

Inductors at UHF: EM Simulation Guides Vector Network Analyzer Measurements

John B. Call
Thales Communications Inc., USA

There are times when it is necessary to measure a component in the lab even though commercial models are available

When designing circuits for operation at UHF frequencies such as filter networks, matching circuits, or dc bias for high frequency circuits, lumped element inductors can achieve a significant space savings when compared to distributed components. Designers using lumped element components at UHF frequencies have many types of models available from component manufacturers or third party vendors that specialize in modeling. These models can be used to build workable circuit simulations that can accelerate initial design efforts. There are times in the design process when it is desirable to measure a component to obtain a model outside the frequency range offered by vendors, to characterize a custom component, or to measure a component in a configuration that is difficult to model. For example, a large inductance value may be needed for a matching circuit and the vendor may recommend using the component to just below the first self resonance, yet the design needs to go beyond the first self resonance. Or a designer has developed a custom inductor and would like to obtain a model that can be useful in simulating the overall design.

There are times when it is desirable and necessary to measure a component in the lab across a broad frequency range even though

commercial models are available. Component testers are useful for verifying the data sheet value of a component, but they may not offer the needed insight into how the component looks in-circuit under mounting conditions that often differ from factory test fixtures. Typical testers are able to measure at a small handful of discrete frequencies and are not capable of providing a complete picture of the parasitics associated with the component. When measuring a custom inductor, the component may not mount properly in a standard test fixture. With most RF/microwave labs having a good vector network analyzer capable of performing a true TRL calibration it is a straightforward matter to generate a custom calibration kit and corresponding test cards for measuring components in a manner representative of the final design. A practical issue not highlighted in the literature is that inductor measurements at high frequencies are affected by parasitics associated with mounting on a planar circuit. Even with careful fixture construction and exceptional measurement techniques it may not be possible to separate the fixture parasitics from the component measurement.

A one-port measurement is preferred for a two-terminal passive component. Simulation results below will show less error for an ideal one-port measurement than for a two-port. Factor in the ability to obtain a good calibration of the vector network analyzer, the added complexity of the test fixture, and the ability

to resolve a small series reactance for a two-port measurement and the error sources for the two-port become more numerous than for a one-port measurement. Using TRL calibration with well constructed calibration standards simplifies the test fixture and calibration tasks while allowing control over the location of the measurement reference plane. Done properly, a one-port S -parameter data file can often be used directly in a simulation to represent the measured component. At other times it may be necessary to extract elements of a component model using the measured data set.

Measuring the Ideal Inductor

In Fig. 1 is a fixture for measuring a lumped element component. The circuit is a coplanar wave guide transmission line of width 50 mils with a gap between the line and ground of 25 mils. The geometry is chosen for ease of construction as no vias are required and mounting/use of the measured component will be similar in the final circuit. The reference plane is pulled back from the end of the line by 50 mils to allow for a component mounting pad. The substrate material is 60 mil FR-4 with a relative permittivity of $\epsilon_r = 4.3$ and a dielectric loss tangent of $\tan \delta = 0.025$. Coplanar wave guide is a convenient choice for this type of circuit as it can be conveniently prototyped without requiring via contact to a backside ground. A Sonnet simulation of this fixture and measurement of an ideal inductor and capacitor is constructed. Sonnet provides the ability to add two port components to the EM simulation. The conductor is loss less. Several simulation experiments will be run with an ideal component in locations Z1, Z2 and Z3 at frequencies from 200 – 1000

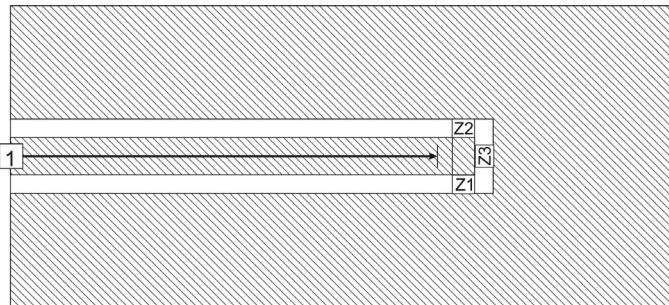


Figure 1 • Simulation circuit for one-port measurement of an inductor.

MHz. The model of the test fixture from a circuit perspective is shown in Fig. 2. Of practical necessity there is a portion of the test fixture that extends beyond the port 1 reference plane to which the device under test (DUT) is attached. The transmission line is approximated as a series of distributed circuit elements across an incremental length Δz [1, 2]. Depending on the frequency range of interest and the location of the reference plane, the transmission line extending from the reference plane to the DUT may be modeled as one or more sections of length

Δz . The geometry of this fixture has been selected such that, from 200 – 1000 MHz, the fixture effects can be modeled with discrete lumped elements for the section Δz . The fixture model will change according to whether the DUT is at location Z1 or Z3. For measurement of a single component the fixture model will be the same for location Z1 and Z2.

