Development stages and final designs of the impedance measurement cells with the SOLT and TRL calibrations

Straight and Dog bone impedance measurement cells

These cells will be used for measurements of the dependence of the high frequency impedance in ferromagnetic wires on the external factors such as magnetic field, temperature, or tensile stress. Both types of the cells can be calibrated using two methods: SOLT and TRL.

SOLT calibration (SHORT, OPEN, LOAD, THRU) with the operation frequency range up to 8 GHz

This calibration method uses three miniature termination standards placed at the ends of the microstrips: SHORT, OPEN, and LOAD 50 Ω. An additional standard THRU will be used as well. Note that Rohde&Schwarz names SOLT as TOSM – THRU, OPEN, SHORT, and MATCH (equals to LOAD).

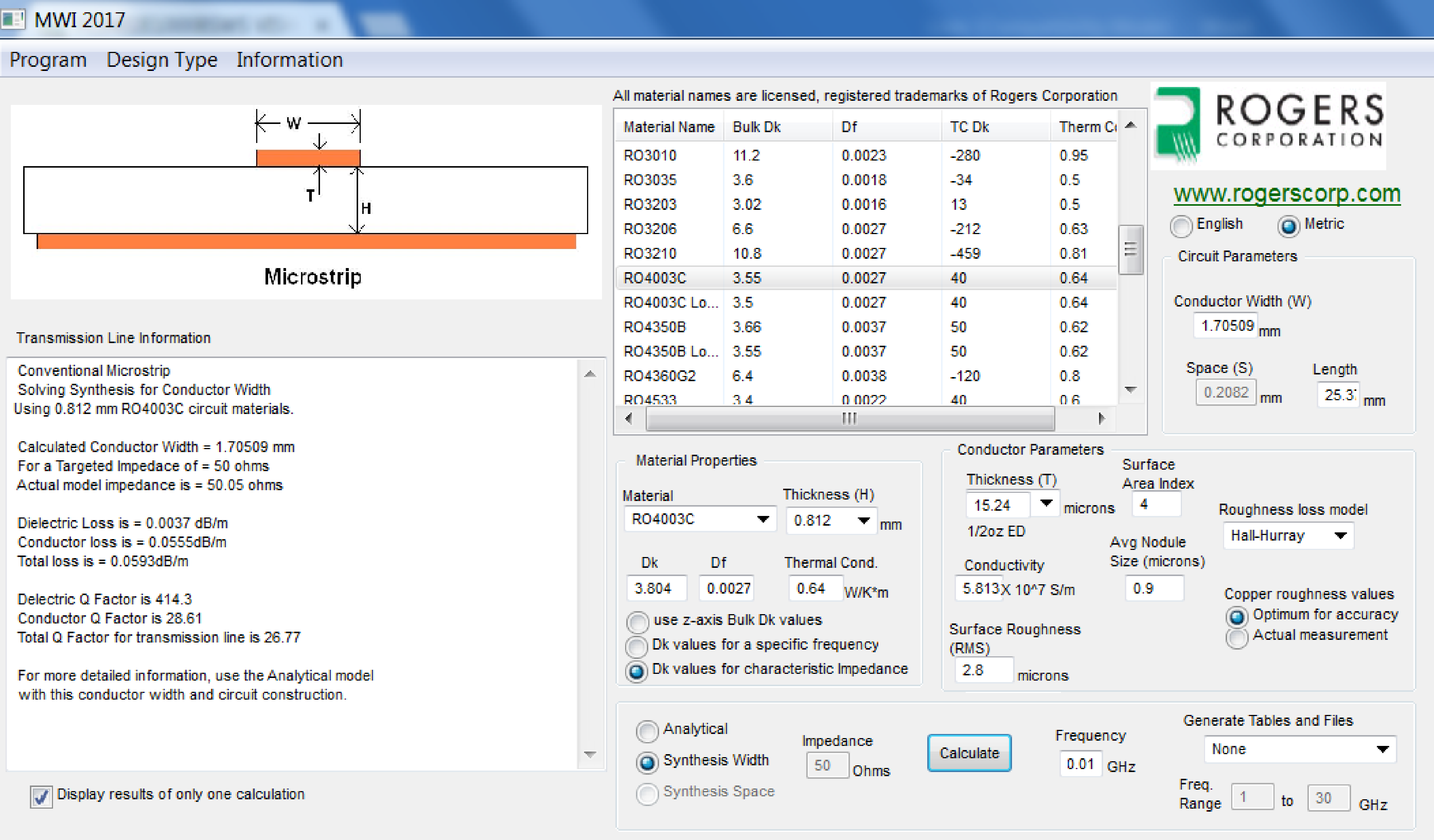
TRL calibration (THRU, REFLECT, LINE) with the operation frequency range 250 MHz – 18 GHz

This calibration method uses three standards: ideal THRU (zero loss and zero electrical length), REFLECT (similar to OPEN in SOLT), and LINE which introduces 20^0 -160^0 phase shift with respect to THRU. In the microwave community, TRL calibration is considered to be most accurate when measuring on the wafers. The only disadvantage of TRL is inability to fabricate very long LINE standards at lower frequencies.

Rogers microwave impedance calculator

It can be run online or free downloaded from Rogers Inc web site:
https://www.globalcommhost.com/rogers/acs/techsupporthub/en/calculatorMWI.php

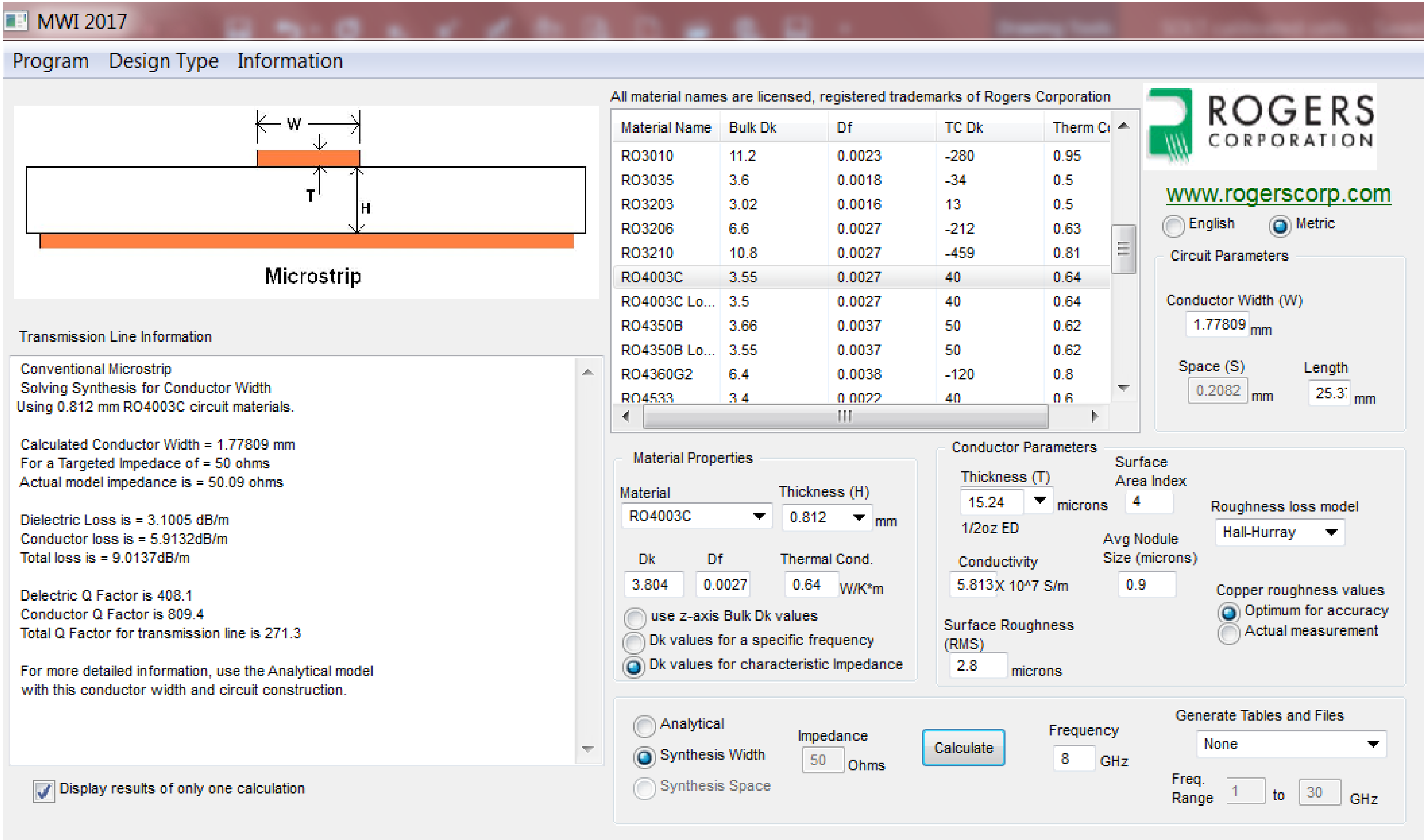
We decided to use Rogers PCBs due a high predictability and stability of their dielectric properties in a wide frequency range. The microstrip parameters will be the same for both the straight and dog bone cells. However, due to a slight frequency dispersion of dielectric properties, the microstrip width cannot be saved constant for all frequencies. We calculate an averaged microstrip width for the frequencies up to 8 GHz – most important range for our applications.



Chosen material: RO4003C

Its thickness: 0.812 mm

Strip width W at the frequency 10 MHz: 1.70509 mm

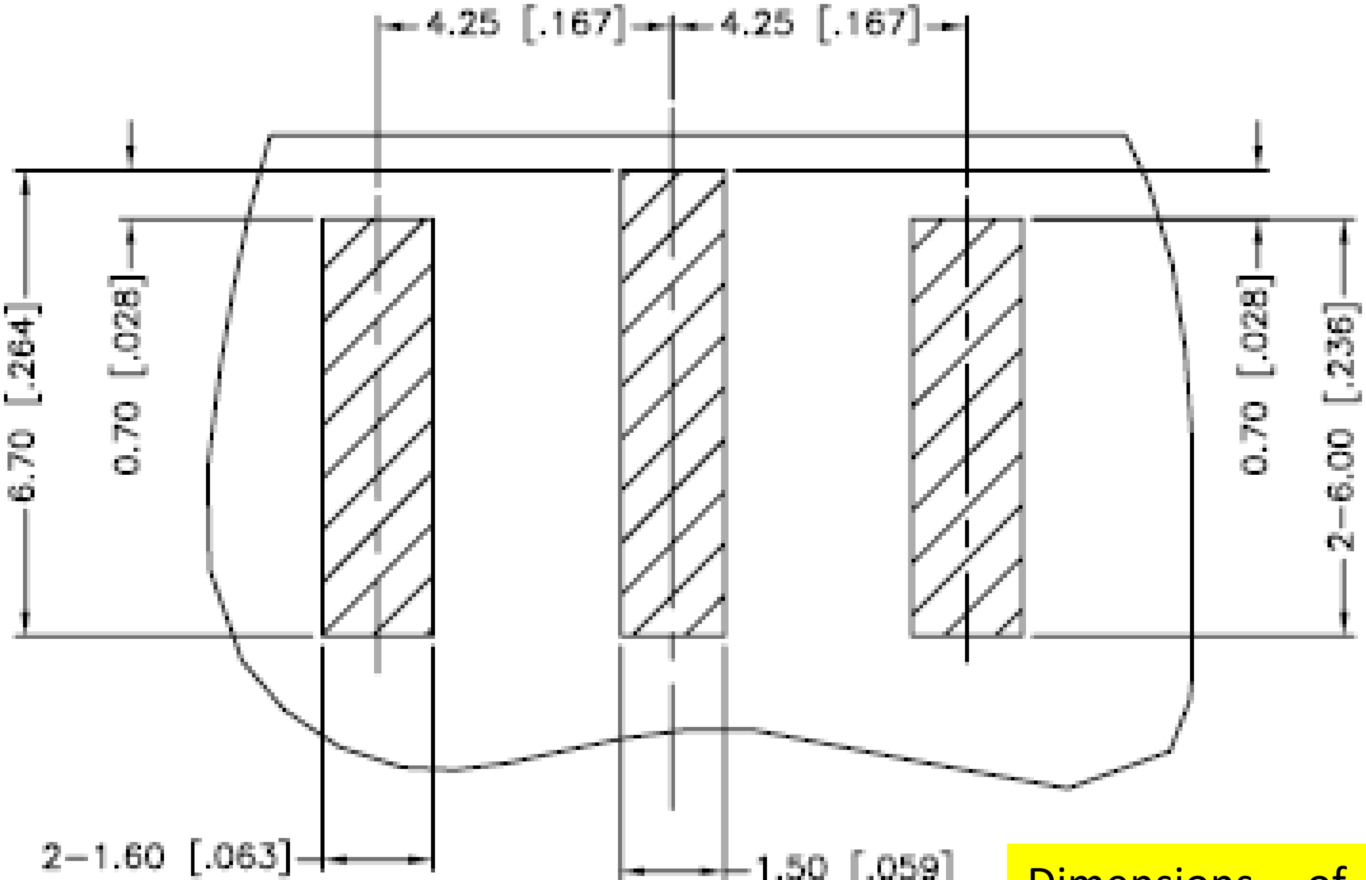


Chosen material: RO4003C

Its thickness: 0.812 mm

Strip width W at the frequency 8 GHz: 1.77809 mm

Using two previous slides, we choose an averaged strip width 1.74159 мм to operate up 8 GHz for RO4003C. Actually, this method of calibration will also work for higher frequencies, but its accuracy will gradually degrade. The chosen connector type: SMA Amphenol 132432. It will be bought from Mouser UK: https://www.mouser.co.uk/ProductDetail/Amphenol-RF/132432?qs=sGAEpiMZZMuLQf%252bEuFsOrk%2fjxZFCvxCSdqBupOAx9pA%3d

