High speed/mm-wave measurement-based model development: uncertainties and model sensitivities

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Abstract

Device models must increasingly cover performance well above 100 GHz and often measurements form a part of that modeling process. Model extraction processes interact with mm-wave measurement uncertainties in sometimes surprising ways that can lead to unexpected model results. This paper explores the uncertainties of broadband S-parameter measurements (noise mechanisms and their distributions, on-wafer calibration characteristics, instrument linearity, etc.) and quasi-linear measurements such as AM/AM and AM/PM (affected by instrument linearity, power accuracy vagaries, nonlinear match oddities, etc.) as well as common model extraction techniques for both passive (primarily isolated and coupled inductors and capacitors) and active devices (primarily transistors). There are cases in the parameter space when uncertainty increases in the former and sensitivity increases in the latter can overlap, making for a worst-case scenario. This paper explores these measurement-based extraction relationships through an analysis of measurement hardware (from kHz range to mm-wave), actual measurement/extractions examples and Monte Carlo simulations. The objective is to outline steps that can be taken to minimize unfortunate uncertainty/sensitivity collisions and hopefully improve the resulting extractions.

Introduction

With increasing needs for fast design success in microwave and mm-wave circuits and systems, more accurate device and circuit models could be very useful. While there are dozens of ways of generating such models (e.g., [1]-[13], measurement-based compact and behavioral models (e.g., [11]-[13]) are quite popular because of their specificity and possibility of including many dependencies. As the underlying measurements (here the focus will be on those made with Vector Network Analyzers, VNAs, and spectrum analyzers) are imperfect, an important question is how do any measurement issues propagate to the final extracted model? As frequencies increase, measurement uncertainties tend to as well and often take on additional dependencies so one might expect a more complicated story with mm-wave models.

For relatively high reflection and transmission levels, noise in S-parameter measurements is largely multiplicative (i.e., not noise floor-based but scales, in absolute terms, with the reflection or transmission wave amplitude) but usually increasing monotonically, in linear terms, with frequency. On-wafer calibration characteristics can vary wildly and popular techniques (line-reflect-match and its derivatives) are dependent on both line quality and on certain characteristics of the match standard behavior. Further complicating the picture, particularly above 100 GHz, are increases in probe-to-probe coupling and the possibility of extraneous modes propagating on either the calibration standard substrate or on the DUT wafer. The outcome is an extra-linear increase in uncertainties at high frequencies with particular strength at low transmission and reflection levels. Instrument linearity effects can also vary wildly (from >30 dBm third order intercept products to ~<10 dBm at 110 GHz depending on the structure) which can be important in active device characterization. For quasi-linear measurements, the receiver linearity (including match linearity) is obviously even more important. The RF power level is often critical for model development and the accuracy of this can be impacted by fundamental vs. integrated power differences particularly at higher mm-wave frequencies.

On the model extraction front, many processes are nonlinear with respect to S-parameters with great sensitivity increases as full reflection or full transmission is approached. This can create interesting situations since the noise of the measurements is locally maximizing in that area so modifying the measurement setup might help. A number of extracted parameters (e.g., inductor Q, gate-drain capacitance, etc.) rely on multiple S-parameters so it can make a difference if the uncertainty in those S-parameters is correlated or uncorrelated. Understanding those relationships might allow one to alter the setup to minimize correlated uncertainty increases. Further, extrapolation is frequently employed and works from some of the higher frequency measurement data where uncertainties may be increasing. Modifying what portions of the data set are most heavily weighted in the extrapolation may be advised.

The first section of this paper will explore some of the more important uncertainty terms and their behaviors. The primary focus will be on VNA measurements on-wafer and in-fixture but some nonlinear measurements will also be considered. The second part of the paper will explore some popular model extraction mechanics (from data to output model) and the sensitivities to various data inputs. This analysis cannot be exhaustive but will look at some common approaches. Finally, some approaches will be discussed for mitigating cases of potentially elevated sensitivity impact.

