke used, however, must not be of too high a quality (to avoid resonance effects). Here it is thoroughly normal to wire Ohmic resistors into the circuit. This not only attenuates but improves the low-pass effect, especially at lower frequencies, where the inductive resistance is known to be still low. This simultaneously provides the protective resistors for the tantalum electrolytic capacitors.

c) The points connected directly to earth (= without capacitor) within an amplifier stage must not just be looped on to the next stage, or an unnecessary oscillation risk will be created.

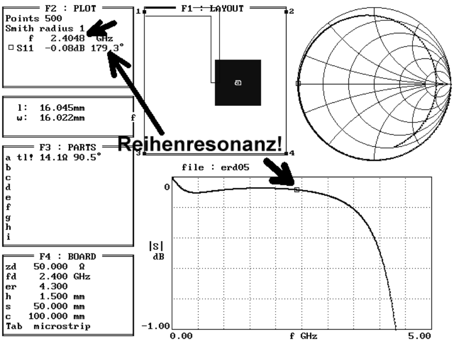
A version mounted on printed circuit boards in accordance with the laws of microwave engineering is ideal, and almost oscillation-proof, even at low frequencies. Coated on both sides, this version thus has an underside which is an integrated earthing and earth level. In the illustration it is shown as a blocked-in rectangle. The direct earthings are brought about through suitable through-platings, which can be implemented in the form of full tubular rivets (silvered, diameter 0.8 mm.).

The screening action of such choke / capacitor layouts at various frequencies can also be determined by means of Puff. But you can find out from [1] how expensive this can be, for example, for a low-noise oscillator.

3.

Earthing Through Microstrop Lines

At higher frequencies, if discrete components slowly fail, we resort to sections of line working at no-load as earth connections. It is well known that such a line represents a short circuit at l = /4.



```
F2 : PLOT
Points 500
Smith radius 1
    f   2.4048  GHz
□ S11  -0.08dB 179.3°

  l:  16.045mm
  w:  16.022mm

F3 : PARTS
a tl! 14.1Ω 90.5°
b
c
d
e
f
g
h
i

F4 : BOARD
zd   50.000  Ω
fd    2.400 GHz
er    4.300
h     1.500 mm
s    50.000 mm
c   100.000 mm
Tab  microstrip
```

Fig. 6: Using a square pad with an edge length l = λ/4 as earthing (Reihenresonanz = Series resonance)

Thus, for example, we can take such a lossy λ/4 line with 50 Ohms for 2.4 GHz and simulate its behaviour for a printed circuit board made from epoxy material FR4, using the data:

Thickness 1.5 mm, r = 4.3 and

Loss factor lt = 0.02

We are less interested in behaviour under resonance here, for it is well known that only the small loss resistance remains behind there (approximately 0.8 Ohms). Use Page Up or Page Down to move the cursor until keying in the equals sign at S11 gives you an inductive or a capacitive reactive component of 5 Ohms (Figs. 4 and 5). Frequencies associated with this are 2,260 MHz and 2,560 MHz earthing would thus be usable only in a narrow range 300 MHz wide. This method is therefore also deliberately used only for narrow-band applications or oscillators.

Things work considerably better, with a broader band, if the section of line working at no-load is altered to such an extent that the length and width are the same and form a square. If we also make the area of this square sufficiently big, the line, because of the low impedance level, acts like a big capacitor with a correspondingly low reactive impedance, even at really low frequencies (i.e. at l < λ/4).

If the frequency is then increased, the λ/4 resonance follows, with the short circuit at the input. Not until this resonance is exceeded does the reactive component increase again, due to the inductive behaviour, but with a low-Ohm line the inductance is indeed also very low. This leads us to expect a decidedly wide usable frequency range, in which the impedance is sufficiently low.

This can be tested for 2,400 MHz using Puff. With the same printed circuit board data as before, the length and width at this frequency are to be given values which correspond to a quarter of the wavelengths. These include lengths and widths of approximately 16 mm. and an impedance level of 14.1 Ohms (Fig. 6).