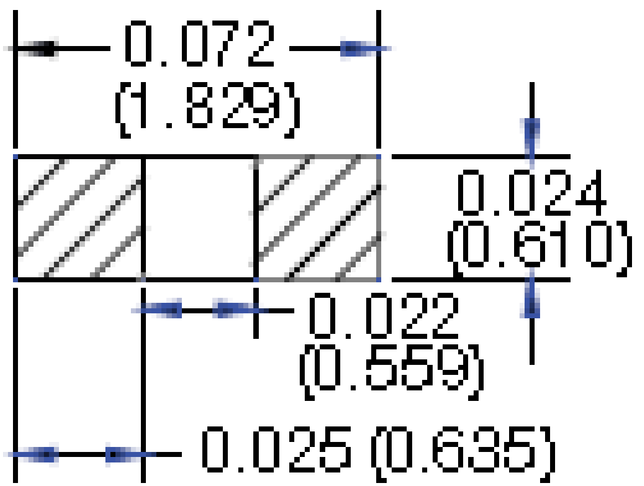


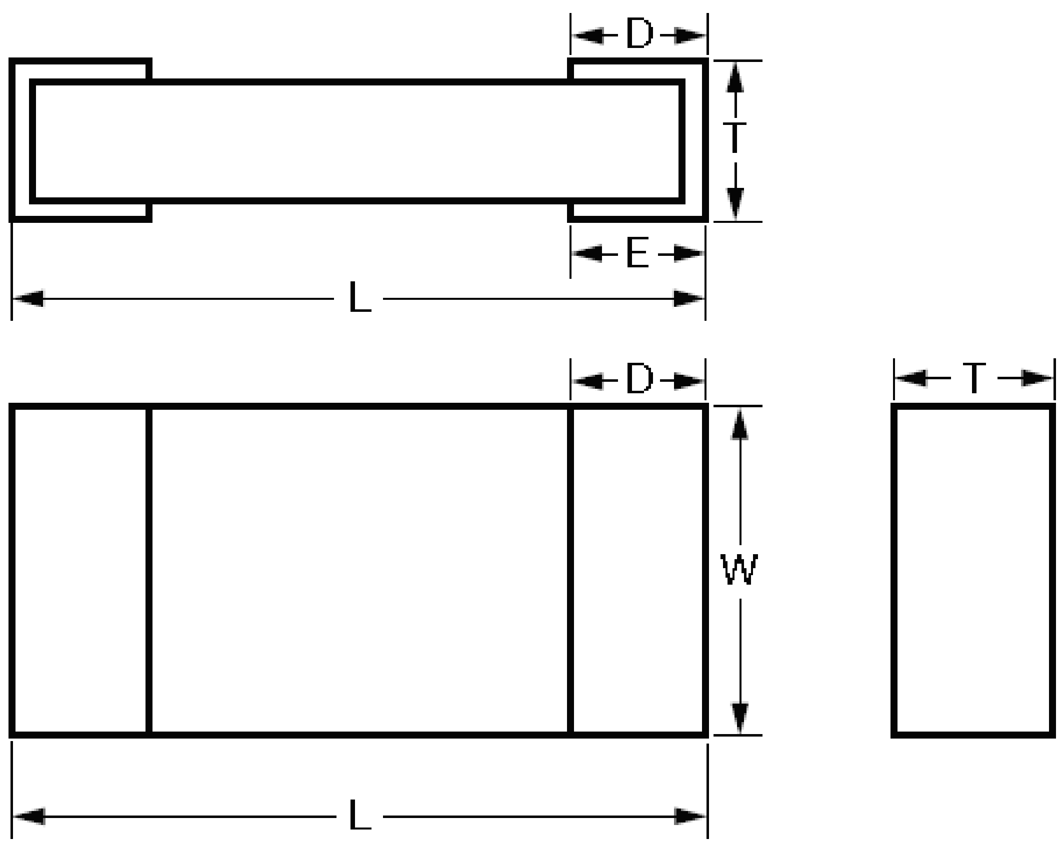
Dimensions of the SMA footprint recommended by Amphenol. Sizes are given in mm and [inches].

For the LOAD termination, we will use two 100 Ω Vishay RF resistors (up to 20 GHz), connected in parallel. Such method allows a reduced parasitic inductance at higher frequencies. The resistors will be bought from Farnell UK: http://uk.farnell.com/vishay/fc0402e1000bsws/resistor-microwave-100r-0402/dp/1109051

CASE SIZE	LENGTH	WIDTH W (± 0.005)	THICKNESS	TOP PAD D (± 0.005)	BOTTOM PAD E (± 0.005)
0402	0.042 ± 0.008 (1.067 ± 0.203)	0.022 (0.559)	0.015 to 0.0015 (0.381 to 0.0381)	0.010 (0.254)	0.010 (0.254)

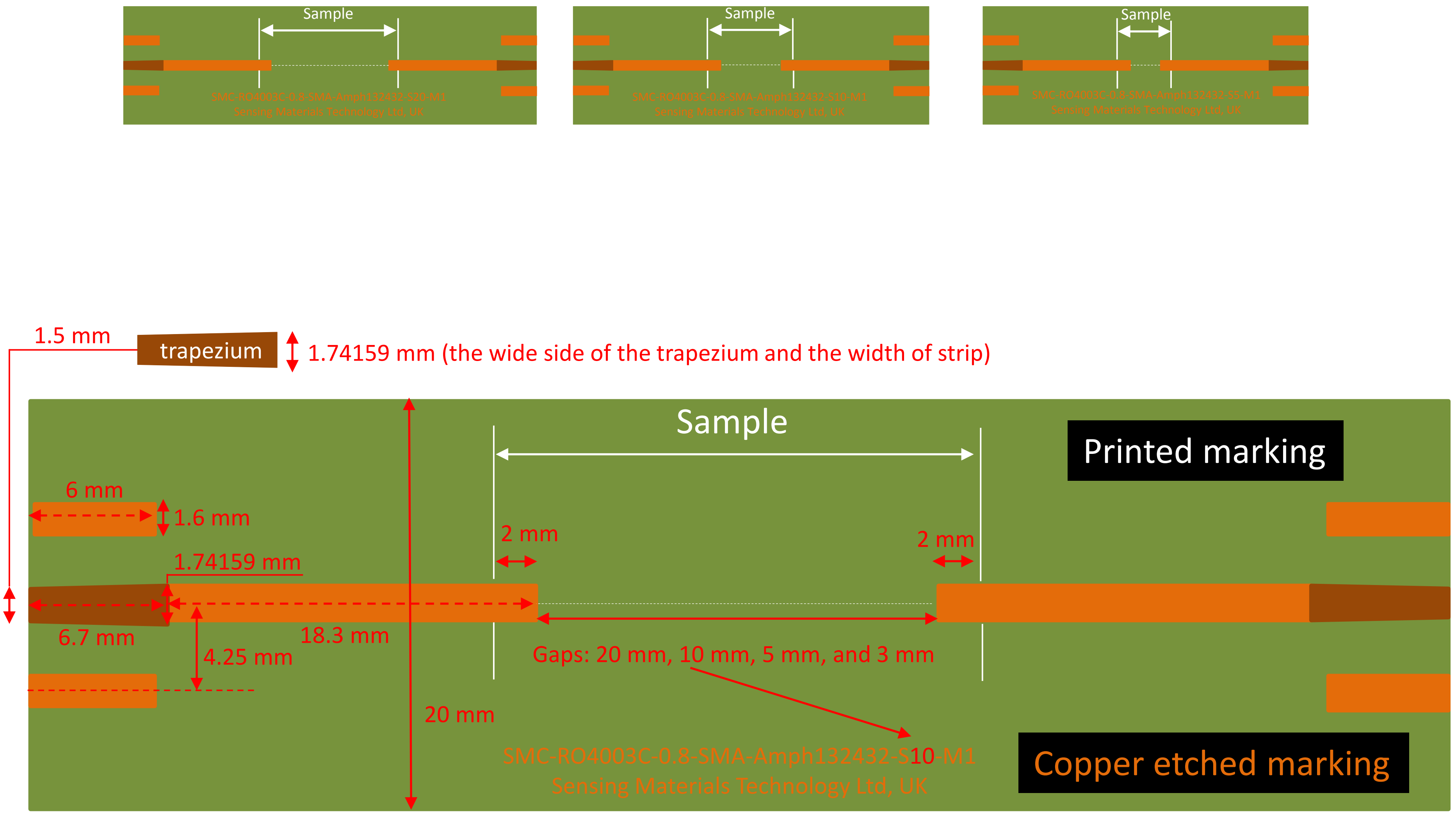
0402 Land Pattern





Sizes in in the brackets are given in mm.

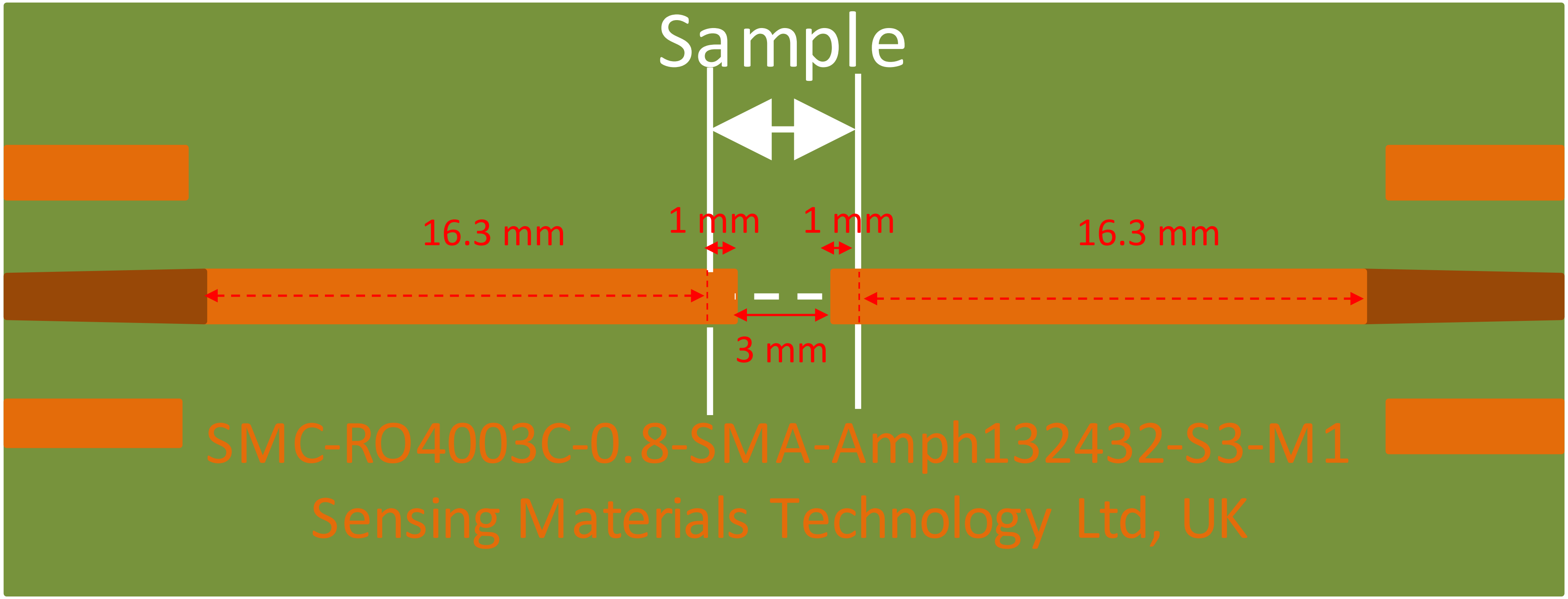
Straight RF cells for the impedance measurements in the presence of magnetic field or heat



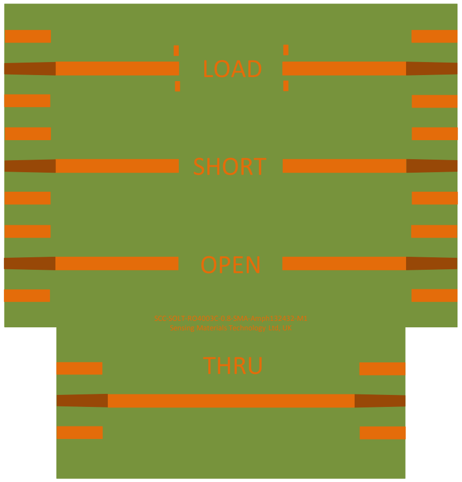
Sample
SMC-RO4003C-0.8-SMA-Amph132432-S20-M1
Sensing Materials Technology Ltd, UK
Sample
SMC-RO4003C-0.8-SMA-Amph132432-S10-M1
Sensing Materials Technology Ltd, UK
Sample
SMC-RO4003C-0.8-SMA-Amph132432-S5-M1
Sensing Materials Technology Ltd, UK
1.5 mm
trapezium
1.74159 mm (the wide side of the trapezium and the width of strip)
Sample
Printed marking
6 mm
1.6 mm
2 mm
2 mm
1.74159 mm
6.7 mm
18.3 mm
4.25 mm
Gaps: 20 mm, 10 mm, 5 mm, and 3 mm
20 mm
SMC-RO4003C-0.8-SMA-Amph132432-S10-M1
Sensing Materials Technology Ltd, UK
Copper etched marking

A very “challenging cell” – extremely short sample (3 mm) and soldering pads (1 mm)

After a proper calibration, a VNA will return correctly measured S-parameters of a sample soldered between the contact pads. These S-parameters can be formally recalculated into the impedance Z. However, such recalculation assumes that Z is a lumped parameter. The latter is fully correct only if the wavelength along the sample is much larger than its longitudinal size. Therefore, for the accurate results, especially when measuring the frequency dispersion of Z, the sample length must be significantly reduced. Here, we face a typical problem in microwave measurements – interpretation of results. Note that S-parameters are always correct (after a proper calibration), but their recalculation into something else may require more sophisticated models. In our project, we have to find new approaches to this problem if the sample size is not meant to be extremely small. The measurement cell proposed below still allows traditional recalculation procedure from S21/12-parameter into the lumped Z up to several GHz. In a VNA, such recalculation can be done in real time.