Measurement uncertainties

Measurement uncertainties for VNA and other instruments commonly used in RF modeling have been covered extensively in the literature (e.g., [14]-[17]). It will be useful, however, to cover some points here with regards to the parts of the parameter space often exercised in modeling efforts. To establish context, consider a generic block diagram of a VNA shown in Fig. 1. There are downconverting receivers with some noise floor and some linearity limit (in the gain compression sense). As of the writing of this paper VNA systems exist with effective third-order intercept products over 30 dBm to 145 GHz and over 20 dBm to 220 GHz. There may be elements of nonlinear match that also contribute to linearity. Broadband systems (to 110 GHz and beyond) often have multiple receivers and source system multiplexed together and those multiplexing sections can introduce frequency-local drift and other effects. A spectrum analyzer can often, in terms of uncertainty components, be thought of as a single receiver in Figure 1 with mechanisms of interest including linearity noise, drift, and absolute power measurement accuracy.

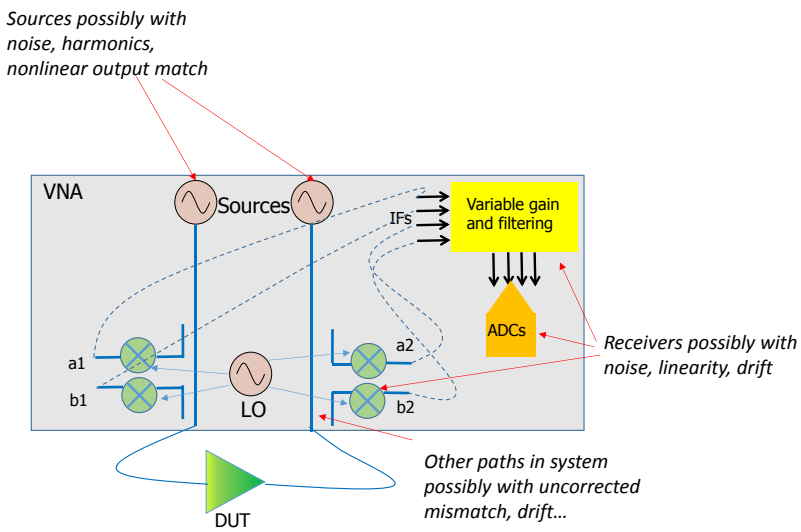


Figure 1. A generic VNA block diagram is shown here. Hardware-related uncertainty elements include noise floor, proportional noise conversion related to the LO, receiver linearity, source harmonics and power accuracy, drift and other items.

The noise on data is an interesting point since there is the noise floor which acts as an additive contributor to the measurement but there is also a multiplicative element that is sometimes less well-understood. Consider close-in phase noise riding on the downconverter LO (close in the sense that it is within an IF bandwidth or resolution bandwidth). The amount of this noise that converts to the IF is at least partially related to the amplitude of the incoming RF signal (and its close-in phase noise) hence the added noise amplitude is multiplicative in part. Thus the absolute noise contribution to the digitized IF is not constant with signal level and indeed becomes constant in dB terms at sufficiently high levels. As an example, trace noise is plotted as a function of power in Fig. 2. At low levels, the noise floor dominates so the amount of noise in dB terms (really dBc terms) decreases with signal level. At higher signal levels, the noise in dB terms becomes roughly constant as the converting phase noise dominates. This measurement is of a ratioed parameter which complicates the analysis somewhat since correlated noise between the test channel and reflected channel cancel but the general trend holds for both wave parameters and ratioed S-parameters.

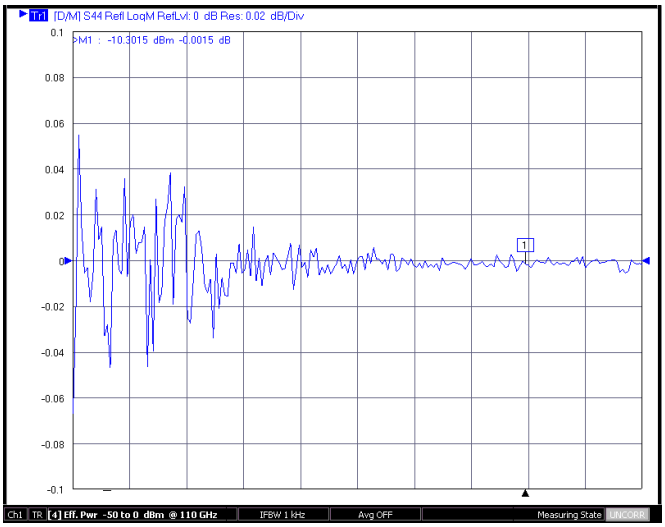


Figure 2. Trace noise as a function of signal level is plotted here for a ratioed parameter. At low signal levels, an additive absolute noise floor dominates while at higher levels, a noise signal largely multiplied on the signal dominates.