If we now simulate the behaviour of this line and look again for the points with a reactive component of 5 Ohms, we find the frequency values 2,867 MHz and 1,973 MHz are suitable. This layout would also fulfil the requirements laid down in a range of approximately 900 MHz. For a direct comparison, the impedance curves (Z) = f(frequency) for the two sections of line have been plotted in a joint diagram (Fig. 7).

The advantage of the square section of line in broad-band applications is once again easy to see here. Unfortunately, it is very cumbersome for practical application at low frequencies.

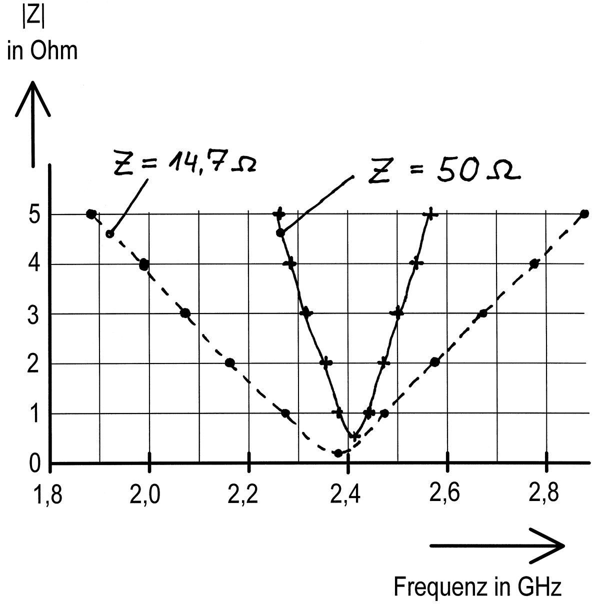


Fig. 7: Impedance curve Z = f(frequency) for the two line sections investigated (Frequenz = Frequency)

4.
Earthing Through Radial Line Stubs

Now let us look more closely at the mysterious circuit segments which are found in nearly all microwave circuits from approximately 2 GHz upwards. Fig. 8 shows such a layout, from type application note 1091. How do such layouts work, and what advantages do they offer?

Interestingly, they are used by almost all microwave circuit developers, although there is scarcely any literature about it. Only an application note from HP [2] taken from the Internet gives them their correct name (radial line stubs) and names some literature sources [3], [4].

It does, though, become apparent very quickly that the calculation formulae given there are very lavish and are tailored more for use on mainframe computers. The HP note also describes the application in a very precise and enlightening manner:

"...problems of location and parasitics of low impedance shunt stubs were solved by using fan-shaped open stubs with the narrow end connected to the main transmission line..."

Consequently, we can introduce a low-Ohmic resistance at a three-dimensionally limited point in the circuit and that is exactly what we want for earthing. I am reminded here of the concept of tapering, which in the past was an important method for broad-band transformations and matchings. We are referring here to lines or cavity conductors which continuously alter their dimensions and thus their wave resistance,

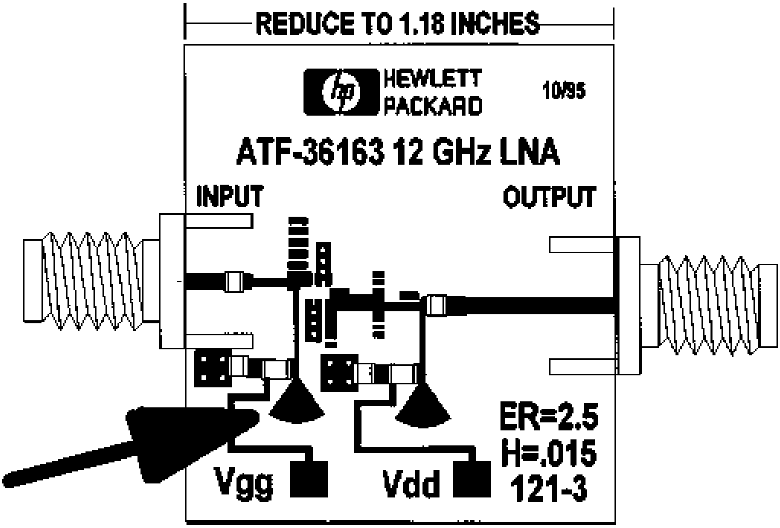


Fig. 8: Application example for radial line stubs (12 GHz LNA)

with the usable frequency range being markedly increased through this measure and the line length required for this reduced.