SOLT calibration board



LOAD

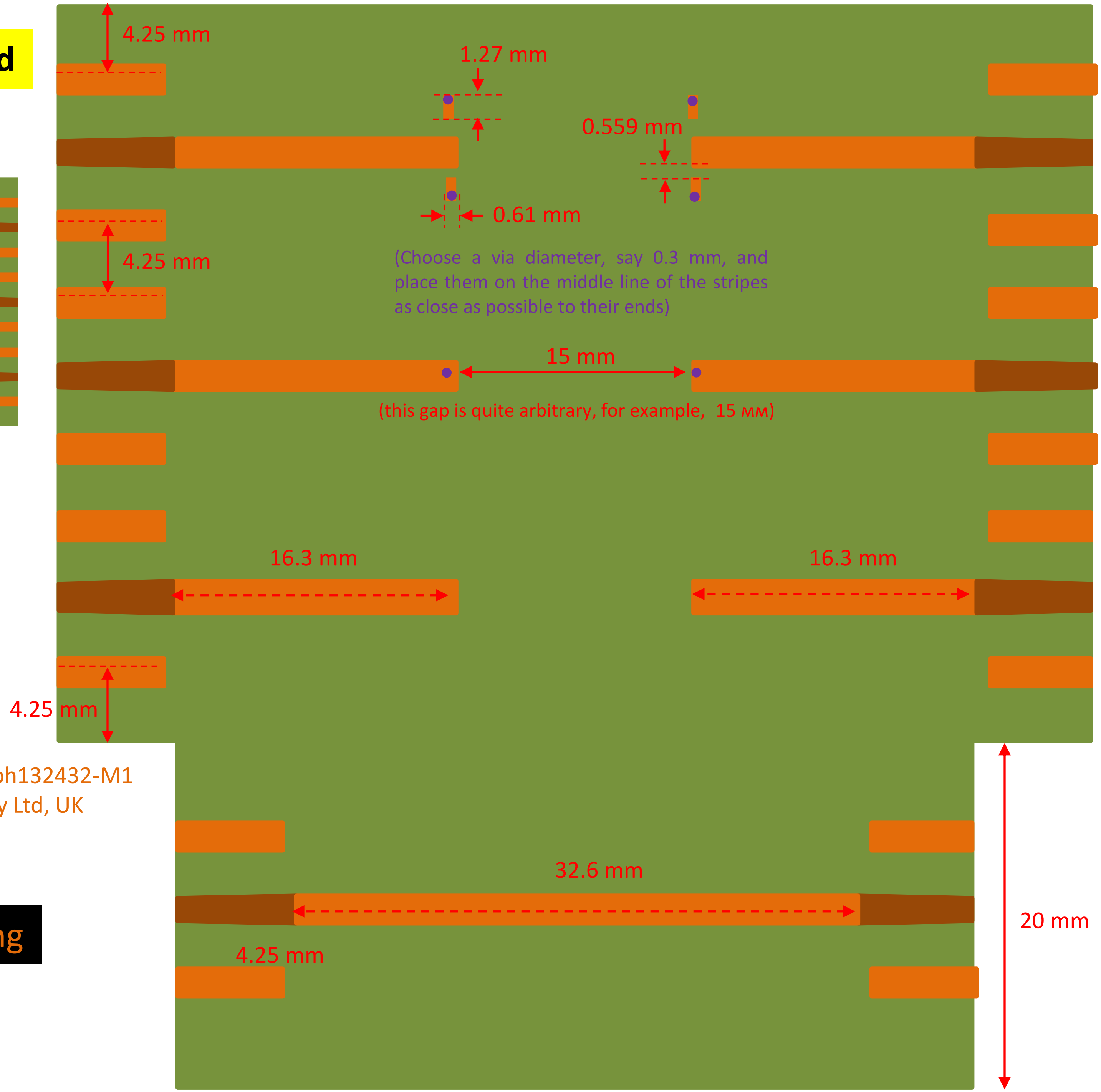
SHORT

OPEN

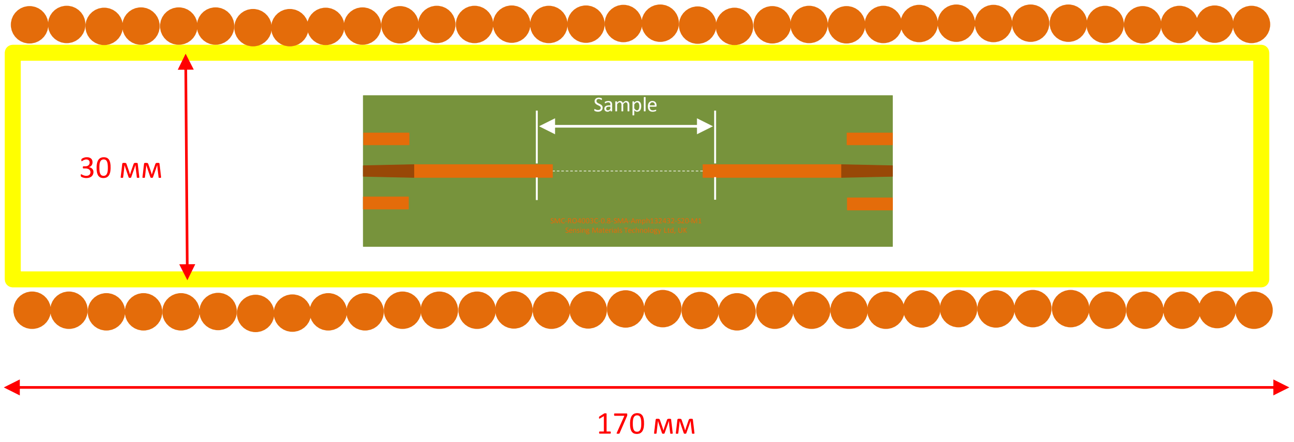
SCC-SOLT-RO4003C-0.8-SMA-Amph132432-M1
Sensing Materials Technology Ltd, UK

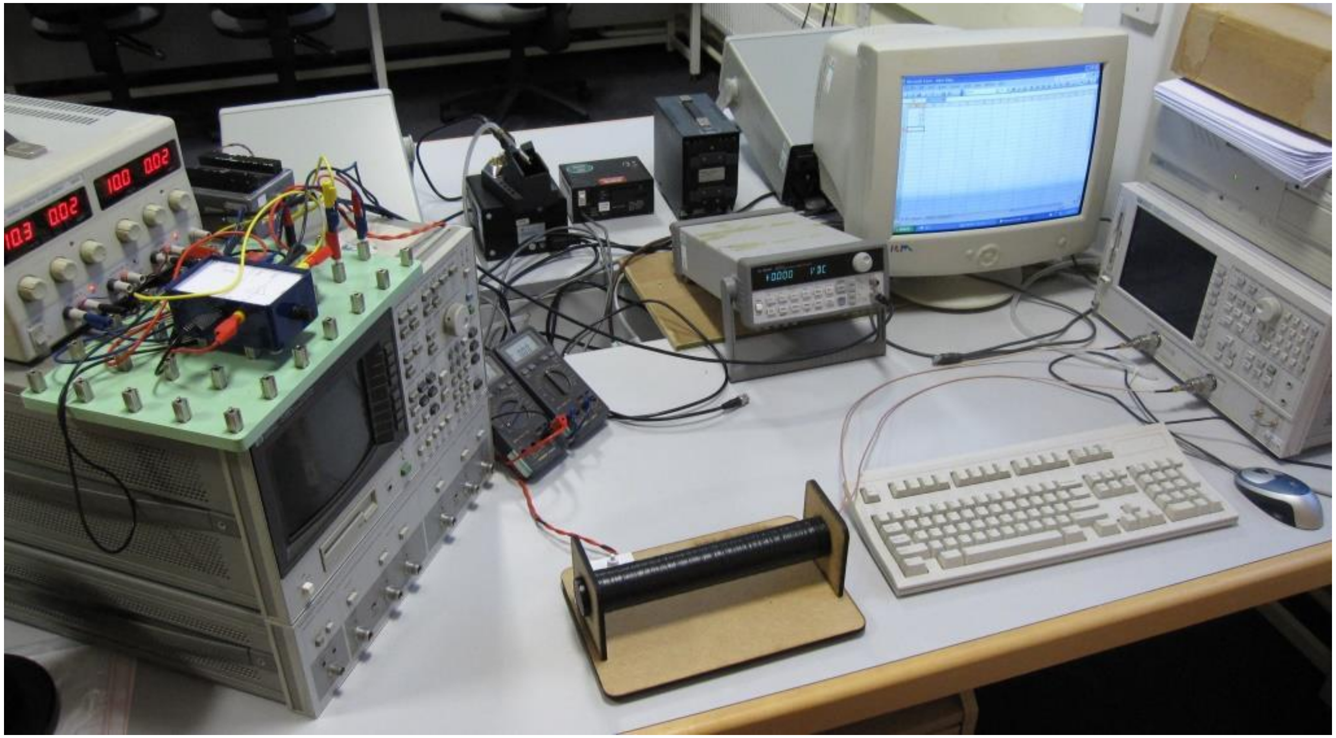
THRU

Copper etched marking

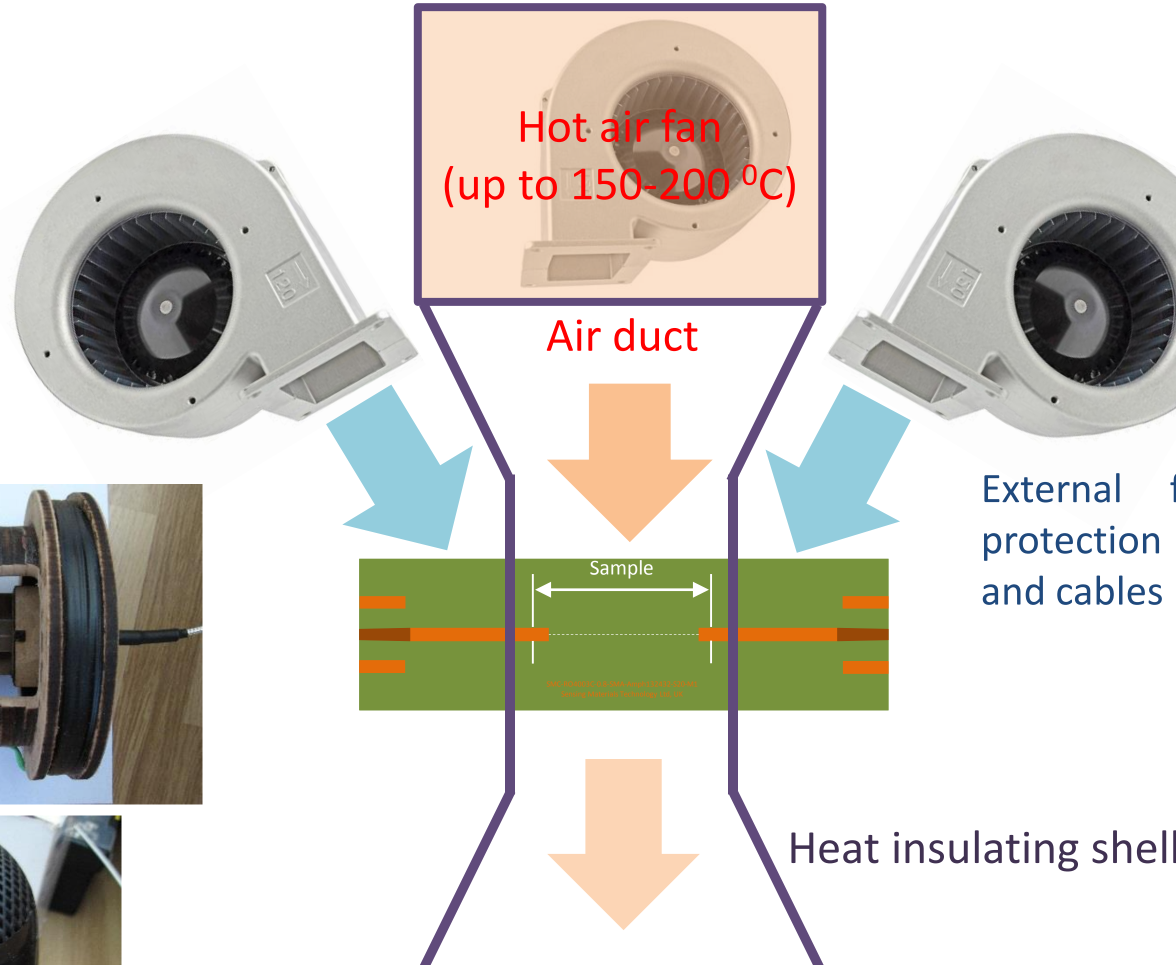


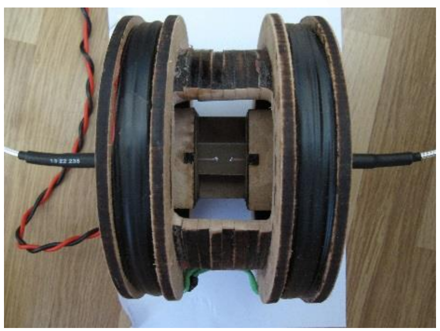
Application: magneto-impedance measurements using a compact solenoid coil





Application: effect of temperature on the magneto-impedance





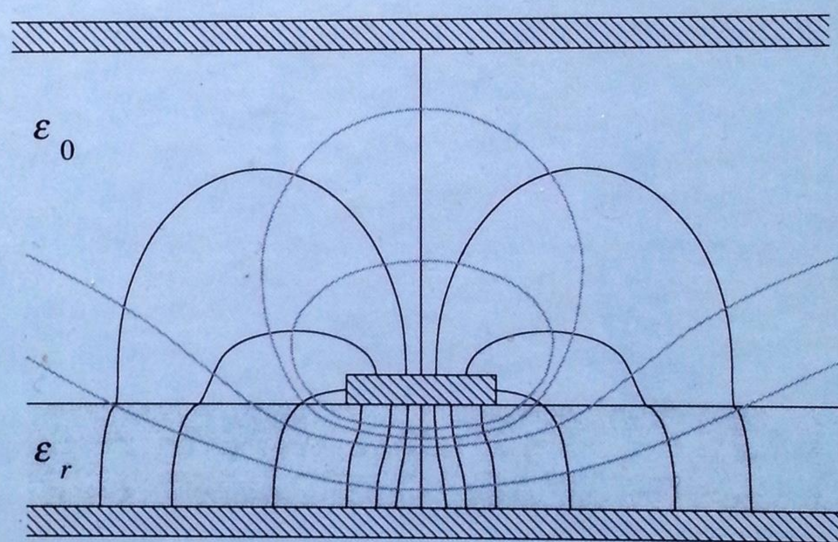
An example of the heat chamber integrated into the Helmholtz coil.

Dog bone RF cells for the stress-impedance measurements

For this type of the cells, we will need a method that allows compensation of the impedance mismatch introduced by a bending of the strip. The necessary information is taken from a book on transmission lines. Also, we can use the online calculator: http://www.finetune.co.jp/~lyuka/technote/ustrip/

Transmission Line Design Handbook

https://www.amazon.com/Transmission-Handbook-Propagation-Microwave-Hardcover/dp/0890064369/ref=sr_1_1?ie=UTF8&qid=1517074427&sr=8-1&keywords=Brian+C.+Wadell



Brian C. Wadell

5.5.1 Microstrip Line Bend

h – thickness of dielectric

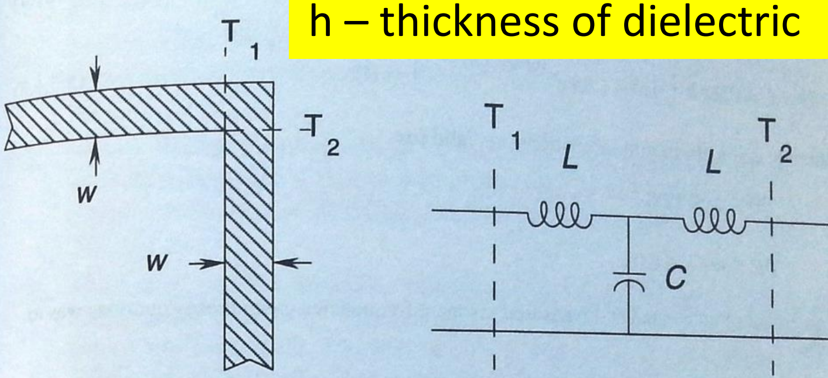


Figure 5.5.1.1: Microstrip Line Abrupt Bend

The discontinuity created by a bend in a microstrip line is given by the model above [2] where:

$$L / h = \frac{100.0}{2.0}\left(4.0 \sqrt{w / h} - 4.21\right) \quad (\mathrm{nH} / \mathrm{m}) \tag{5.5.1.1}$$

and for w / h < 1:

$$C / w = \frac{(14.0\ \varepsilon_r + 12.5)(w / h) - (1.83\ \varepsilon_r - 2.25)}{\sqrt{w / h}} + \frac{0.02\ \varepsilon_r}{w / h} \quad (\mathrm{pF} / \mathrm{m}) \tag{5.5.1.2}$$

or for w / h ≥ 1:

$$C / w = (9.5\ \varepsilon_r + 1.25)(w / h) + 5.2\ \varepsilon_r + 7.0 \quad (\mathrm{pF/m}) \tag{5.5.1.3}$$

These equations are accurate to within 5% for

$$2.5 \le \varepsilon_r \le 15.0$$

$$0.1 \le w / h \le 5.0$$

Kirschning [7] gives equations for C and L as

$$C = 0.001\ h \left[(10.35\ \varepsilon_r + 2.5)(w / h)^2 + (2.6\ \varepsilon_r + 5.64)(w / h) \right] \quad (\mathrm{pF}) \tag{5.5.1.4}$$

$$L = 0.22\ h \left[1.0 - 1.35\ e^{-0.18 (w / h)^{1.39}} \right] \quad (\mathrm{nH}) \tag{5.5.1.5}$$

where w and h are in mm. Equations are valid for

$$2.0 \le \varepsilon_r \le 13.0$$

$$0.2 \le w / h \le 6.0$$

Agreement of calculated and measured resonant frequencies using these equations was to 0.3%.