From a measurement setup point-of-view, one would want a high enough level to get out of the noise floor limited zone. If the DUT cannot be driven that hard (e.g., a bare transistor), padding down the drive after the reference coupler (using internal step attenuators if available) can help optimize trace noise.

Repeatability can be a dominant term, particularly on-wafer and for mm-wave measurements (e.g., [18]). In coax with well-maintained connectors, repeatability can be in the -50 to -60 dB range (as measured by a difference in linear reflection expressed in dB terms) even to 145 GHz. In fixtured or on-wafer scenarios, where most model extraction is done, this often reduces to the -30 to -45 dB range (getting worse at higher mm-wave frequencies) depending on details of the connection/probing process. Looking at composite VNA uncertainty curves (Fig. 3, including all of the mechanisms discussed in this section among others), the repeatability change can clearly have a significant effect. The behaviors in Fig. 3 are with a semi-automatic probe station and improvement is, of course, possible with fully automatic probing and regular refreshes/replacements of probe tips and calibration substrates.

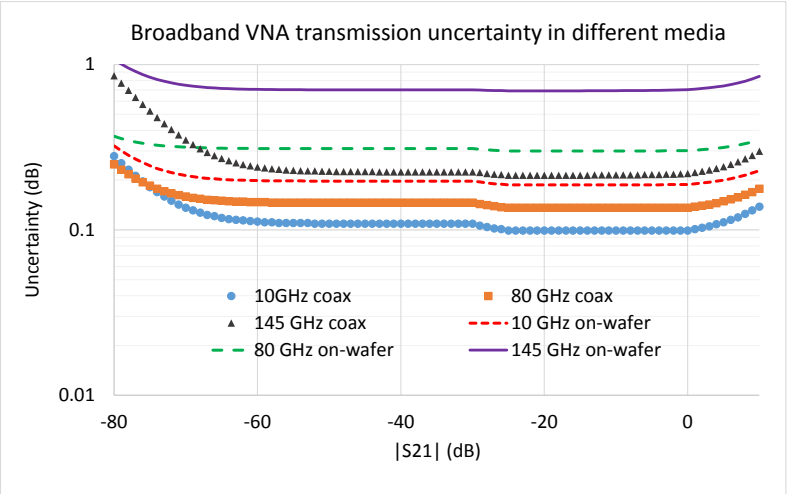


Figure 3. Composite uncertainty curves for a transmission measurement (DUT assumed well-matched) are plotted here versus frequency and the connection environment. Practical repeatability levels can have a dominant impact on the outcome.

Another aspect of repeatability, mainly in on-wafer measurements, is related to probe-to-probe crosstalk. Depending on the probes and DUT layout, these levels can reach ~-30 dB at 110 GHz and may be influenced by neighboring structures on wafer (if surface modes are excited) or things such as the position of the microscope objective (if radiated coupling is

relevant). These environmental dependencies can have a dramatic impact on S-parameter data particularly resonances (as might be used to extract certain device parasitics). As an example, a transmission resonance is used for an inductance extraction based on the resonant frequency and an already extracted capacitance. Aside from the question of how that capacitance uncertainty propagates through, -40 dB crosstalk of random phase can result in ~1% uncertainty in the resonance frequency (see Fig. 4). Assuming a simple single-pole resonance, this leads to a 2% inductance uncertainty just from probe crosstalk. Improvements in this behavior can be challenging (involving perhaps different probing geometries or layouts) but if narrowband crosstalk improvements are possible (e.g., with absorbers if radiated coupling) then working around the device self-resonance is one approach.

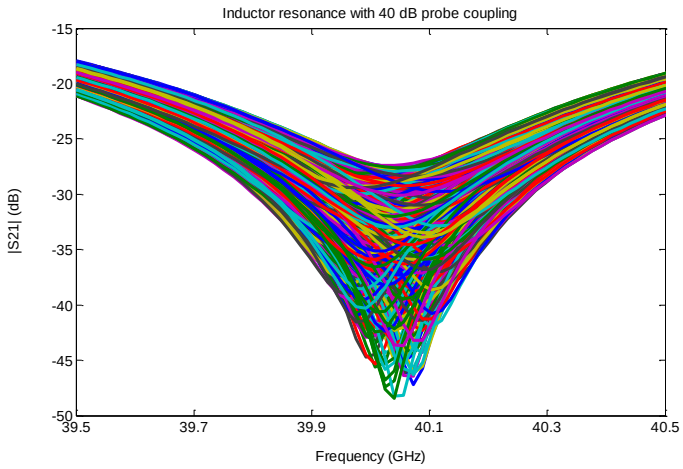


Figure 4. The effect of randomly phased -40 dB probe crosstalk on a measured resonance is plotted here using a Monte Carlo simulation. The resonant frequency is being used as part of the extraction and a significant spread in values are possible.