First a few basic observations:

a) Because the line end is under no-load here too, the associated series resonance must consequently be introduced somewhere (recognisable at the 180° phase angle at S11) with the short circuit at the input. The knife-sharp tip of the structure is very advantageous here, so that the earthing can be positioned precisely at the desired spot on the printed circuit board.

b) If we make the central angle very small, the construction becomes more and more like a normal but narrow and also high-Ohmic microstrip line, with its narrow-band resonance. So increasing the central angle must produce the opposite effect, namely broad-band low impedance.

c) And, assuming the function of tapering has been described correctly, the layout would even have to turn out markedly shorter than a standard λ/4 line!

We can now use a little trick in our investigation with Puff:

One of these radial line stubs is chopped up into many short line sections, but all with the same length, which are wired in series. The line width of each partial section is then selected to be small enough to obtain the best approximation to the original structure.

For a frequency f = 2,400 MHz, we select the radius of the associated full circle to be precisely the same as the λ/4 line length in the previous example, i.e. approximately 16 mm.. Let a piece be cut out of the full circle with a central angle of 45 degrees.

It is helpful if the entire structure can be

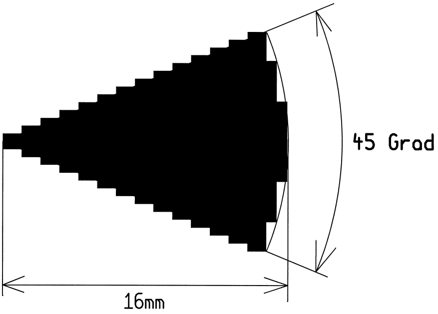
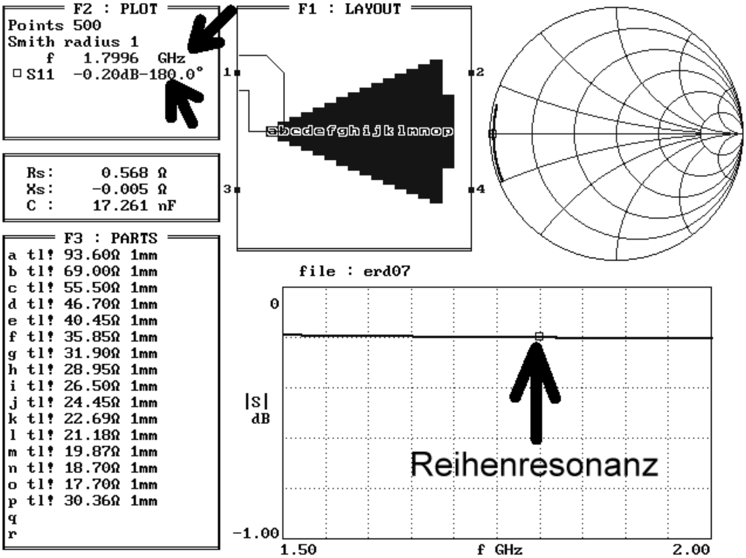


Fig. 9: Radial line stub assembled from 16 elements (Grad = Degrees)

drawn out with a scale of 10:1 on millimetre paper or squared paper, and the dimensions determined for the strip line conductor widths required or they can simply be worked out on a pocket calculator.

We are working here with 16 elements in all (N.B.: you can list a maximum of only 18 components in field F3 of Puff!). Each element has a length of 1 mm. and the widths are graduated in such a way that the best approximation to the ideal stub is obtained (Fig. 9). The following individual data for the simulation can be measured out from the drawing of the line sections or calculated with the help of the geometrical formulae. The numbering for the line sections starts on the left, i.e. at the tip of the circle sector (Table 1).