REFERENCES

[1] Douville, Rene J.P., and David S. James, "Experimental Study of Microstrip Bends and Their Compensation," *IEEE Transactions on Microwave Theory and Techniques*, Vol. MTT-26, No. 3, March 1978, pp. 175–181.

[2] Garg, Ramesh, and I.J. Bahl, "Microstrip discontinuities," *International Journal of Electronics*, Vol. 45, No. 1, 1978, pp. 81–87.

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5.5.2 Optimal Right-Angle Mitered Bend

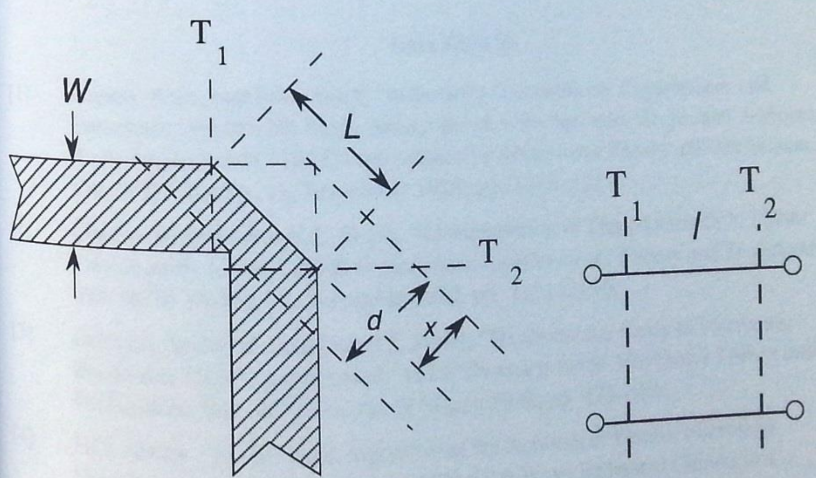


Figure 5.5.2.1: **Optimal Microstrip Line Mitered Bend**

The optimal mitre occurs for:

$$M = 52 + 65\,e^{-1.35\,w/h} = \frac{100\,x}{d} \quad (\%) \qquad (5.5.2.1)$$

http://www.finetune.co.jp/~lyuka/technote/ustrip/

or, more conveniently,

$$\frac{L}{w} = \sqrt{2}\,[1.04 + 1.3\,e^{-1.35\,w/h}] \qquad (5.5.2.2)$$

$$d - x = \sqrt{2}\,w - \frac{L}{2} \qquad (5.5.2.3)$$

These equations are valid for $w/h \geq .25$, $\varepsilon_r \leq 25$, ±4 (sic, %?) accuracy. In Chadha [2], the optimal mitre is found to be $w/h = 2.895$. Chadha also recommends that for bends made of lines of unequal widths that more material be removed from the wider line. Figure 5.5.3.2 compares the references.

The equivalent electrical length of the mitre, l, is calculable from arbitrary bend equations of [1] by setting the angle α to 90° resulting in

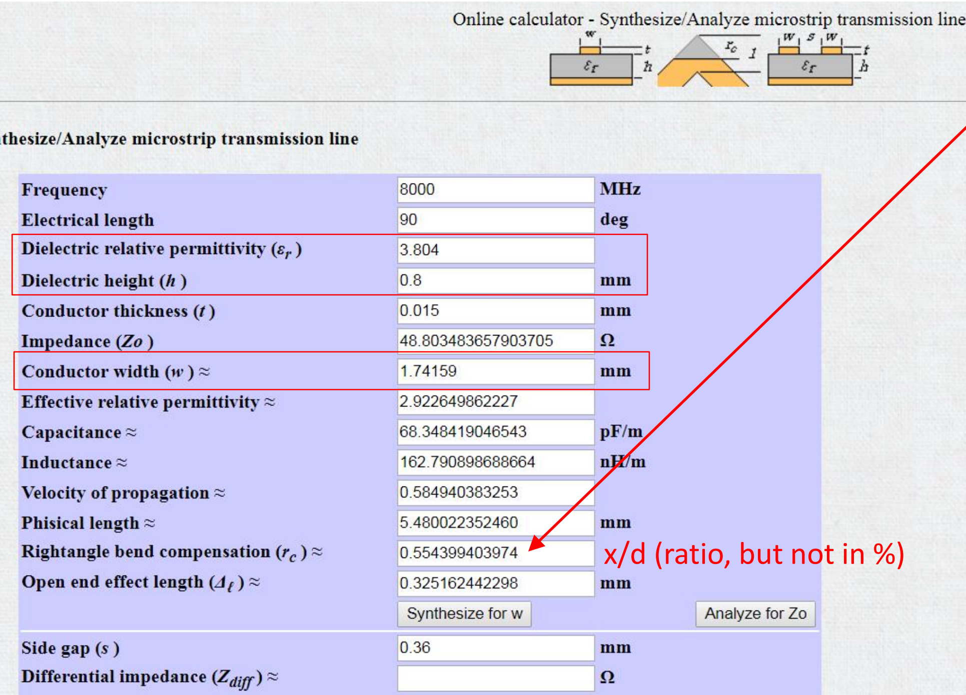
$$l = \frac{L}{\sqrt{2}} \qquad (5.5.2.4)$$

Other techniques of equalizing the bend are shown in Figure 5.5.2.2 below. Analytic equations for these structures are not available. Douville [3] shows that the 45° double bend is always better than a single right-angle bend.

We will use this optimisation

Be careful when using online calculators: sometimes it is not entirely clear what they are calculating

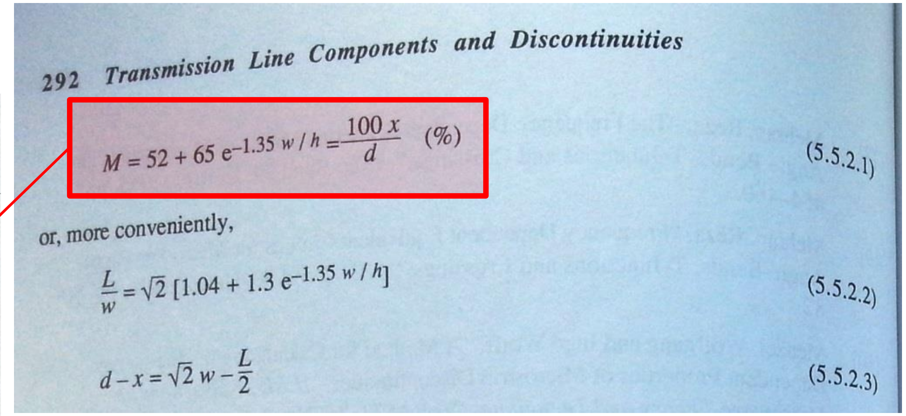
http://www.finetune.co.jp/~lyuka/technote/ustrip/


Online calculator - Synthesize/Analyze microstrip transmission line

thesize/Analyze microstrip transmission line

Frequency	8000	MHz
Electrical length	90	deg
Dielectric relative permittivity (ε_r)	3.804	
Dielectric height (h)	0.8	mm
Conductor thickness (t)	0.015	mm
Impedance (Zo)	48.803483657903705	Ω
Conductor width (w) ≈	1.74159	mm
Effective relative permittivity ≈	2.922649862227	
Capacitance ≈	68.348419046543	pF/m
Inductance ≈	162.790898688664	nH/m
Velocity of propagation ≈	0.584940383253	
Phisical length ≈	5.480022352460	mm
Rightangle bend compensation (r_c) ≈	0.554399403974	
Open end effect length (Δ_ℓ) ≈	0.325162442298	mm
	Synthesize for w	Analyze for Zo
Side gap (s)	0.36	mm
Differential impedance (Z_{diff}) ≈		Ω

x/d (ratio, but not in %)

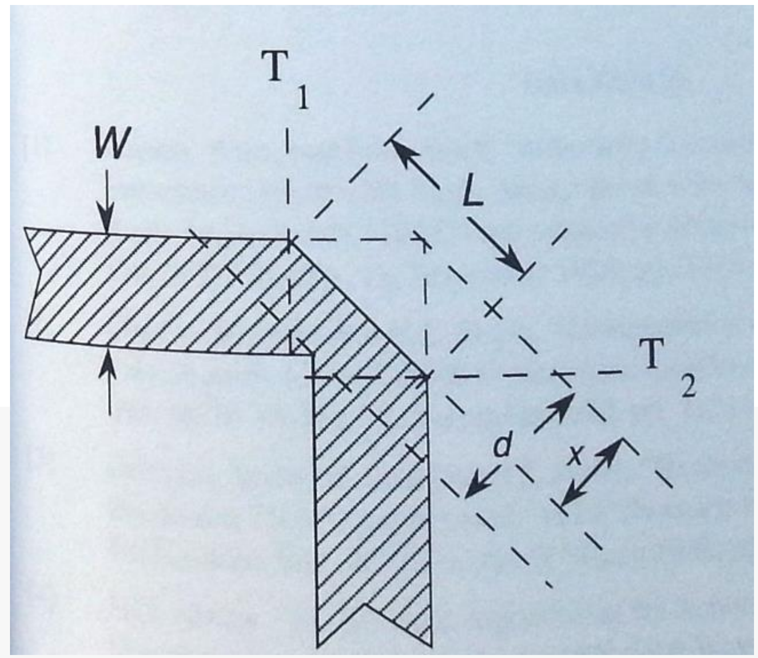

292 *Transmission Line Components and Discontinuities*

$$M = 52 + 65\,e^{-1.35\,w/h} = \frac{100\,x}{d} \quad (\%) \qquad (5.5.2.1)$$

or, more conveniently,

$$\frac{L}{w} = \sqrt{2}\,[1.04 + 1.3\,e^{-1.35\,w/h}] \qquad (5.5.2.2)$$

$$d - x = \sqrt{2}\,w - \frac{L}{2} \qquad (5.5.2.3)$$




```python
# Program in Python
import math
w=1.74159 # strip width in mm
h=0.8 # PCB thickness in mm
d=2**0.5*w # diagonal in mm
x=d*(0.52+0.65*math.exp(-1.35*w/h)) # cut off from the corner in mm
print('x=',x,'mm')
```

x= 1.365474753822523 mm ← This value will be used for our cells

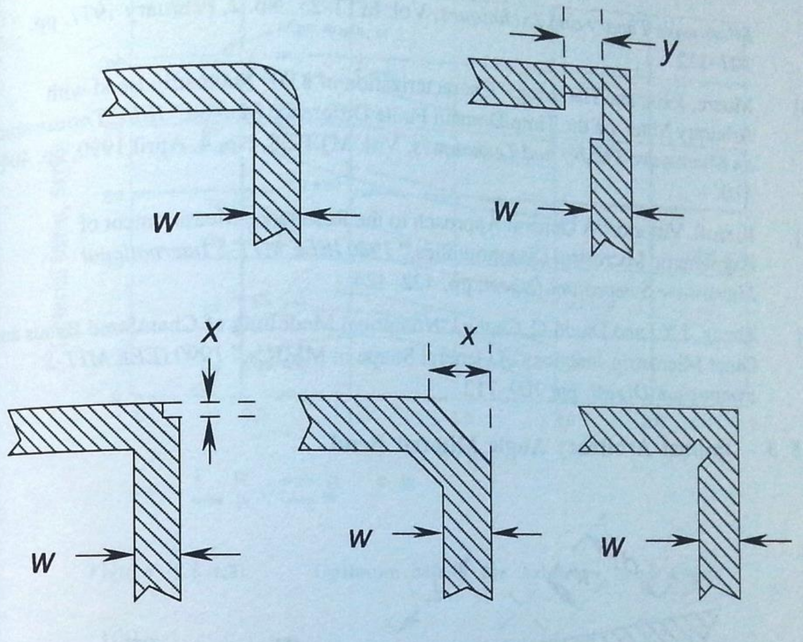


Figure 5.5.2.2: Various Techniques of Bend Compensation

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5.5.3 Optimal Arbitrary Angle Mitered Bend

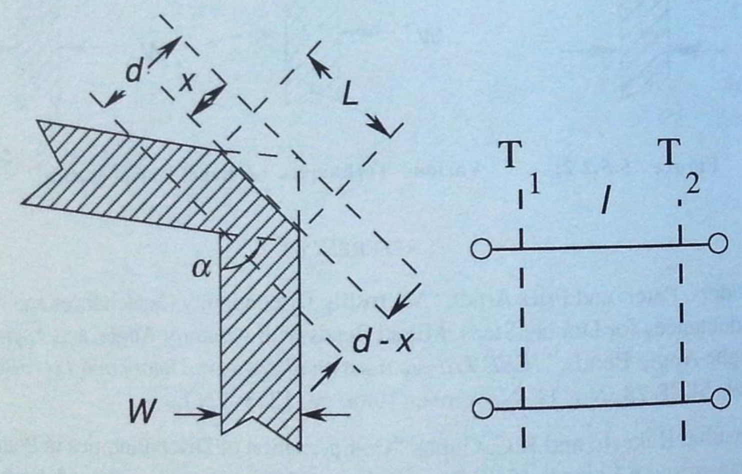

Figure 5.5.3.1: Arbitrary Mitered Bend

The optimal mitre dimensions are shown in Figure 5.5.3.2 for a range of angles and w / h ratios. This figure combines the results of a number of references for comparison and convenience.