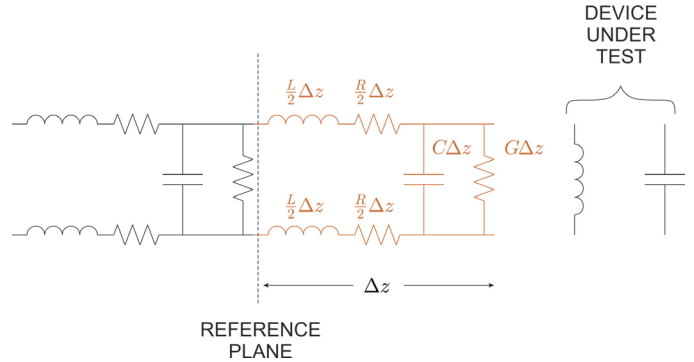


Figure 2 • Circuit element representation of test fixture showing transmission line structure, reference plane, and device under test.

The experiment will be to determine the effects of measuring an ideal inductor with a value of 3.4 nH and a capacitor with a value of 15.6 nH. The simulation is run with a component in a given location. The S -parameters from the simulation will be used to deembed a model of the ideal component. This will allow comparison of the expected measurement results with the actual circuit.

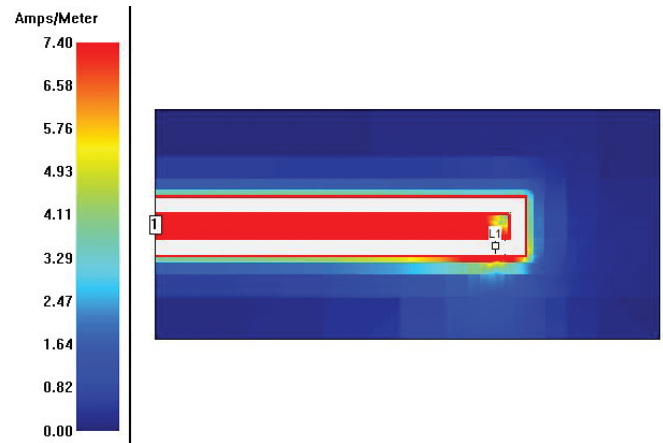


Figure 3 • Current density at 600 MHz for ideal inductor measurement at location Z1.

The inductor is simulated in position Z1. The current density at 600 MHz for the inductor measurement is shown in Fig. 3. Observe that the asymmetry of the circuit configuration results in higher current density on the ground plane nearest the inductor. In the absence of the component the current distribution on the ground plane to the right of the reference plane would be symmetrical.

The S -parameters of the measurement are taken at the reference plane shown in Fig. 1. These S -parameters are fit to two models as shown in Fig. 4, (a) an ideal inductor and (b) a lossy inductor model. The ideal inductor model represents connection of the ideal component directly to the reference plane while neglecting the fixture transmission line for the section labeled Δz in Fig. 2. The lossy inductor model is most useful for fitting measurement of real components, but for the purpose at hand the fixture parasitics included in the S -parameter data will be accounted for by the additional elements R_S , R_P , and C_P . The model fit of the ideal inductor is good, the S -parameters show a decrease in magnitude with increasing frequency indicating a loss mechanism. The lossy model gives a better fit across frequency with the circuit elements C_P and R_P providing the frequency dependent loss (an extra model element to account for skin-effect losses in a real inductor has been omitted from Fig. 4b). At 600 MHz the measured Q of the ideal inductor is 7000 due entirely to the dielectric losses. For a practical inductor the skin-effect losses in the windings will be significantly higher than these dielectric losses. The model parameters are tabulated in Table I. Note that the apparent measured value of the inductor is approximately 0.5 nH higher than the known value. For the one-port measurement of Fig. 1 the conclusion is that the extra length between the reference plane and the component adds this