Now the hard work begins. You can certainly key in the lengths of the microstrip line sections in the Puff F3 field directly in mm. but not their mechanical widths. To do that, you first have to assign any impedance to each of the 16 sections and vary it until the desired conductor width is set as a

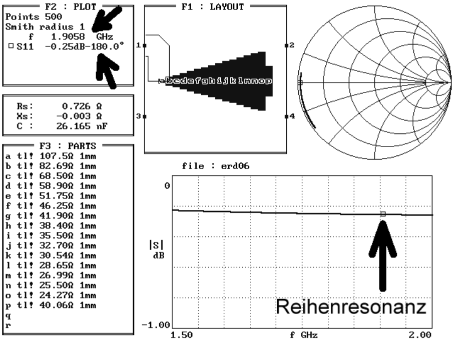


```
F2 : PLOT
Points 500
Smith radius 1
   f    1.7996   GHz
□ S11   -0.20dB-180.0°
```

```
Rs:      0.568 Ω
Xs:     -0.005 Ω
C :     17.261 nF
```

```
F3 : PARTS
a tl! 93.60Ω 1mm
b tl! 69.00Ω 1mm
c tl! 55.50Ω 1mm
d tl! 46.70Ω 1mm
e tl! 40.45Ω 1mm
f tl! 35.85Ω 1mm
g tl! 31.90Ω 1mm
h tl! 28.95Ω 1mm
i tl! 26.50Ω 1mm
j tl! 24.45Ω 1mm
k tl! 22.69Ω 1mm
l tl! 21.18Ω 1mm
m tl! 19.87Ω 1mm
n tl! 18.70Ω 1mm
o tl! 17.70Ω 1mm
p tl! 30.36Ω 1mm
q
r
```

Fig. 10: Radial line stub with a central angle of 45°: S 11 curve between 1.5 and 2.5 GHz (Reihenresonanz = Series resonance)

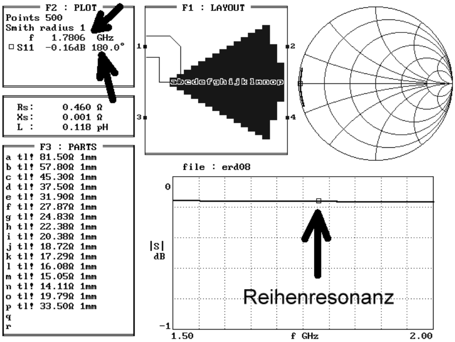


```
F2 : PLOT
Points 500
Smith radius 1
    f   1.9058  GHz
□ S11  -0.25dB-180.0°

  Rs:     0.726 Ω
  Xs:    -0.003 Ω
  C :    26.165 nF

F3 : PARTS
a tl! 107.5Ω 1mm
b tl! 82.69Ω 1mm
c tl! 68.50Ω 1mm
d tl! 58.90Ω 1mm
e tl! 51.75Ω 1mm
f tl! 46.25Ω 1mm
g tl! 41.90Ω 1mm
h tl! 38.40Ω 1mm
i tl! 35.50Ω 1mm
j tl! 32.70Ω 1mm
k tl! 30.54Ω 1mm
l tl! 28.65Ω 1mm
m tl! 26.99Ω 1mm
n tl! 25.50Ω 1mm
o tl! 24.27Ω 1mm
p tl! 40.06Ω 1mm
q
r
```

