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A SIMPLE APPROACH TO SIGNAL VIA STUBS FOR COAXIAL PCB CONNECTOR LAUNCHES

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Introduction

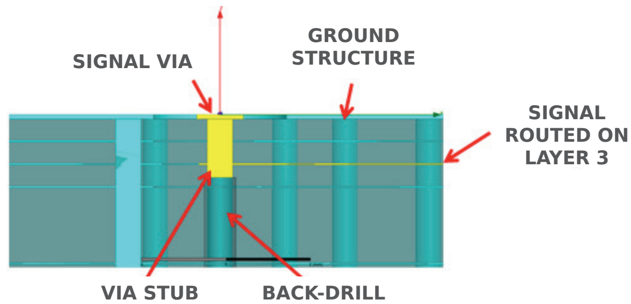
In my work as a Signal Integrity/RF engineer, I have many customers who request assistance in designing their printed circuit board (PCB) launches for coaxial test connectors. (For example, see Molex part number 73387-0020.) Customers typically use these connectors on PCBs designed to test another product such as a backplane or I/O connector. They want to maximize the bandwidth of the test connector and PCB launch in order to get the clearest picture of the intended device under test (DUT). With that goal in mind, I ask a series of questions so that I understand what the customer is attempting to do. Those questions are:

1. What is the PCB material?
2. What is the PCB stack?
3. What is the transmission line structure?
4. On what layer or layers will signals be routed?
5. For signals on internal layers, will signal vias be back-drilled or blind?
6. If the signal vias will be back-drilled, what is the longest possible stub length?
7. What is the desired performance (typically return loss or VSWR) for the test connector and launch?

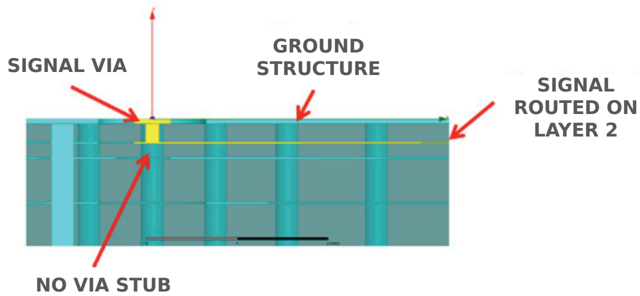
MOLEX 73387-0020

In designing PCBs, I will generally see three types of vias: through-hole, back-drilled and blind. Examples are shown below. The purposes of this paper are to present a relatively simple approach to modeling back-drilled vias and to suggest a rough rule of thumb *for those who do not use or have access to electrical modeling tools.* Therefore, I shall address the final three questions from the above list.

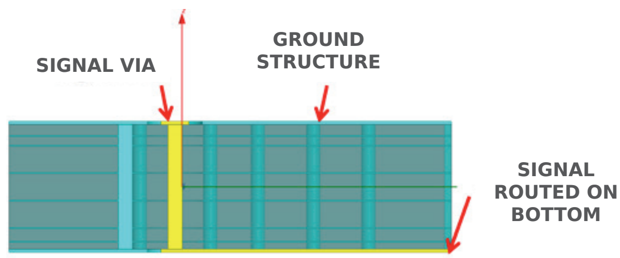
With respect to question 7, I shall define a "good" launch as a connector/launch that has a return loss of 20dB or better at the maximum frequency of interest. Additionally, I shall define the bandwidth of a launch as the maximum frequency at which the return loss crosses 20dB.



BACK-DRILLED VIA WITH SIGNAL ROUTED ON LAYER 3



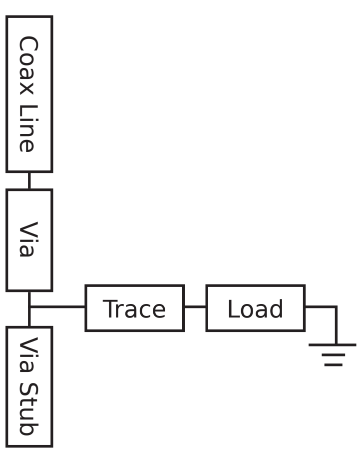
BLIND VIA WITH SIGNAL ROUTED ON LAYER 2



PLATED THROUGH-HOLE VIA
(ALL NON-FUNCTIONAL PADS REMOVED)

Model

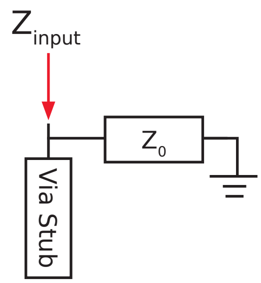
I shall start with a basic transmission line diagram of our connector and launch. The connector is described as a coaxial line.