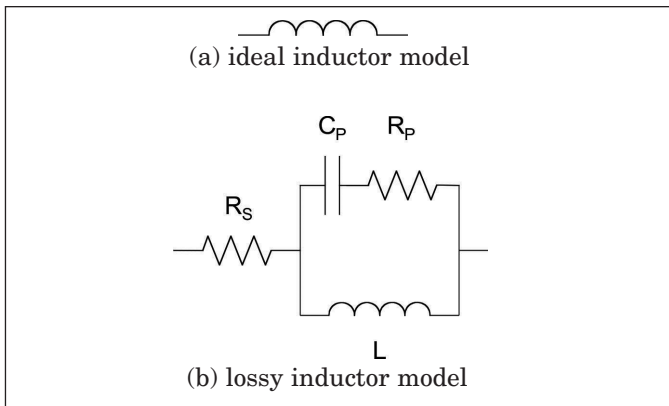


Figure 4 • Inductor models used to fit EM simulation results.

additional series parasitic inductance. In Fig. 2 the length Δz has two associated parasitic inductances. In the ideal case each inductance is $\frac{1}{2} \Delta z$ with one inductance associated with conductor path and the other in the ground path. By inspection of Fig. 1 it is expected that not only will the inductance $\frac{1}{2} \Delta z$ change depending on measurement location $Z1$ and $Z3$, but the conductor path and ground path inductance will most likely not be equal as the physical distance between the DUT terminal and the measurement reference plane is longer on the ground path than for the conductor path.

The variation in the fixture inductance $\frac{1}{2} \Delta z$ depending upon the location of the component is readily apparent by moving the inductor to location $Z3$ and extracting the model from the simulation results. The current density at 600 MHz for the inductor at location $Z3$ is shown in Fig. 5. While the symmetry of the ground plane currents are preserved with the inductor at location $Z3$ compared to that of location $Z1$, the physical length between the grounded terminal of the inductor and the reference plane has increased. The model parameters are tabulated in Table II. The apparent measured value of the 3.4 nH inductor is approximately 0.8 nH higher than the known value. There is an additional 0.3 nH parasitic embedded in the measurement.

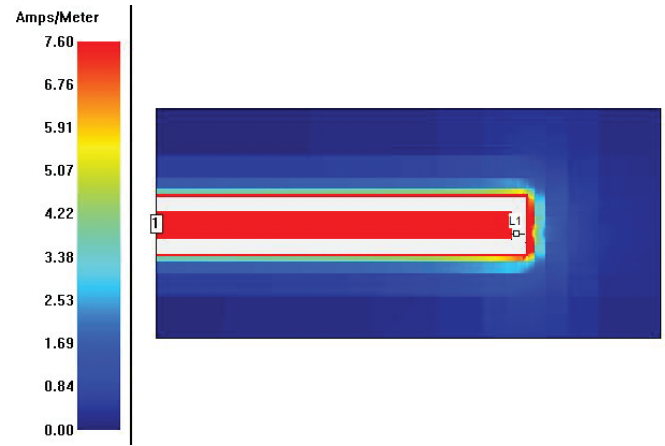


Figure 5 • Current density at 600 MHz for ideal inductor measurement at location $Z3$.

Inductor model values extracted from EM simulation

Model	Element	Value
ideal	L	3.92 nH
lossy	L	3.85 nH
	C_P	0.17 pF
	R_P	14.7 Ω
	R_S	0.113 m Ω

Table I. Location $Z1$ model.

Model	Element	Value
ideal	L	4.16 nH
lossy	L	4.09 nH
	C_P	0.15 pF
	R_P	16.1 Ω
	R_S	0.114 m Ω

Table II. Location $Z3$ model.

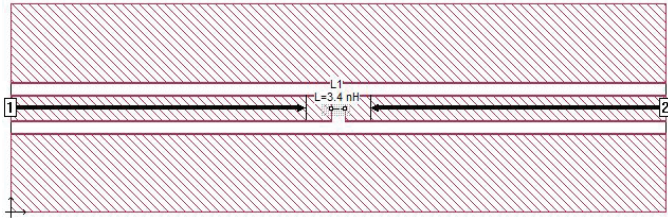


Figure 6 • Simulation circuit for two-port measurement of the inductor.

Consider measuring the same inductor as a two-port and extracting models. The Sonnet simulation for the two-port configuration is shown in Fig. 6 with the resultant model tabulated in Table III. The apparent value of the inductor is approximately 1 nH higher than the known value. The unknown test fixture structure corresponding to Δz of Fig. 2 is approximately doubled for the two-port configuration, though the contribution of the transmission line inductance $\frac{1}{2} \Delta z$ appears to be reduced for each port compared with that of location Z3 as the inductance error is not greater than 1 nH. For a lab measurement the inductance error will be compounded by the uncertainty associated with construction and calibration of a two-port fixture. Consider these results carefully and proceed with caution when directly inserting a measured two-port file (or model extracted from a two-port file) into a simulation that compensates for transmission line effects up to the terminals of the component. The potential for magnifying the uncertainty of the simulation is increased by accounting for the surrounding circuit parasitics twice, once in the measurement and the second time in the simulation. The one-port approach has the potential for less error in the measured data.