Fig. 11: S 11 curve with a 30° central angle (Reihenresonanz = Series

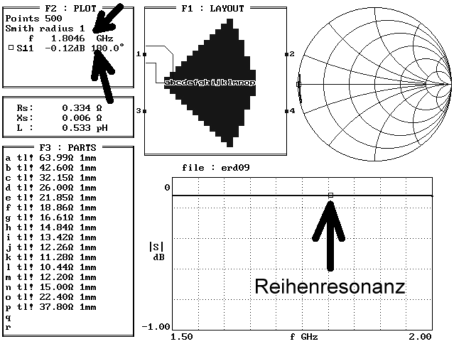


```
F2 : PLOT
Points 500
Smith radius 1
    f   1.7806  GHz
□ S11  -0.16dB 180.0°

  Rs:     0.460 Ω
  Xs:     0.001 Ω
  L :     0.118 pH

F3 : PARTS
a tl! 81.50Ω 1mm
b tl! 57.80Ω 1mm
c tl! 45.30Ω 1mm
d tl! 37.50Ω 1mm
e tl! 31.90Ω 1mm
f tl! 27.87Ω 1mm
g tl! 24.83Ω 1mm
h tl! 22.38Ω 1mm
i tl! 20.38Ω 1mm
j tl! 18.72Ω 1mm
k tl! 17.29Ω 1mm
l tl! 16.08Ω 1mm
m tl! 15.05Ω 1mm
n tl! 14.11Ω 1mm
o tl! 19.79Ω 1mm
p tl! 33.50Ω 1mm
q
r
```

Fig. 12: S 11 curve with a 60° central angle (Reihenresonanz = Series resonance)



```
F2 : PLOT
Points 500
Smith radius 1
    f   1.8046  GHz
□ S11  -0.12dB 180.0°

  Rs:     0.334 Ω
  Xs:     0.006 Ω
  L :     0.533 pH

F3 : PARTS
a tl! 63.99Ω 1mm
b tl! 42.60Ω 1mm
c tl! 32.15Ω 1mm
d tl! 26.00Ω 1mm
e tl! 21.85Ω 1mm
f tl! 18.86Ω 1mm
g tl! 16.61Ω 1mm
h tl! 14.84Ω 1mm
i tl! 13.42Ω 1mm
j tl! 12.26Ω 1mm
k tl! 11.28Ω 1mm
l tl! 10.44Ω 1mm
m tl! 12.20Ω 1mm
n tl! 15.00Ω 1mm
o tl! 22.40Ω 1mm
p tl! 37.80Ω 1mm
q
r
```

Fig. 13: Radial line stub with a 60° central angle (Reihenresonanz =

reaction when the equals sign is keyed in. Then the stub structure is assembled in field F1, and the simulations are begun for the range 1.5 2.5 GHz.

Not only does Fig. 10 show the complete list of stripline sections, together with their associated impedances, in field F3, but also the resonance frequency of the earthing (= reactive component precisely 0 Ohms, and thus phase angle of 180° at S11) can be read off at approximately 1.8 GHz. Thus, for a central stub angle of 45°, we have a mechanical reduction of the layout of at least approximately 25%, as against the standard λ/4 line.

Just for interest, the calculations and

Leitungsstück Nr.:	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16
Breite in mm	0,816	1,63	2,45	3,265	4,08	4,90	5,71	6,53	7,35	8,16	8,98	9.80	10,61	11,43	12,24	6,12
Länge in mm	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0

Table 1: (Leitungstuck = Line section, Breite = Width, Länge = Length

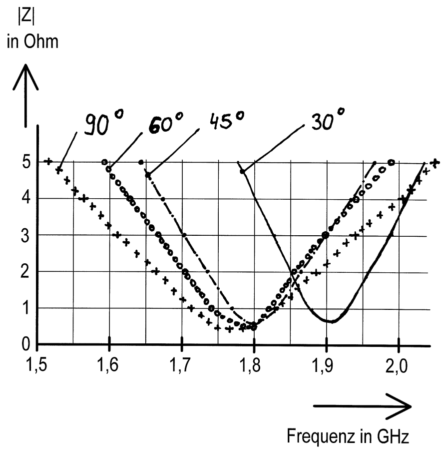


Fig. 14: Z = f(frequency) for radial line stubs with 30/45/60 and 90° central angles (Frequenz = Frequency)

simulations were repeated again for the central angles 30°, 60° and 90°, in order to obtain a feel for optimal dimensioning. Figs. 11, 12, 13 now show both the required impedances of the line sections for the three central angles investigated and also the associated simulations with the resonance frequencies, which now differ only insignificantly.