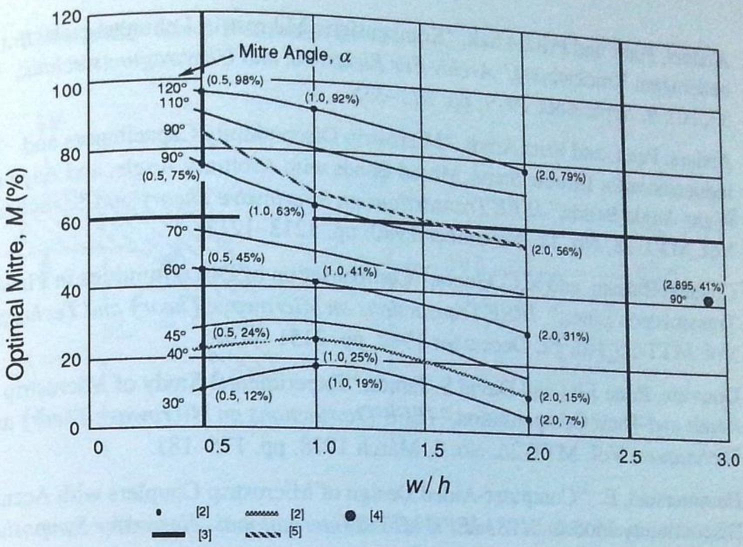

Figure 5.5.3.2: Optimum Mitres for Arbitrary Bend Angles

$$M = \frac{100\,x}{d} \quad (\%) \tag{5.5.3.1}$$

$$d - x = \sqrt{2}\,w\left(1.0 - \frac{M}{100}\right) \tag{5.5.3.2}$$

The arbitrary angle bend can be compensated by taking material off of the outside of the corner. This can be done by mitering. Figure 5.5.3.2 shows the optimal mitre percentage to achieve a matched impedance as a function of the line's w/h ratio. The accuracy is expected to be within a few percent [3]. For comparison the equation of [5] is plotted for the optimal 90° mitred bend. The other references are also shown.

The electrical length of the bend is calculated:

$$\frac{l_k}{w} = \frac{2\,M}{100\,\sin\alpha} \tag{5.5.3.3}$$

REFERENCES

[1] Anders, Peter, and Fritz Arndt, "Beliebig abgeknickte Mikrostrip-Leitungen mit bogenförmigen Übergang," *Archiv Für Elektronik und Übertragungstechnik*, Band 33, Heft 3, March 1979, pp. 93–99.

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5.5.4 Microstrip Line Rounded Bend

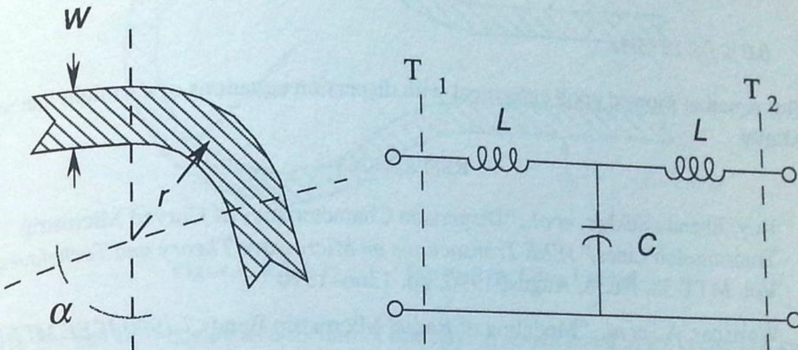


Figure 5.5.4.1: Microstrip Line Rounded Bend

Curves of model parameters are available in Weisshar *et al.* [2]; however, no closed-form equations have been derived to date. The dispersion relative to a straight line has been derived by curve-fitting experimental results in Roy *et al.* [1] and are:

$$\varepsilon_{eff,bend}(f) = \varepsilon_r - \frac{\varepsilon_r - \varepsilon_{eff}(0)\exp\left(\dfrac{\sqrt[4]{h/r}}{12.0}\right)}{1.0 + \left(\dfrac{f}{f_c}\right)^2} \tag{5.5.4.1}$$

where

$$f_c = \frac{c}{4.0\,h\sqrt{\varepsilon_r - 1.0}} \tag{5.5.4.2}$$

$\varepsilon_{eff}(0)$ is defined in (3.5.1.3)

r is the mean radius of the curve

which is valid for

$2.0 \le \varepsilon_r \le 12.0$

$0.8 \le w/h \le 5.0$

$\dfrac{\lambda_g}{4.0\,\pi\,h} < r/h \le \infty$

$0.0 \le f \le 12$ GHz

This equation showed good agreement with dispersion equations of a straight line when r is infinity.

REFERENCES

[1] Roy, Jibendu Sekhar, *et al.*, "Dispersion Characteristics of Curved Microstrip Transmission Lines," *IEEE Transactions on Microwave Theory and Techniques*, Vol. MTT-38, No. 8, August 1990, pp. 1366–1370.

[2] Weisshar, A., *et al.*, "Modeling of Radial Microstrip Bends," *1990 IEEE MTT-S Symposium Digest*, pp. 1051–1054.

5.5.5 Optimal Microstrip Line Rounded Bend

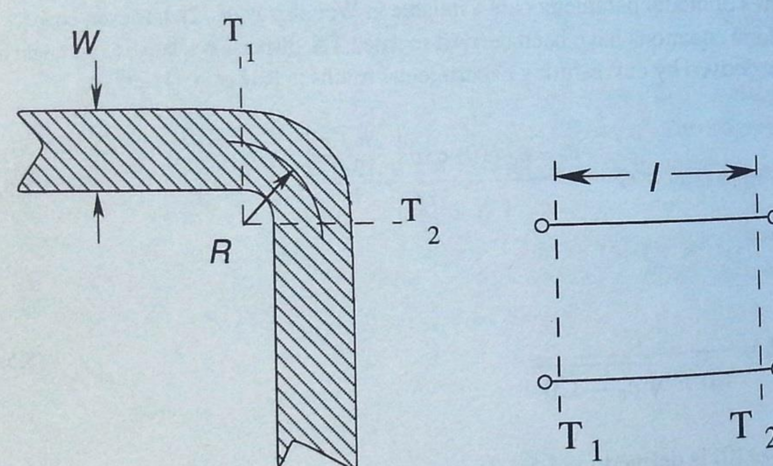


Figure 5.5.5.1: Optimal Microstrip Line Rounded Bend

As an alternative to the mitred corners of the previous sections, a microstrip line may be gradually bent to make a corner. In [2] it is stated that $R \ge 4.0\,h$ will give VSWR < 1.05. Generally, this is not a space efficient technique to use. U-bends can be made in a similar fashion as shown in Figure 5.5.5.2 below.

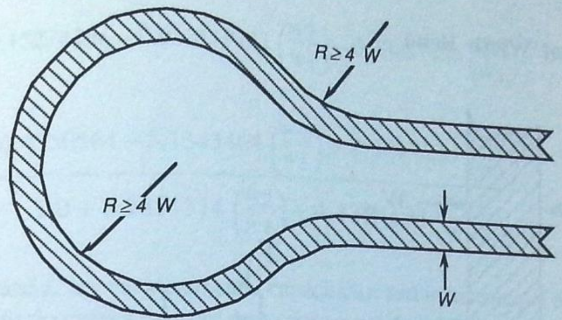


Figure 5.5.5.2: Microstrip Line U-bend

The line can be modeled as a section of line having length equal to the center line length of the curved section. This model does not include the changes in ε_{eff} and the effects of coupling between parts of the curve.

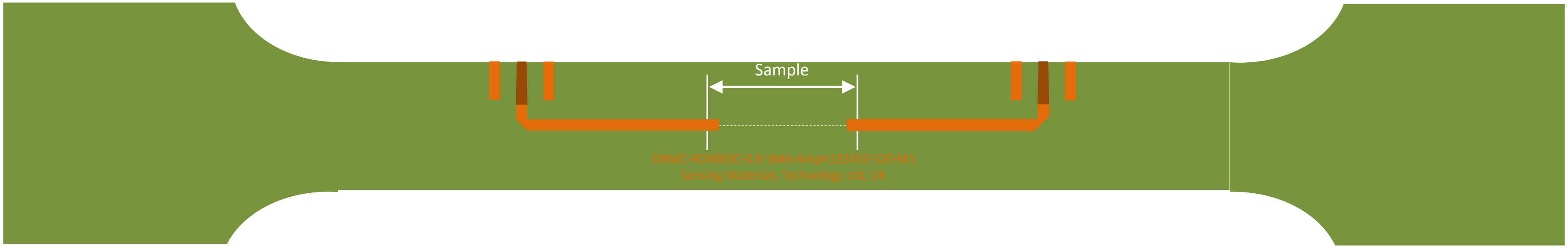
REFERENCES

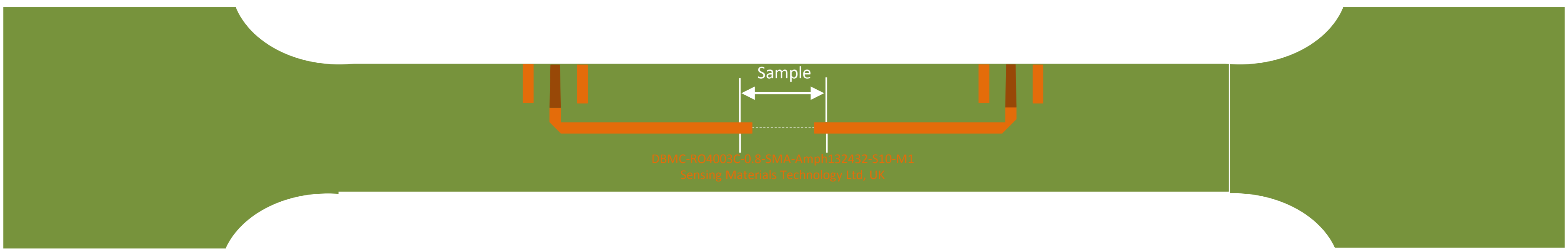
[1] Anders, Peter, and Fritz Arndt, "Beliebig abgeknickte Mikrostrip-Leitungen mit bogenförmigen Übergang," *AEÜ*, Band 33, Heft 3, March 1979, pp. 93–99.

[2] Niehenke, Edward C., "Additional Notes Microwave Transmission Media Microstrip Discontinuities."

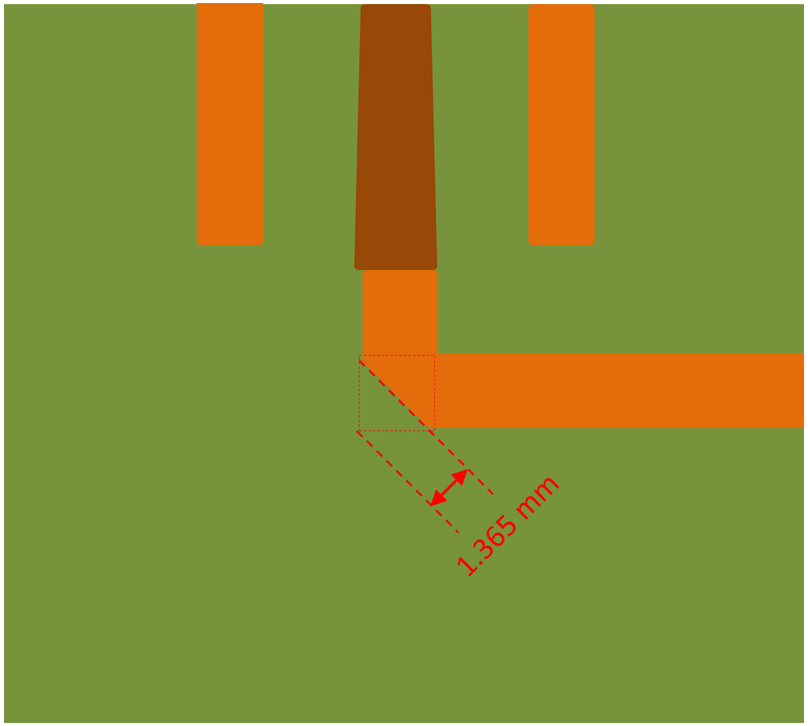
[3] Roy, Jibendu Sekhar, *et al*., "Dispersion Characteristics of Curved Microstrip Transmission Lines," *IEEE Transactions on Microwave Theory and Techniques*, Vol. MTT-38, No. 8, August 1990, pp. 1366–1370.

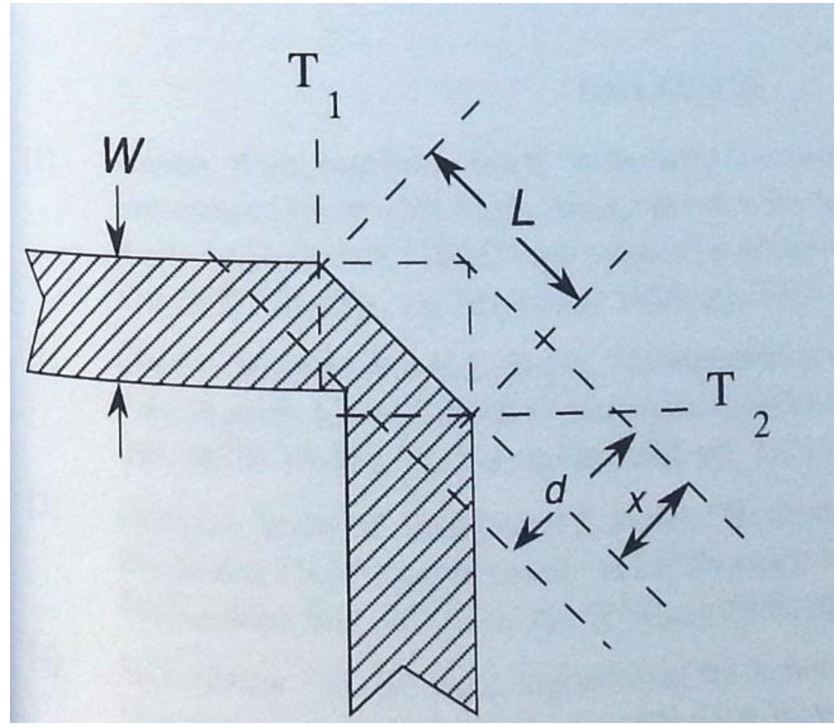
[4] Weisshaar, A., *et al*., "Modeling of Radial Microstrip Bends," *1990 IEEE MTT-S International Microwave Symposium Digest*, pp. 1051–1054.