Model	Element	Value
ideal	L	4.36 nH
lossy	L	4.30 nH
	C_P	0.13 pF
	R_P	18.3 Ω
	R_S	0.114 m Ω

Table III • Model values for 3.4 nH inductor from two-port measurement as shown in Fig. 6.

A series of simulations is run with the ideal inductor varying from 1 nH to 30 nH at locations Z1 and Z3. The ideal inductor model is extracted from the simulation S-parameters and the difference between the known inductance value and the model value along with the percent error is plotted in Fig. 7. The parasitic series inductance increases with the known inductance value and the percent error approaches 5% with location Z1 being closer to the actual inductance value. For small inductance values the error is limited by the transmission line inductances related to $\frac{1}{2} \Delta z$. The apparent

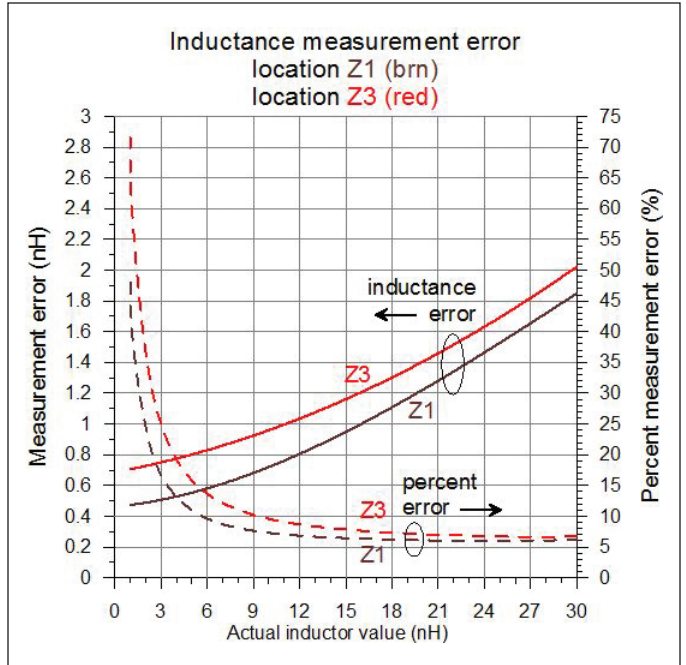


Figure 7. Inductance error for measurement in test fixture of Fig. 1.

increase in inductance error from 0.5 nH to 1.8 nH with increasing inductance value is due to the transmission line element $C\Delta z$ shown in Fig. 2 in parallel with the DUT. The apparent inductance value is pulled higher by the parallel resonance or the DUT inductance and $C\Delta z$. Extending the measurement frequency to 10 GHz will reveal the self resonance of the DUT with $C\Delta z$. Unfortunately, measurement of the inductor from 200 – 1000 MHz is not sufficient to unmask the contribution of $C\Delta z$, and its contribution to the deembedded value is to increase the measurement error. The best achievable measurement for an unknown ideal inductor between 1 and 30 nH at 200-1000 MHz without deembedding the transmission line will have an error of 0.5 to 2.0 nH.

Can the fixture effects be deembedded from the component measurement? The practical answer is no. The fixture effects can be minimized by careful design of the fixture, but they cannot be removed. Measurement of the inductor at low frequencies, where the fixture effect is negligible, improves the result for an ideal inductor, but the useful information about parasitics of the type shown in the lossy model of Fig. 4b will be lost.

Measuring the Ideal Capacitor

Simulation of an ideal capacitor from 200 - 1000 MHz in locations Z1 and Z3 provides an estimate for $C\Delta z$. A 15.6 pF capacitor is simulated in location Z2. Location Z2 is equivalent to location Z1 for a single component measurement. For clarity when measuring multiple components the capacitor is positioned at location Z2. The resulting S-parameters of the EM simulation are fit to a