To be able to carry out a direct comparison between the characteristics of the 4 stub versions investigated, a summary diagram was drawn up for the impedance curve of the four different stub versions, using Puff, in the frequency range from 1.5 to 2 GHz (Fig. 14). If we look closely at this picture, we can obtain the following knowledge from it:

a) With a central angle of 30°, the taper effect is already weakening markedly, and the behaviour corresponds more and more to the normal λ/4 line. This can be seen, above all, from the fact that the reduction is no longer so effective and the band width is diminishing.

b) Between 45° and 90°, the reductions are almost identical in the order of 25% - but the broad-bandness naturally increases with the central angle, as was conjectured.

Practical tips:

As a developer, you can lay in a stock of such radial stubs for the main frequency ranges in which you work.

The procedure required for this can be demonstrated using the example of the design frequency of 2,400 MHz already used previously.

Step 1:

Decide on a central angle between 45° and 90° - for example, 60°.

Step 2:

The design frequency can now be selected to be higher by a factor of 1.25. Thus we can enter 3.0 GHz in field F4.

Step 3:

Decide on the mechanical length required for a standard λ/4 line for the printed circuit board material used and for the frequency of 3 GHz. This gives the radius of the full circle from which the stub will be cut out.

Step 4:

Now sketch the stub, replace it by a series of max. 18 series-wired microstrip line sections see above and enter these sections in field F3. Then comes the wearisome task of determining the matching impedance for each line section.

Step 5:

Now assemble the stub in field F1 from the individual line sections and then start the simulation for S 11.

N.B.:

Don't be afraid of carrying out two separate simulations, one for a narrow band width of 2 to 3 GHz and another for a broad-band range of 0 to 10 GHz. Thus for a subsequent practical application you can have the option of investigating its characteristics even outside the range for which it is intended to be used and, for example, checking for oscillation tendencies. If you have used only the narrow-band version of stub here, Puff immediately reacts with a corresponding error message if the preset frequency range is exceeded.

Step 6:

Now switch to a text processing program and print out the *.puf-file just produced, together with any device file existing for an FET or transistor from the Puff directory. If we lay the two print-outs directly next to one another, we can recognise which parts of the stub file must be altered or deleted to make a device file from it.

The device file created in this way can then be stored under a suitable name (ending in *.dev) in the Puff directory. It will then be available as a component for future campaigns.

Step 7:

If, though, you want to turn your attention to another frequency range, you have to repeat the entire procedure using the corresponding data.

Anyone who, like me, tends to make more and more use of the ARRL Radio Designer (in parallel with Puff) to determine the noise figure, the stability factor, k, or an impedance curve, etc., will now have rather more work to do. You can certainly integrate the S-parameter file for the stub generated by Puff into the Radio Designer, but it does need rather more extensive preliminary work:

Not only do all superfluous parts of the Puff file have to be deleted but the unit GHz has to be added to every line of the S-parameter listed following the frequency value. The altered file is then copied into the ARD file currently being worked on to do this, see [5].

5.

Literature

[1] Kraus, Gunthard, DG8GB Design and realisation of microwave circuits, Part 9. VHF Reports, no. 2 / 1998, pp. 119 ff.

[2] Broadband Microstrip Mixer Design Application Note 976 from Hewlett Packard. On Internet under: http://www.hp.com/HP-COMP/rf

[3] Vinding, J.R. Radial Line Stubs as Elements in Strip Line Circuits. Nerem Record, pp. 108-109, 1967

[4] Schneider, M.V. Microstrip Lines for Microwave Integrated Circuits. Bell System Technical Journal, Vol. 48, no. 5, May-June, 1969

[5] Kraus, Gunthard, DG8GB and Zimmermann, Andreas, DG3SAZ. Low-noise pre-amplifier for 137 MHz NOAA weather satellite range and / or 145 MHz 2-m. amateur radio band / Part 2. VHF Reports, no. 1 / 1999, pp. 37 ff.