Sample
DBMC-RO4003C-0.8-SMA-Amph132432-S20-M1
Sensing Materials Technology Ltd, UK


Sample
DBMC-RO4003C-0.8-SMA-Amph132432-S10-M1
Sensing Materials Technology Ltd, UK


Sample
DBMC-RO4003C-0.8-SMA-Amph132432-S5-M1
Sensing Materials Technology Ltd, UK

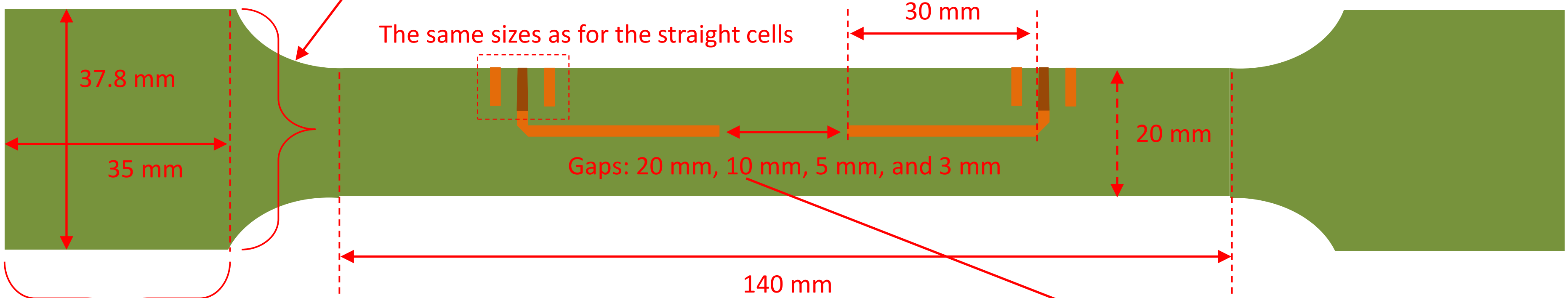




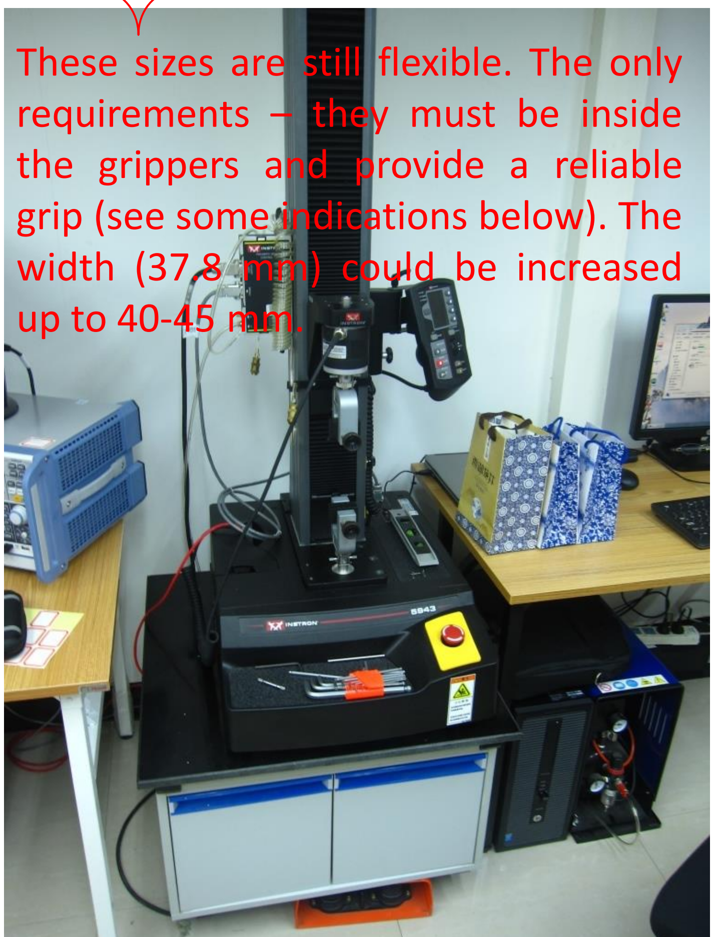

```python
# Program in Python
import math
w=1.74159 # strip width in mm
h=0.8 # PCB thickness in mm
d=2**0.5*w # diagonal in mm
x=d*(0.52+0.65*math.exp(-1.35*w/h)) # cut off from the corner in mm
print('x=',x,'mm')
```

x= 1.365474753822523 mm ← This value will be used for our cells

Propose a smooth transition

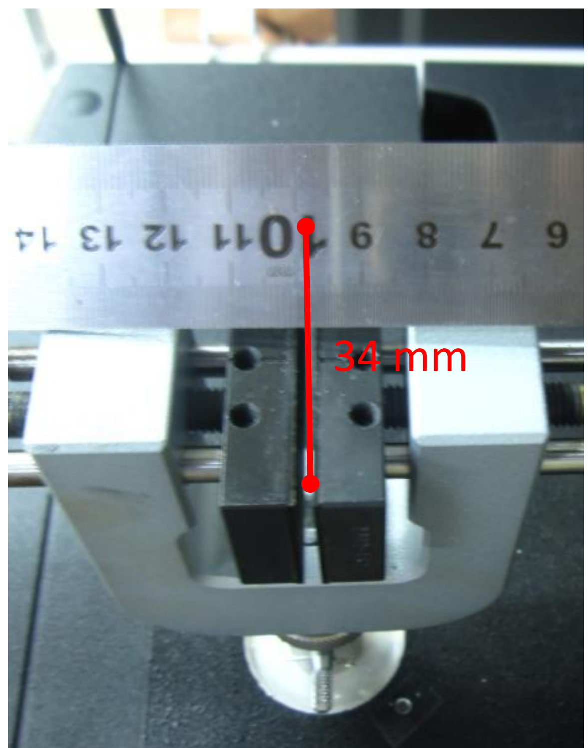


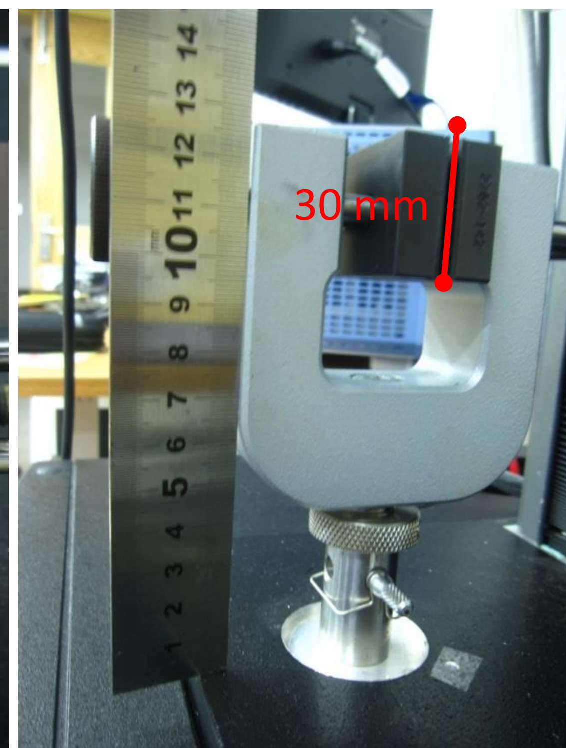
These sizes are still flexible. The only requirements – they must be inside the grippers and provide a reliable grip (see some indications below). The width (37.8 mm) could be increased up to 40-45 mm.

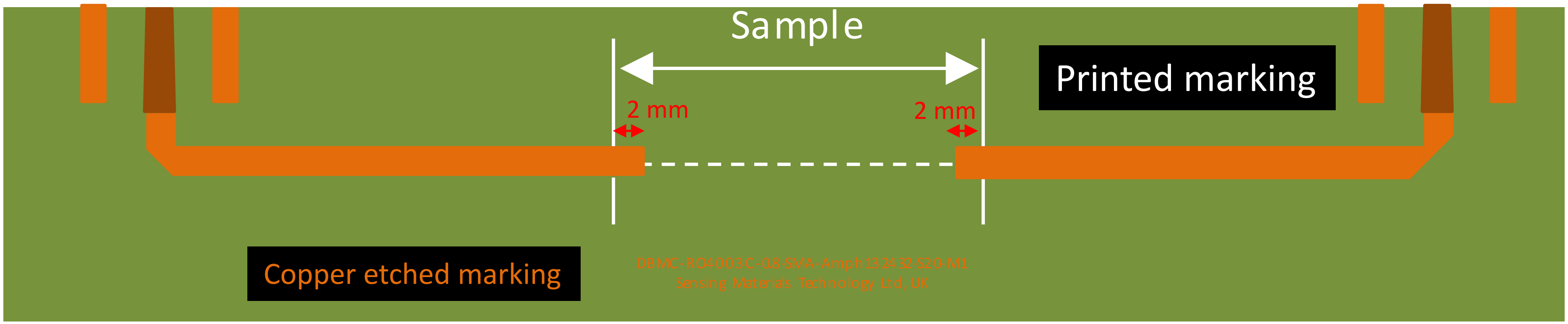


DBMC-RO4003C-0.8-SMA-Amph132432-S10-M1
Sensing Materials Technology Ltd, UK

Copper etched marking

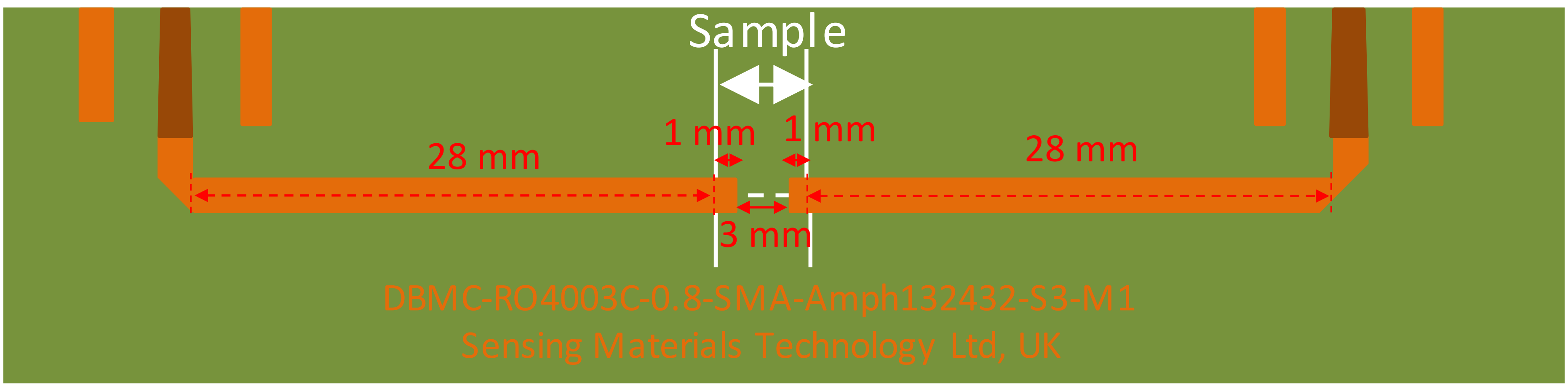


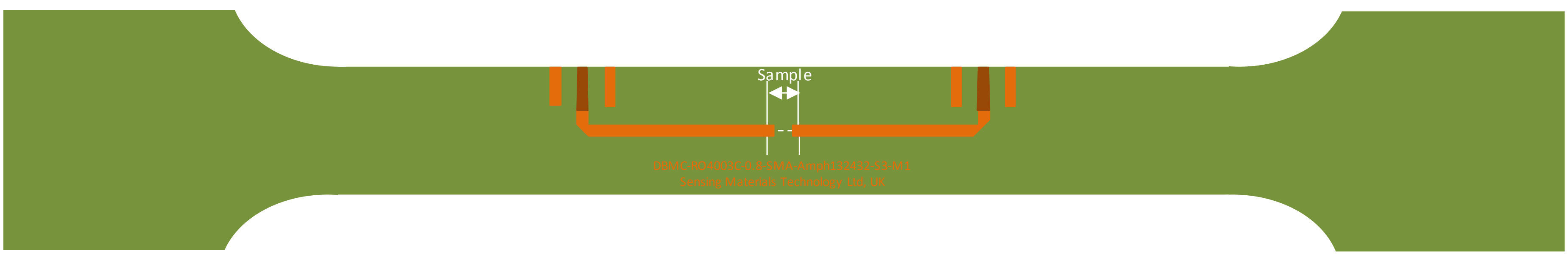



Sample
2 mm
2 mm
Printed marking
Copper etched marking
DBMC-RO4003C-0.8-SMA-Amph132432-S20-M1
Sensing Materials Technology Ltd, UK

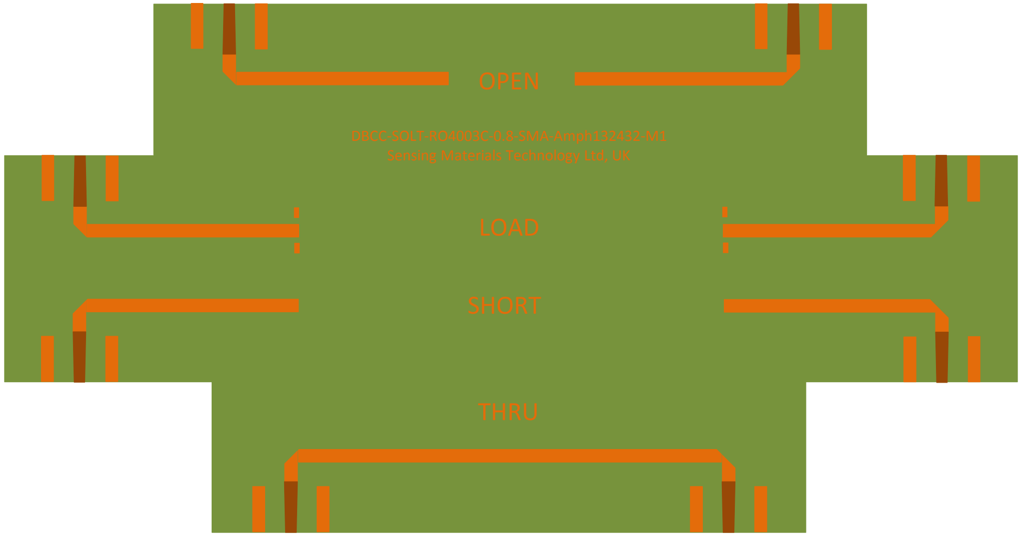
A very “challenging cell” – extremely short sample (3 mm) and soldering pads (1 mm)

After a proper calibration, a VNA will return correctly measured S-parameters of a sample soldered between the contact pads. These S-parameters can be formally recalculated into the impedance Z. However, such recalculation assumes that Z is a lumped parameter. The latter is fully correct only if the wavelength along the sample is much larger than its longitudinal size. Therefore, for the accurate results, especially when measuring the frequency dispersion of Z, the sample length must be significantly reduced. Here, we face a typical problem in microwave measurements – interpretation of results. Note that S-parameters are always correct (after a proper calibration), but their recalculation into something else may require more sophisticated models. In our project, we have to find new approaches to this problem if the sample size is not meant to be extremely small. The measurement cell proposed below still allows traditional recalculation procedure from S21/12-parameter into the lumped Z up to several GHz. In a VNA, such recalculation can be done in real time.