EM SIMULATION

simple capacitor model as illustrated in Fig. 8. The model values are tabulated in Table IV. Similarly, the capacitor is simulated at location Z3 with the resultant model values tabulated in Table V. For both locations there is approximately a 0.1 pF error in the capacitance measurement due to $C\Delta z$ of the fixture transmission line parasitic. The inductance value for each location corresponds to the inductance error noted earlier, where the Z1 location was observed to be around 0.5 nH and the Z3 location was approximately 0.8 nH. The estimate of $C\Delta z$ and $\frac{1}{2} \Delta z$ obtained from the capacitor measurement could be used to compensate for errors in the inductor measurement. The parasitic series inductance and uncertainty in capacitance value of a real capacitor confounds the ability to obtain estimates of $C\Delta z$ and $\frac{1}{2} \Delta z$. This example would require a 15 pF capacitor with a tolerance $\ll \pm 0.1$ pF in order to resolve $C\Delta z$. Extending the simulation of the inductor at location Z1 out to 10 GHz and extracting the transmission line parasitics of the fixture between the reference plane and the fixture yields the values of $C\Delta z = 0.16$ pF and $L\Delta z = 0.48$ nH. This level of certainty would require an accurate one-port calibration at the reference plane from 0.2 to 10 GHz, a difficult proposition at best. It would also require measurement of a DUT fully characterized across the same frequency range. With an EM simulation tool such as Sonnet it is possible to obtain estimates of these values, but it is a challenge to translate this approach to improving lab measurements.

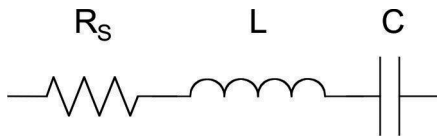
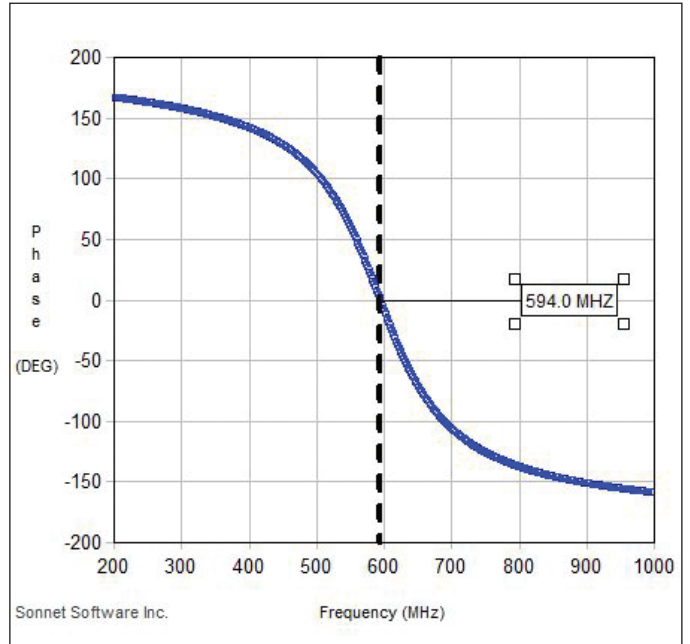


Figure 8 • Capacitor model used to fit EM simulation results.

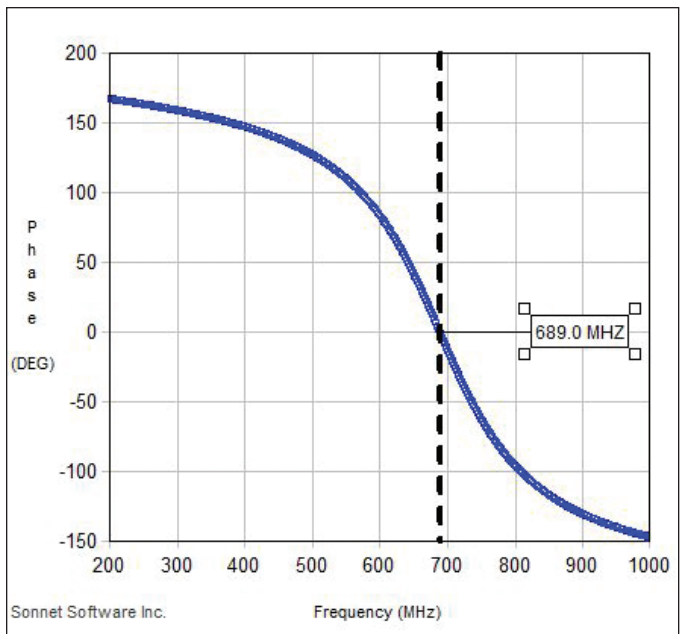
Capacitor model values extracted from EM simulation			
Element	Value	Element	Value
C	15.7 pF	C	15.7 pF
L	0.47 nH	L	0.70 nH
R_S	3.4 mΩ	R_S	3.2 mΩ

Table IV. Location Z2. Table V. Location Z3.