In order to simplify the model so that we may deal specifically with the via stub, we assume that the trace and load become a perfect load and that the coax line and via are ideal transmission lines. In other words, the all elements are lossless and they have the same characteristic impedance.

$$Z_{Coax}=Z_{Via}=Z_{Trace}=Z_{Stub}=Z_{Load}=Z_0$$

It is acknowledged that these assumptions are not terribly realistic. For example, the trace has no reference plane as it exits the via and we are ignoring any stray capacitance from pads. However, the assumptions do allow a clear look at the impact of the via, or open circuit, stub. We can now look at the input impedance of the parallel combination of the stub and the perfect load.



Now, we examine the open circuit stub alone. The input impedance of the stub is:[1]

$$Z_{in}(l) = Z_0 \frac{Z_L + jZ_0 \tan(\beta l)}{Z_0 + jZ_L \tan(\beta l)}$$

where

l = length of the stub

$$Z_L = \infty$$

$$\beta = \frac{2\pi}{\lambda}$$

This expression simplifies to:

$$Z_{in}(l) = -jZ_0 \cot\left(\frac{2\pi l}{\lambda}\right)$$

In the argument of the cotangent, it is helpful to substitute

$$\frac{1}{\lambda} = \frac{f\sqrt{\varepsilon_r}}{c}$$

Which gives us the desired form of the open circuit stub's input impedance.

$$Z_{in}(l) = -jZ_0 \cot\left(\frac{2\pi l f\sqrt{\varepsilon_r}}{c}\right)$$

This input impedance is in parallel with an ideal load. The input impedance of the parallel combination is:

$$Zin(l) = \frac{-Z_0 * jZ_0 \cot\left(\frac{2\pi l f\sqrt{\varepsilon_r}}{c}\right)}{Z_0 - jZ_0 \cot\left(\frac{2\pi l f\sqrt{\varepsilon_r}}{c}\right)}$$

Which simplifies to:

$$Zin(l) = \frac{jZ_0 \cot\left(\frac{2\pi l f\sqrt{\varepsilon_r}}{c}\right)}{1 - j \cot\left(\frac{2\pi l f\sqrt{\varepsilon_r}}{c}\right)}$$

When the denominator is rationalized (multiply the equation by $\frac{1+j\cot\left(\frac{2\pi l f\sqrt{\varepsilon_r}}{c}\right)}{1+j\cot\left(\frac{2\pi l f\sqrt{\varepsilon_r}}{c}\right)}$) *and simplified, we end up with an input impedance of*

$$Zin(l) = \frac{Z0}{2}\left\{1 + \cos\left(\frac{2\pi(2f)l\sqrt{\varepsilon_r}}{c}\right) - j\sin\left(\frac{2\pi(2f)l\sqrt{\varepsilon_r}}{c}\right)\right\}$$

This equation may be implemented easily in a spreadsheet.

Stub Length (in)	Relative Dielectric	c (m/s)	Z0 (ohms)
0.010	3.5	3.00E+08	50

Frequency (GHz)	Re(Z)	Im(Z)	Mag(Z)
0.1	50.0000	-0.0490	50.0000
1	49.9952	-0.4901	49.9976
5	49.8800	-2.4467	49.9400
10	49.5211	-4.8699	49.7600
15	48.9268	-7.2463	49.4605
20	48.1027	-9.5532	49.0422
25	47.0569	-11.7683	48.5061
30	45.7993	-13.8705	47.8536
35	44.3420	-15.8395	47.0861
40	42.6990	-17.6564	46.2055

I have run models of Molex SMA (73251-3480), 2.92mm (73252-0090) and 2.40mm (73387-0020) compression mount test connectors with various routings, PCB materials and stub lengths in ANSYS® HFSS™.

Rule of Thumb: Based on the compiled data, an estimate of the maximum bandwidth of the connector and launch, due to the open circuit stub, is the frequency at which the magnitude of the impedance (column Mag(Z)) crosses 48 ohms for a 50 ohm system. (72 ohms for a 75 ohm system)

In the table below, the Spreadsheet 48 ohms column gives the frequency in GHz at which the Mag(Z) column in the spreadsheet is approximately 48 ohms. The Model RL 20dB column gives the frequency in GHz at which the HFSS™ model return loss crossed 20dB. (See graphs at the end of this paper.) Each model consisted of a connector and a PCB with 5mm of stripline trace.