SOLT calibration board



OPEN

DBCC-SOLT-RO4003C-0.8-SMA-Amph132432-M1
Sensing Materials Technology Ltd, UK

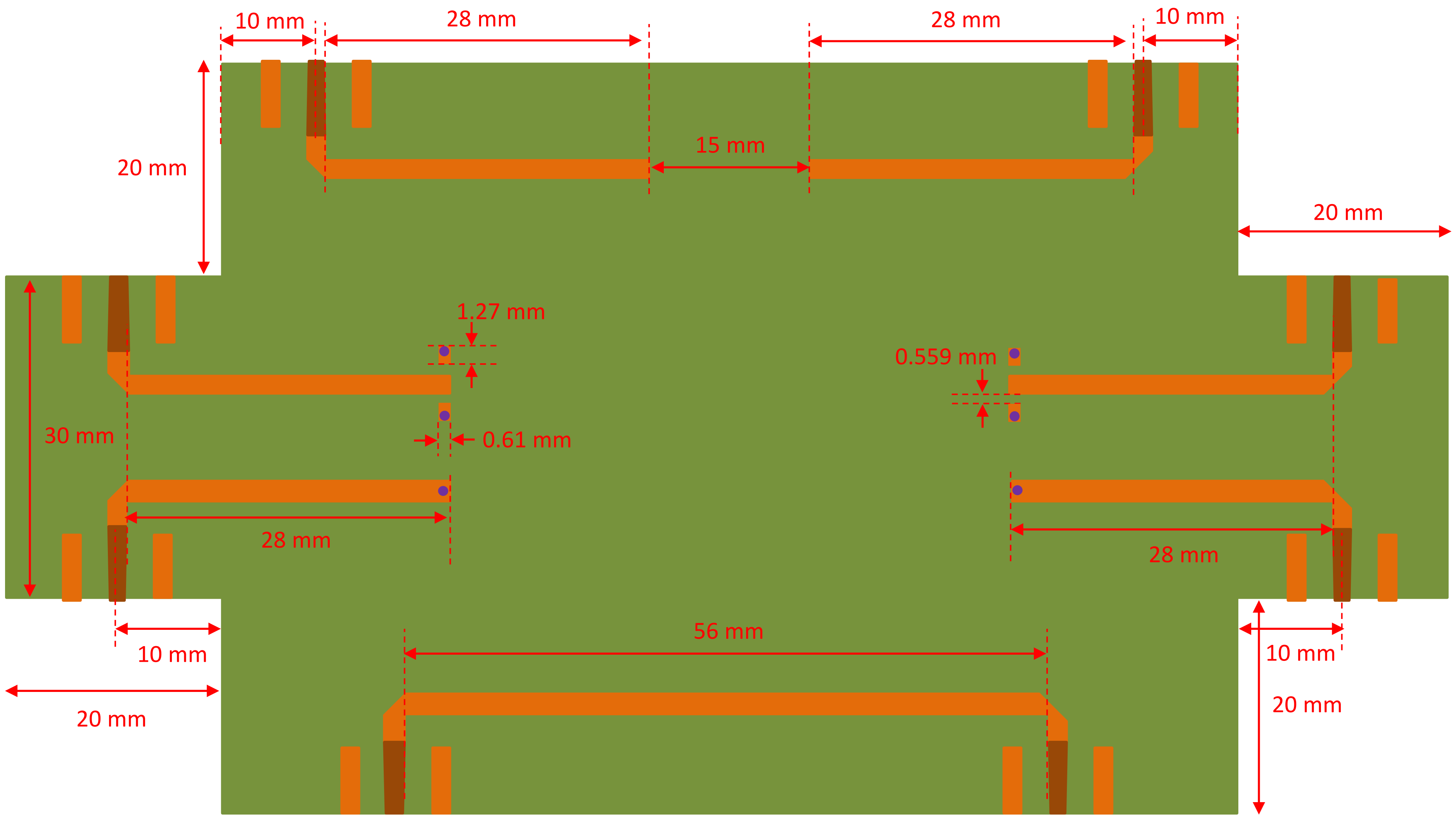
LOAD

SHORT

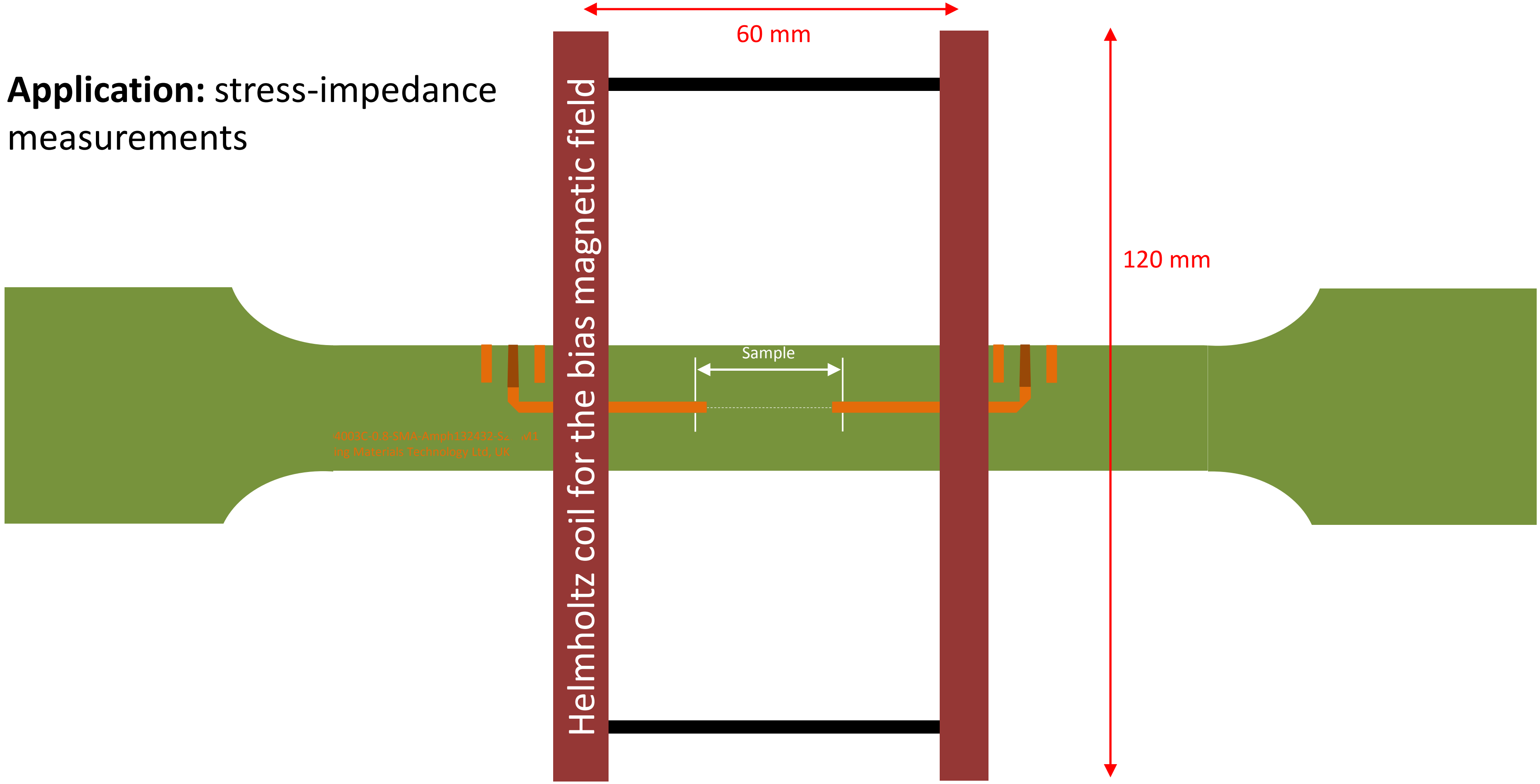
THRU

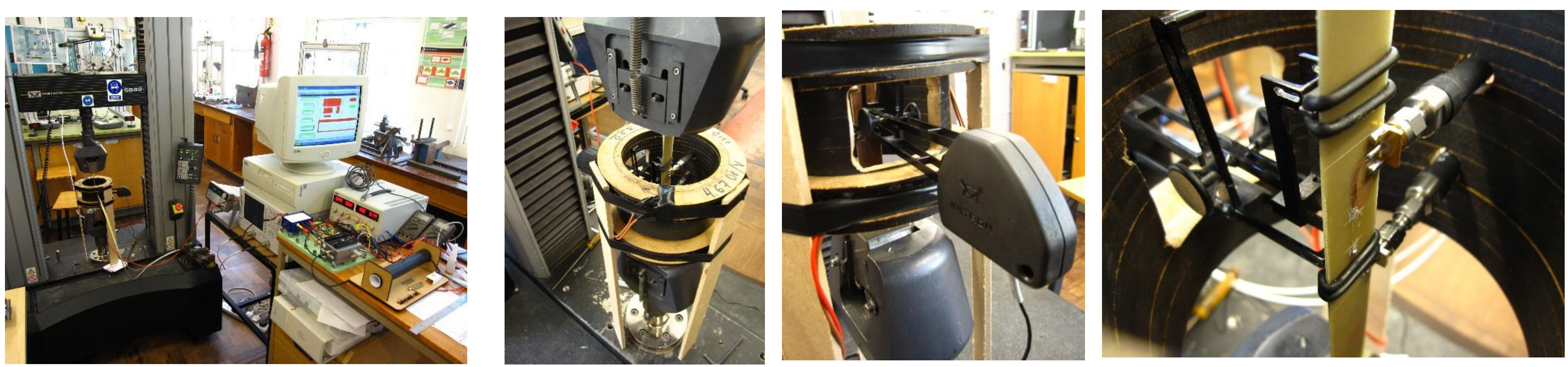
Copper etched marking

SOLT calibration board



Application: stress-impedance measurements





TRL calibration boards

When calibrating with TRL, absolutely the same measurement cells (straight or dog bone) will be used, but the calibration boards are now completely different.

An extraction form Rohde&Schwarz ZNB VNA user manual:

A TRL calibration requires the two-port standards THRU and LINE, which are both assumed to be ideally matched. Beyond that, THRU must be lossless, and its length must be exactly known. The length of LINE standard must be known approximately. Furthermore, a reflecting one-port standard (Reflect) is needed. The magnitude of the reflection coefficient of the Reflect standard can be unknown but must be nonzero; its phase must be roughly known (90 deg). The magnitude and phase of the reflection coefficient must be the same at both test ports. TRL calibration is especially useful for DUTs in planar line technology (e.g. test fixtures, on-wafer measurements) where it is difficult to design and connect accurately modelled OPEN, SHORT or MATCH standards.

The system of equations solved to derive the error terms is such that singularities occur whenever the length difference ΔL between THRU and LINE is an integer multiple of half of the wave length. As a rule, singularities are avoided with sufficient accuracy if the phase shift resulting from the electrical length difference between THRU and LINE standard is between 20° and 160°. This corresponds to a ratio of 1:8 for the start and stop frequency of the calibrated sweep range.

Calculations for LINE

For a phase difference, we have:

$$\varphi = 2\pi f \Delta t \qquad (1)$$

where f is a frequency and Δt is a time interval. In our case, the time interval means a time delay caused by the wave propagation along LINE as compared to THRU:

$$\Delta t = \frac{l}{v} = \frac{l\sqrt{\varepsilon_{eff}}}{c} \qquad (2)$$

where l is the length of “LINE – THRU” (extension), $v = c/\sqrt{\varepsilon_{eff}}$ is the wave propagation speed along LINE, c is the speed of light in vacuum, and ε_{eff} is the effective dielectric constant as if the wave propagated in free space. Some online impedance calculators provides ε_{eff} for a selected structure, for example, this one: https://www.pasternack.com/t-calculator-microstrip.aspx. Putting (2) into (1), we obtain:

$$\varphi = \frac{2\pi f \sqrt{\varepsilon_{eff}}\, l}{c} \qquad (3)$$

$$f = \frac{\varphi c}{2\pi\sqrt{\varepsilon_{eff}}\, l} \qquad (4)$$

$$l = \frac{\varphi c}{2\pi\sqrt{\varepsilon_{eff}}\, f} \qquad (5)$$

For $\varphi = \pi/9$ (20^0) and $\varphi = 8\pi/9$ (160^0), we obtain:

$$f_{low} = \frac{c}{18\sqrt{\varepsilon_{eff}}\, l} \qquad (6)$$

$$f_{high} = \frac{4c}{9\sqrt{\varepsilon_{eff}}\, l} \qquad (7)$$

$$\frac{f_{high}}{f_{low}} = 8 \qquad (8)$$

As the design rules for LINE, we will use the following equations:

$f \in [f_{low}, 8 f_{low}]$ - selected frequency range (9)

$$l[mm] = \frac{c[mm/s]}{18\sqrt{\varepsilon_{eff}}\, f_{low}[Hz]} \approx \frac{10^{11}}{6\sqrt{\varepsilon_{eff}}\, f_{low}[Hz]} \qquad (10)$$

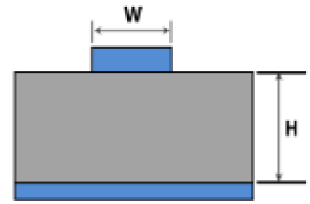
Pasternack's calculator (https://www.pasternack.com/t-calculator-microstrip.aspx)

Dielectric Constant ▼ | 3.804

Width: 1.74159 | Millimeters ▼

Height: 0.812 | Millimeters ▼

Calculate



Result:

Width/Height:	**2.145**
Effective Dielectric Constant:	**2.948**
Impedance:	**50.02 Ω**

As the design rules for LINE, we will use the following equations:

$f \in [f_{low}, 8 f_{low}]$ - selected frequency range

$$l[mm] = \frac{c[mm/s]}{18\sqrt{\varepsilon_{eff}}\, f_{low}[Hz]} \approx \frac{10^{11}}{6\sqrt{\varepsilon_{eff}}\, f_{low}[Hz]}$$

2.948

$$If\ \left(\frac{W}{H}\right) < 1:$$

$$\varepsilon_{eff} = \frac{\varepsilon_R+1}{2} + \frac{\varepsilon_R-1}{2}\left[\frac{1}{\sqrt{1+12\left(\frac{H}{W}\right)}} + 0.04\left(1-\left(\frac{W}{H}\right)\right)^2\right]$$

$$Z_0 = \frac{60}{\sqrt{\varepsilon_{eff}}}\ln\left(8\left(\frac{H}{W}\right) + 0.25\left(\frac{W}{H}\right)\right)$$

$$If\ \left(\frac{W}{H}\right) > 1:$$

$$\varepsilon_{eff} = \frac{\varepsilon_R+1}{2} + \left[\frac{\varepsilon_R-1}{2\sqrt{1+12\left(\frac{H}{W}\right)}}\right]; \quad Z_0 = \frac{120\pi}{\sqrt{\varepsilon_{eff}}\left[\frac{W}{H}+1.393+\frac{2}{3}\ln\left(\frac{W}{H}+1.444\right)\right]}$$

For three overlapping frequency ranges, we obtain:

[250 MHz, 2 GHz] $\Rightarrow$ LINE1 = THRU+38.83 mm

[500 MHz, 4 GHz] $\Rightarrow$ LINE2 = THRU+19.41 mm

[3 GHz, 18 GHz] $\Rightarrow$ LINE3 = THRU+3.24 mm

$$l[mm] \approx \frac{10^{11}}{6\sqrt{2.948}\, f_{low}[Hz]}$$

These calculations can be checked using the online calculator: http://www.emtalk.com/mscalc.php