When the capacitor simulations are repeated with values ranging from 1 to 50 pF from 200 - 1000 MHz the resulting model extraction shows a near constant error in the capacitance value ranging from 0.10 to 0.15 pF. The error in capacitance value is bounded at 15% for 1 pF and drops below 1% at 50 pF. The series inductance, while not fixed, remains in the neighborhood of 0.4 nH for location Z1 and 0.7 nH for location Z3. If the series inductance of the capacitor could be bounded it may be possible to use information from the capacitor measurement to improve



(a) Inductor at Z1 with capacitor at Z2, $f_{res} = 594$ MHz



(b) Inductor and capacitor at Z3, $f_{res} = 689$ MHz

Figure 9 • Phase of S_{11} for the parallel measurement of 3.4 nH and 15.6 pF.

the error associated with a subsequent inductance measurement by bounding the contribution of the fixture effects to the result.

Measuring the Ideal Inductor and Capacitor at Parallel Resonance

Consider simulating the measurement of the 3.4 nH inductor and the 15.6 pF capacitor in parallel. The capac-

itor value was chosen to resonate with the inductor of interest at $f_{res} = 691$ MHz which is in the frequency range of design interest. The first simulation places the inductor at location Z1 and the capacitor at location Z2. The second simulation places the inductor and capacitor both at location Z3. From the simulation S-parameters the frequency at which S_{11} passes through 0° is where the parallel resonance occurs. The graphs of S_{11} phase and the frequency of parallel resonance for each simulation is shown in Fig. 9. Using the resonant frequency f_{res} and knowledge of the capacitor value an estimate of the inductance value can be obtained. Table VI gives the results for measurement at each location. The inductance error with the components at locations Z1 and Z2 is 1.20 nH. The error for location Z3 is 0.02 nH. This means that the error in the measurement of the ideal inductance value is limited by the degree to which the capacitance value is known and the series inductance of the capacitance itself. The contribution of the fixture to the measurement error has been minimized, while facilitating an inductance measurement in a frequency range of interest.

The current density of the resonant circuits for both cases are shown in Figs. 10 and 11. When the inductor and capacitor are at locations Z1 and Z2 the resonant current on the ground path must travel around the transmission line structure as seen in Fig. 10. This circuitous route results in an additional 1.2 nH of inductance in series with the desired inductance. With the inductor and capacitor at location Z3 the resonant current on the ground path is confined to the area immediately in the vicinity of the components. The effect of the fixture on resonant frequency is minimized.

Capacitance (known)	C = 15.6 pF	
Component location	Z1, Z2	Z3
Resonant frequency f_{res}	594 MHz	689 MHz
Inductance value	4.60 nH	3.42 nH
Inductance error	1.20 nH	0.02 nH

Table VI • Obtaining the inductance value from the resonant frequency and capacitor value.

Applying Simulation Results to Lab Measurements

A lab measurement technique using a good quality capacitor with parasitic series inductance bounded by the construction geometry would introduce additional uncertainty into the measurement on top of the uncertainty in the capacitance value itself. From the simulation results in Table IV it should be possible to measure the capaci-

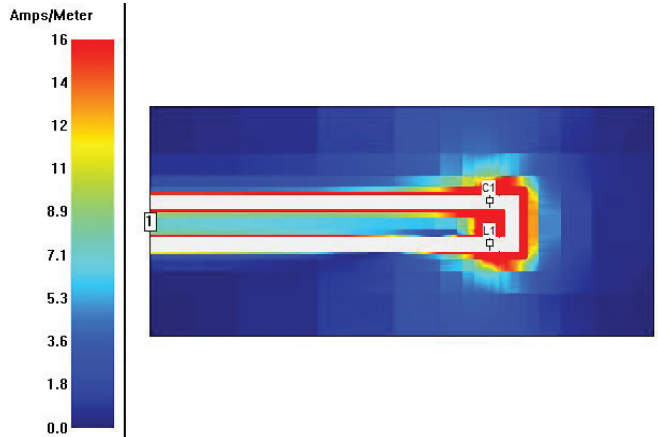


Figure 10 • Current density at 594 MHz for inductor at location Z1 and capacitor at location Z2.