	Connector: SMA			Connector: 2.92mm			Connector: 2.40mm		
	PCB relative dielectric: 4.6			PCB relative dielectric: 3.5			PCB relative dielectric: 3.6		
Stub (mm)	Spread-sheet 48 ohms	Model RL 20dB	% Error	Spread-sheet 48 ohms	Model RL 20dB	% Error	Spread-sheet 48 ohms	Model RL 20dB	% Error
0.15	42	12.8	228.1	48	34.3	39.9	48	37.8	27.0
0.25	25	12.3	103.3	29	28	3.6	28	31.2	-10.3
0.35	18	12	50.0	21	22.4	-6.2	20	26.2	-23.7
0.45	14	11.4	22.8	16	16.8	-4.8	16	21.2	-24.5
0.55	11	10.9	0.9	13	12.5	4.0	13	17.4	-25.3

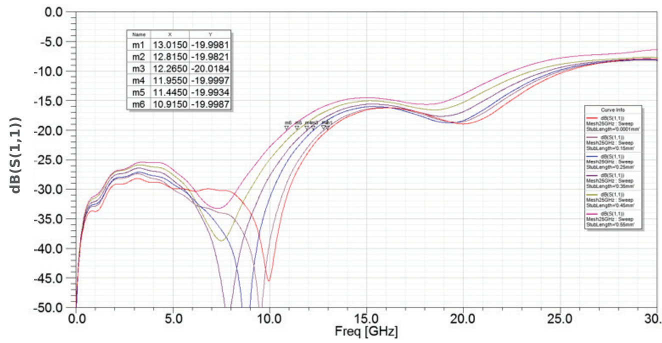
RED = MAG(PERCENT ERROR) IS GREATER THAN 30; BLUE = MAG(PERCENT ERROR) IS LESS THAN 30

All the bandwidth estimates (Spreadsheet 48 ohms) for the 2.4mm connector are within 30% of the actual frequency at which the model return loss was 20dB. The shortest stub length limit is out at 48GHz, which is nearly the rated bandwidth of the 2.40mm interface itself.

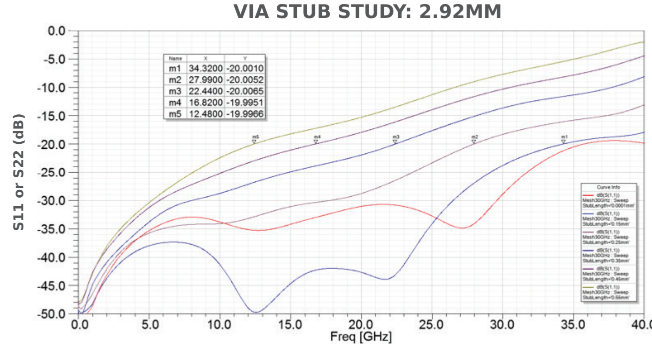
For the 2.92mm connector, the limit for the 0.15mm stub is also 48GHz, which is beyond the rated bandwidth of the 2.92mm interface. In other words, the open circuit stub is not the limiting factor in the model.

This particular SMA launch was a first attempt at a design with a blind via that would eventually work up to 20 GHz. For the sake of this investigation, via stubs were added to the model. Again, the rule of thumb is way off with the shorter stubs because there are other problems that limit the launch bandwidth more than the stub as these lengths

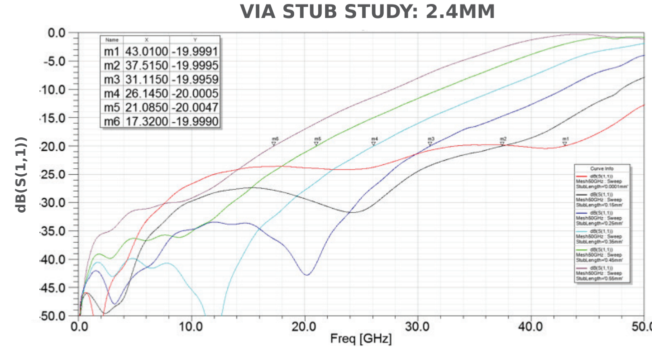
VIA STUB STUDY: SMA



VIA STUB STUDY: 2.92MM



VIA STUB STUDY: 2.4MM



Conclusion

1. The open circuit stub of a back-drilled via must not be overlooked when a system bandwidth should be maximized.
2. The “rule of thumb” stated above appears to be a reasonable estimate of the bandwidth limit imposed by an open-circuit stub. Again, this estimate is intended for those who do not use or have access to electrical modeling tools.

[1] *Hayt, William H., Jr., Engineering Electromagnetics, 5th ed., McGraw-Hill: New York, 1989, p386*

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