RO4003C

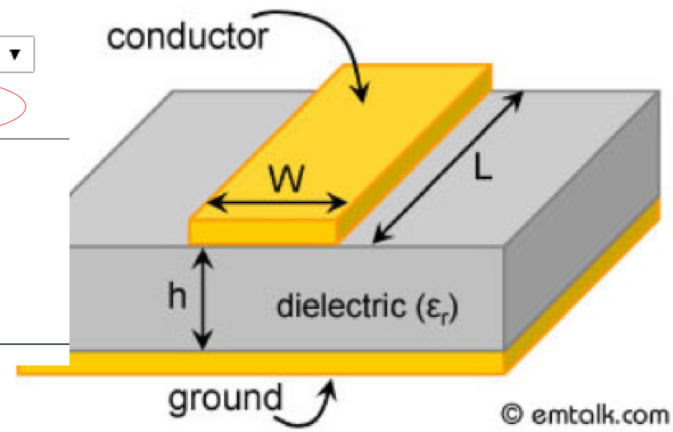
Dielectric Constant (ε_r): 3.804

Dielectric Height (h): 0.812 mm

Frequency: 2 GHz

Electrical Parameters			Physical Parameters		
Zo:	50.0144159238	Ω	Width (W):	1.74159	mm
Elec. Length:	160.006600076	deg	Length (L):	38.83	mm

Analyze



Push "Analyze"

[250 MHz, 2 GHz]

$$\frac{10^{11}}{6\times\sqrt{2.948}\times 2.5\times 10^{8}} = 38.83$$

Substrate Parameters

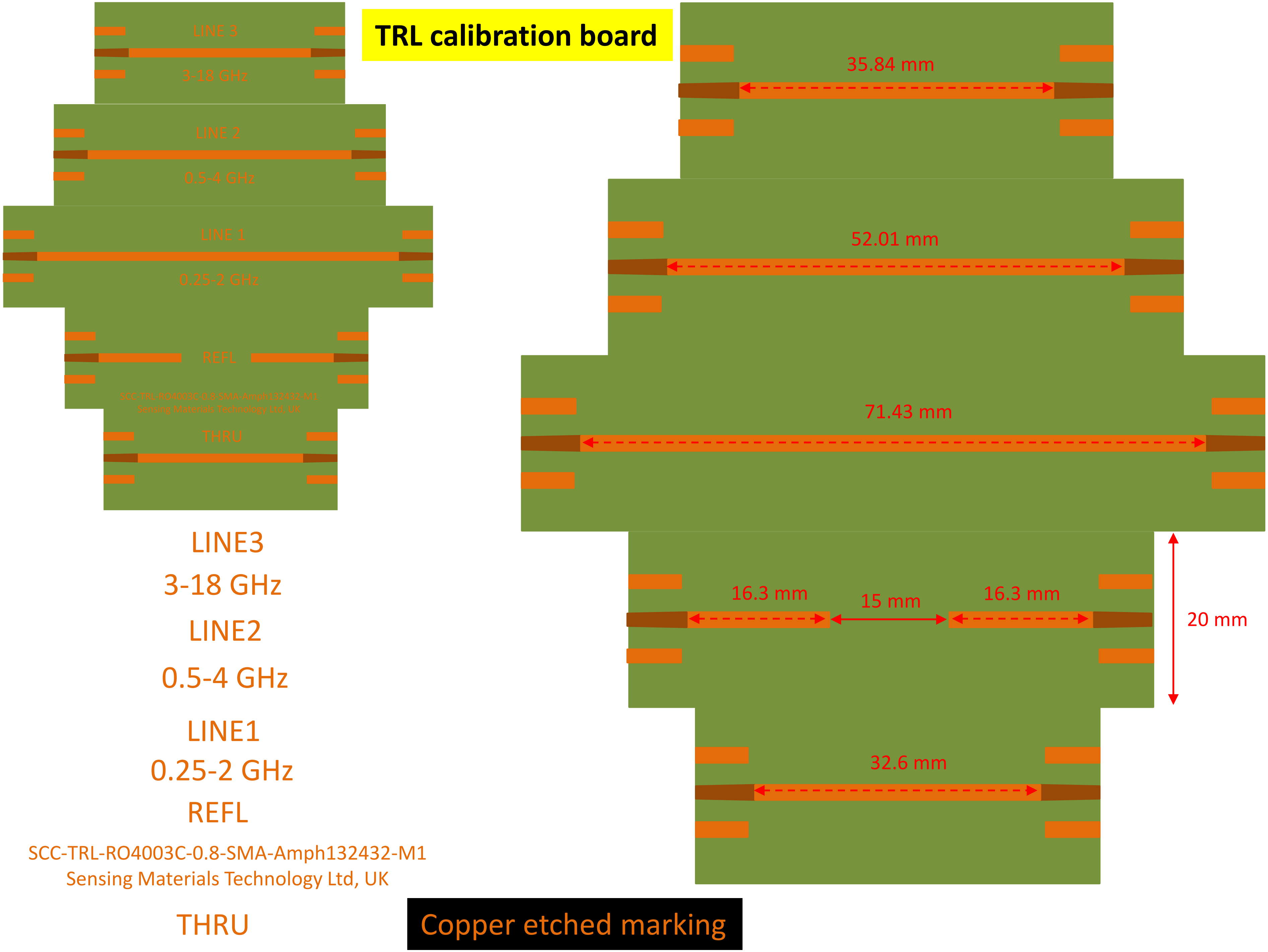
Dielectric Constant (ε_r): 3.804

Dielectric Height (h): 0.812 mm

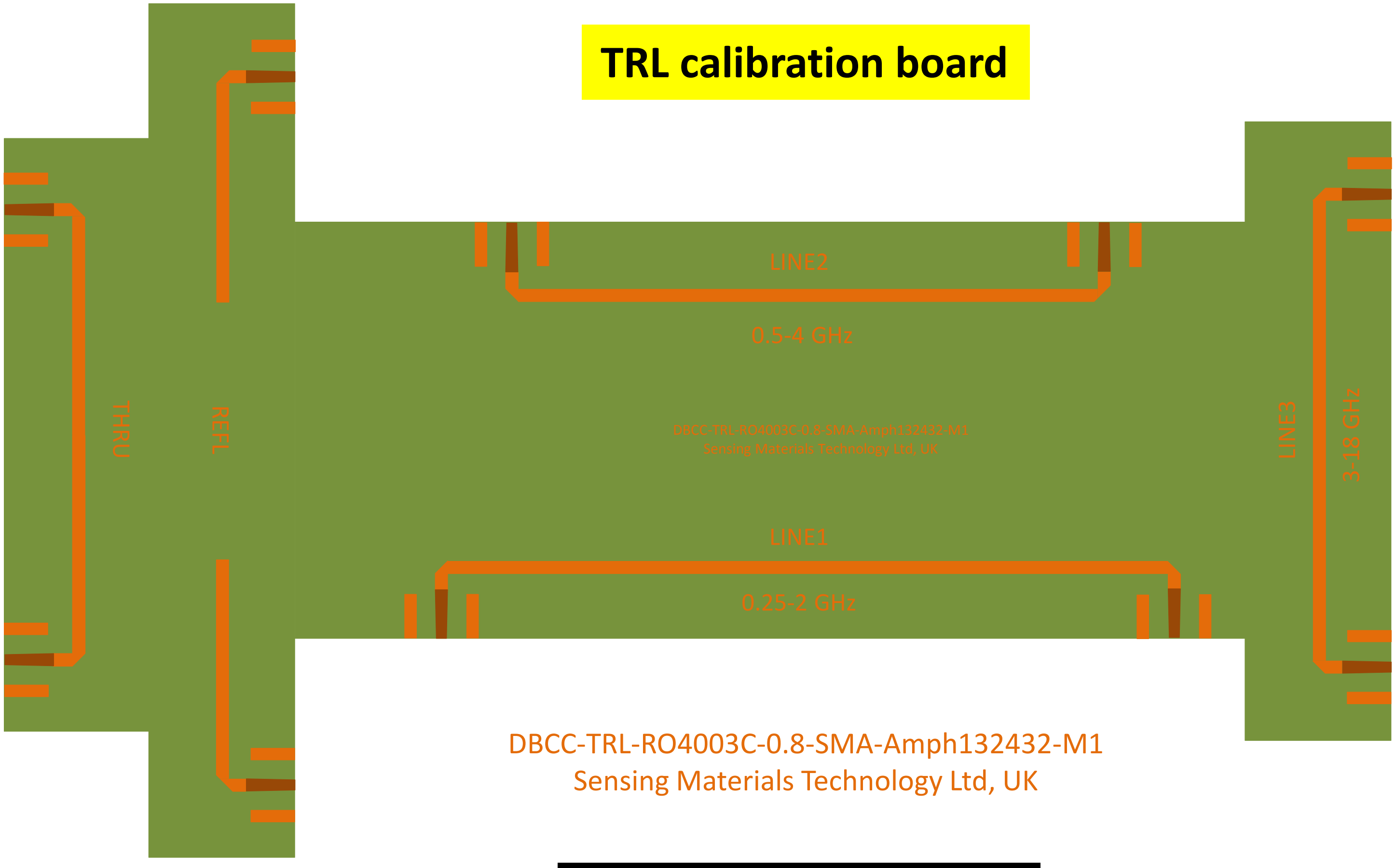
Frequency: 0.25 GHz

Electrical Parameters			Physical Parameters		
Zo:	50.0144159238	Ω	Width (W):	1.74159	mm
Elec. Length:	20.0008250095	deg	Length (L):	38.83	mm

Analyze

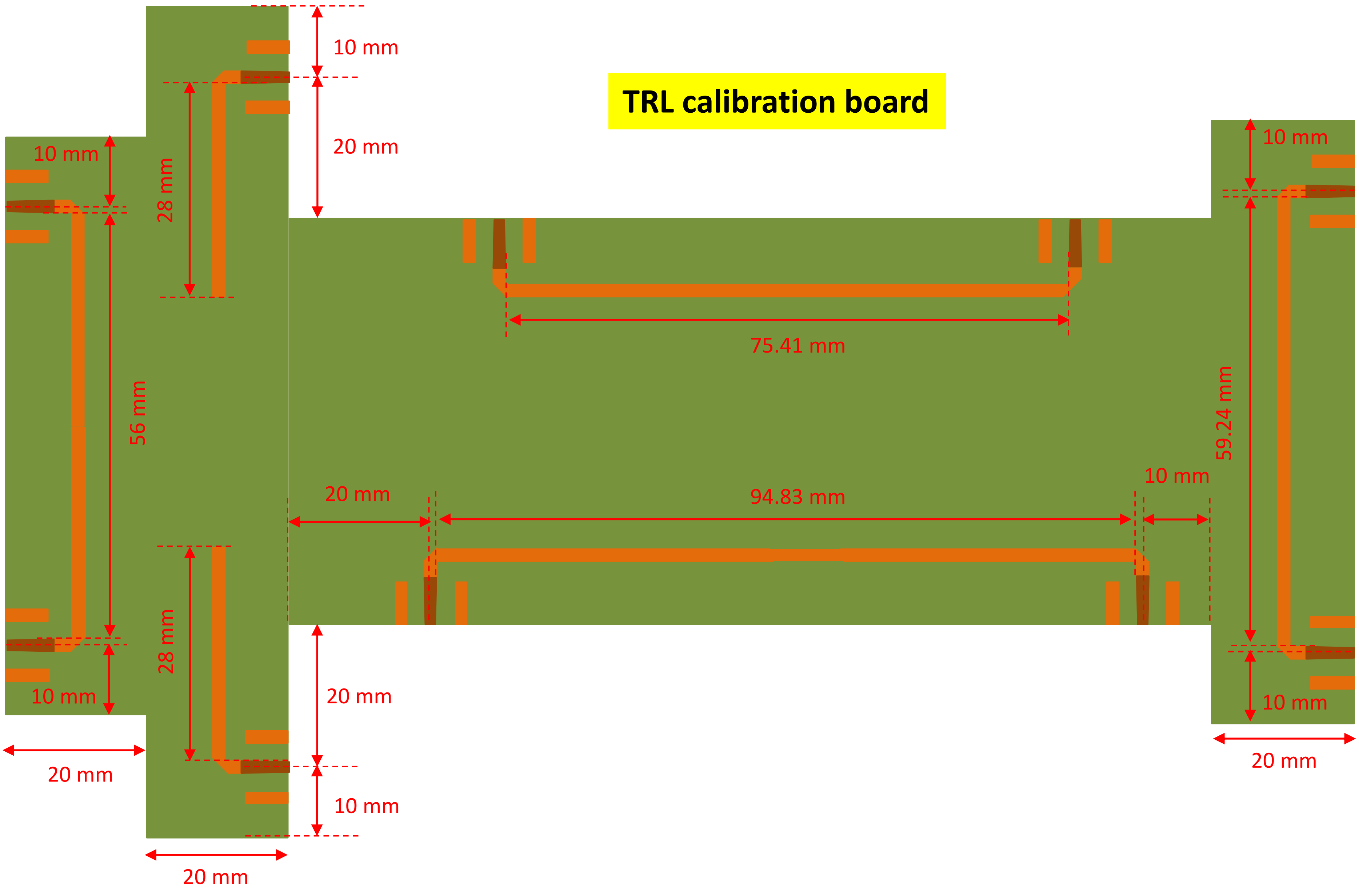

TRL calibration board
LINE 3
3-18 GHz
LINE 2
0.5-4 GHz
LINE 1
0.25-2 GHz
REFL
THRU
35.84 mm
52.01 mm
71.43 mm
16.3 mm
15 mm
16.3 mm
20 mm
32.6 mm
LINE3
3-18 GHz
LINE2
0.5-4 GHz
LINE1
0.25-2 GHz
REFL
SCC-TRL-RO4003C-0.8-SMA-Amph132432-M1
Sensing Materials Technology Ltd, UK
THRU
Copper etched marking

TRL calibration board



DBCC-TRL-RO4003C-0.8-SMA-Amph132432-M1
Sensing Materials Technology Ltd, UK

Copper etched marking



TRL calibration board
10 mm
20 mm
10 mm
28 mm
75.41 mm
56 mm
59.24 mm
10 mm
20 mm
94.83 mm
10 mm
28 mm
20 mm
10 mm
10 mm
20 mm
20 mm
10 mm
20 mm

Cells' taggings

SMC-RO4003C-0.8-SMA-Amph132432-S10-M1: Straight Measurement Cell, RO4003C material, PCB thickness 0.8 mm, SMA connector Amphenol 132432, Sample length 10 mm, Model 1

SCC-SOLT-RO4003C-0.8-SMA-Amph132432-M1: Straight Calibration Cell, SOLT calibration method, RO4003C material, PCB thickness 0.8 mm, SMA connector Amphenol 132432, Model 1

SCC-TRL-RO4003C-0.8-SMA-Amph132432-M1: Straight Calibration Cell, TRL calibration method, RO4003C material, PCB thickness 0.8 mm, SMA connector Amphenol 132432, Model 1

DBMC-RO4003C-0.8-SMA-Amph132432-S10-M1: Dog Bone Measurement Cell, RO4003C material, PCB thickness 0.8 mm, SMA connector Amphenol 132432, Sample length 10 mm, Model 1

DBCC-SOLT-RO4003C-0.8-SMA-Amph132432-M1: Dog Bone Calibration Cell, SOLT calibration method, RO4003C material, PCB thickness 0.8 mm, SMA connector Amphenol 132432, Model 1

DBCC-TRL-RO4003C-0.8-SMA-Amph132432-M1: Dog Bone Calibration Cell, TRL calibration method, RO4003C material, PCB thickness 0.8 mm, SMA connector Amphenol 132432, Model 1

Paper models of the cells in their natural size