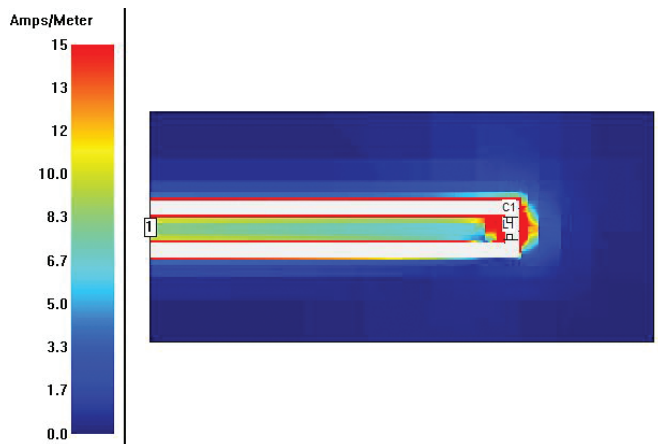


Figure 11 • Current density at 689 MHz for inductor and capacitor at location Z3.

tance value of a 15 pF capacitor to within 1% in the presence of the fixture parasitics. An AVX ACCU-P 0603 15 pF thin-film chip capacitor has a self resonance frequency of approximately 1.5 GHz which gives a series inductance parasitic on the order of 0.75 nH. The manufacturer's data sheet for this series of parts indicates the self inductance is approximately 0.78 nH [3]. The manufacturing process for thin-film chip capacitors is able to maintain tight control over the dimensions and material structure so that the variance of the self inductance should correspond to the variance in package geometry with the implication being that the inductance should be well bounded and consistent across different capacitor values within the same package outline.

The capacitor could be measured first in location Z2 and the capacitance extracted from the measurement. The inductor and capacitor could then be measured at location Z3 as a parallel resonant circuit. The measured resonant frequency could be skewed by the additional series parasitic inductance of the capacitor. For an inductor of approximately 3 - 5 nH it is expected that the 15 pF

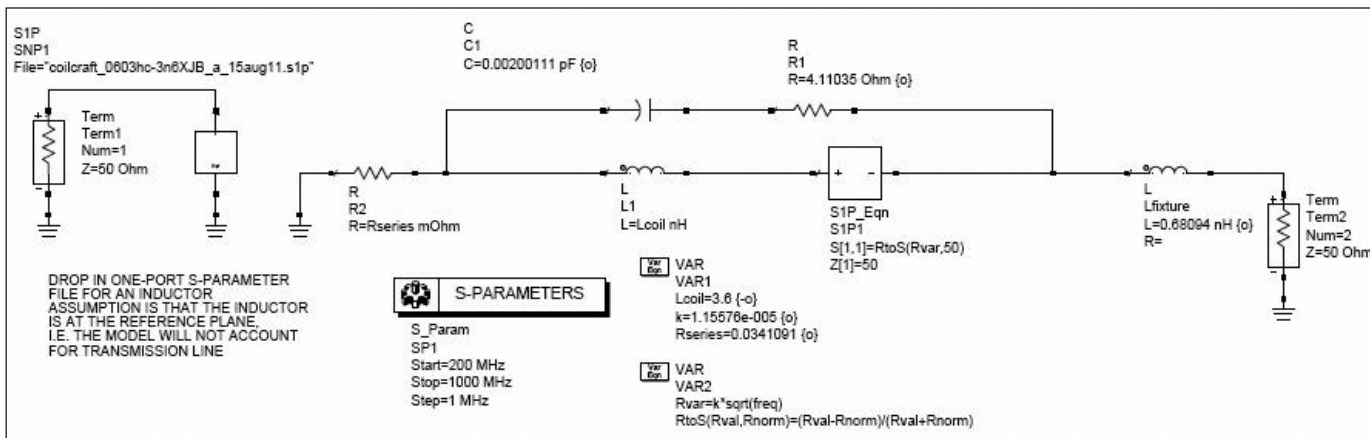


Figure 12 • Extracting inductor model from one-port S-parameter measurement data.

capacitor should be at least an order of magnitude larger than the parallel parasitic capacitance. Using the measured capacitance value and resonant frequency the inductance of the measured circuit can be obtained. An estimate of the unknown inductor value can be obtained by subtracting out the series inductance of the capacitor. The actual inductor value will be bounded by the uncertainty on the capacitor series inductance and the measurement of the capacitance value. The final step would be to measure the inductor in location Z1 and use the inductor value obtained from the resonant frequency measurement to separate out the fixture parasitics from the model of the inductor based on the S-parameter measurements.

In the lab an Agilent N5230A Vector Network Analyzer is calibrated from 200 – 1000 MHz using a TRL calibration kit and associated measurement fixture designed to achieve a calibrated measurement reference plane at the location indicated in Fig. 1. The measurement objective will be to derive a model for a Coilcraft 0603 chip inductor (0603HC-3N6XJB) across the frequency range of measurement.

Measuring the bare fixture across 200 - 1000 MHz and extracting a model of the parasitics yields a capacitance measurement of 0.12 pF corresponding to $C\Delta z$ in Fig. 2. While this value is of limited use, it indicates that the fixture construction is consistent with the simulation

Element	Value
C	15.9 pF
L	0.97 nH
R_S	59 mΩ

Table VII • Capacitor model values extracted from measurement.

model and may help bound the error in the subsequent measurements. Next an 0603 thin-film chip capacitor (AVX 06033J150GBT) is measured in location Z2. The model values extracted from the measurement are tabulated in Table VII.

With the capacitor and inductor at location Z3, S_{11} is measured. The circuit exhibits a parallel resonance at 603 MHz. Using the measured capacitor value from Table VII the corresponding inductance of the resonant circuit is 4.40 nH. Accounting for the series inductance of the capacitor using a low estimate of 0.75 nH yields an inductor value of 3.63 nH. If the fixture $C\Delta z$ parasitic is removed from the capacitance measurement or a slightly larger value for the series parasitic inductance of the capacitance is used the resulting estimate of the inductor value ranges between 3.6 nH and 3.7 nH. The inductor is measured in location Z1 and a model fit to the measured S-parameter data.

The inductor model extracted is based on the model used by Coilcraft [4] with an additional inductance $L_{fixture}$ in series to account for the $L\Delta z$ of the fixture. To compensate for frequency dependent losses primarily associated with skin effect the model includes a frequency dependent term

$$R_{VAR} = k\sqrt{f}$$

where f is the frequency in Hertz. When fitting a model to S-parameters measured across a broad frequency range this term provides a better fit to the data than using a fixed resistance. The ADS implementation of the model used for extraction is illustrated in Fig. 12 and the resultant values are tabulated in Table VIII. For comparison, the model extracted directly from the S-parameters is tabulated in Table IX. For the direct fit case the $L\Delta z$ series parasitic of the fixture cannot be separated from the measurement and so the parasitic element $L_{fixture}$ is not included in the extraction. The other model elements differ only slightly between the two approaches.

The parasitic R_2 would account for fixed resistive losses in the inductor and would be important for lower frequencies. From S-parameters measured at 200 - 1000 MHz there is insufficient information to separate out the fixed series resistance from the skin-effect losses so the value of R_2 shown in Tables VIII and IX has no practical use in this example. If the inductance had a ferrite core, the element

Inductance extracted using resonant frequency, other elements fit from S -parameters

Element	Value
L	3.6 nH
C_1	0.002 pF
R_1	4.1 Ω
R_2	0.03 m Ω
$k (R_{VAR})$	1.16×10^{-5}
$L_{fixture}$	0.68 nH

Table VIII

Inductance direct fit to S -parameters

Element	Value
L	4.3 nH
C_1	0.002 pF
R_1	3.2 Ω
R_2	0.09 m Ω
$k (R_{VAR})$	1.16×10^{-5}
$L_{fixture}$	NA

Table IX

R_2 could be used to account for the ferrite loss to a first order and then the frequency dependent losses of the core absorbed into R_{VAR} . Depending on the ferrite core material and the span of frequency for the extraction this may yield useful results, but more loss elements will be needed for a true fit of a ferrite loaded inductor. A measurement of the dc resistance could also be used to fix R_2 . It should be acknowledged that the extracted values for this inductor do vary from the model parameters provided by the manufacturer in the product technical notes [4].

Summary

Using EM simulation of ideal components the possible error sources in measurement of a two-terminal passive component were explored. The insight gained from the simulations can be directly applied to measurements in the lab. To remove the effects of fixture parasitics in the measurement of an inductor, additional measurements are taken of a capacitor and the parallel resonance of the capacitor and inductor. From these measurements it is possible to separate out the component inductance from

the fixture transmission line inductance $L\Delta z$ to improve the overall accuracy of the extracted model.

References

1. Inan, U. S. and Inan, A. S., *Engineering Electromagnetics*, Addison Wesley Longman Inc. 1999.
2. Collin, R. E., *Foundations for Microwave Engineering 2nd Edition*, Wiley-IEEE Press, 2000.
3. AVX RF Microwave Products databook Version 11.6.
4. Coilcraft "SPICE Model-0603HC", Document 158-1, Cary, Illinois, 16 March 2009.

Author Information

John B. Call is a Staff Engineer with Thales Communications Incorporated USA in Clarksburg, Maryland where he designs handheld and mobile radios for mission critical applications. He received his B.S.E.E. degree from the University of Utah and his M.S.E.E. degree from Virginia Polytechnic Institute and State University. Interested readers may contact him by e-mail at: john.call@thalescomminc.com.