Yet another repeatability term is drift defined as response change after calibration/de-embedding time independent of any change in the DUT. This parameter is a strong function of the setup (external cable run lengths, intervening network presence) and environment. It is not unusual for drift to be predominantly in phase (if the mechanism is cable change) but, through the calibration and de-embedding, this can transfer into magnitude of the S-parameters. Nor is it unusual for the amount of drift to monotonically increase with frequency. It can happen that drift is heightened in local frequency regimes due to reflectometer degradations, the presence of a diplexer in front of the measurement reflectometers (e.g., [22]) or other causes. This leads to a local uncertainty increase, particularly in broadband measurement systems, and a non-intuitive level of variation in model parameters. If hardware improvements are not possible, interpolation of surrounding data is sometimes done on the S-parameter level.

Calibration and de-embedding processes are often a concern and they can indeed be determinative of uncertainty in certain cases (e.g., [13]). It is possible the line impedances are inconsistent or standards definitions are incorrect but a well-executed calibration in defined media is rarely the dominant uncertainty term is model extraction measurements. There is not space in this paper to work through common uncertainty sensitivities with calibrations but many excellent references exist (e.g., [19]-[20]). Certain details of the final de-embedding process can, however, be quite important:

- Many parasitic extractions rely on the phase of an S-parameter so proper reference plane placement matters (sometimes even probe placement matters). A series 10 pH inductance only generates a S_{21} phase shift of 4 degrees at 110 GHz. A probe placement difference of 10 μm (probe pads or often on the order of 50 μm in size) can generate a S_{21} phase difference of about the same amount (depending on substrate, etc.). Pre-determination of offset lengths (using reference calibrations) can sometimes help here.

- For de-embedding on-wafer interconnects, sometimes standards at the inner plane are not well-defined which can imperil accuracy of that last segment. Often, the only de-embedding needed is for feed lines which does help reduce the sensitivities. It is a very simple example, but consider the effect of a standards issue (50 μm error in offset lengths) and system drift (5C temperature change) for different feed structures. The Monte Carlo simulation results are shown in Fig. 5. As the network gets lossier or more mismatched, the de-embedding process

has to work harder and the same distortions to the measurement have increased impact on the de-embedded result. There are many de-embedding approaches available and some are tailored to specific network loss/mismatch or standards availability profiles. Searching for a more optimal method can sometimes help.

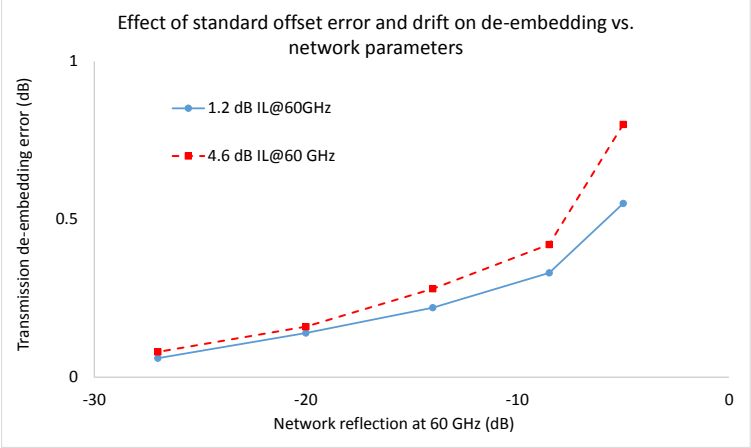


Figure 5. Lossier and more poorly matched de-embedding networks can increase overall uncertainties as the sensitivities to drift or errors on standards can increase. Example increases are shown here for several structures being de-embedded for the same errors on reflection standards (offset lengths) and the same amount of drift.

Quasi-linear measurements are increasingly an important part of modeling exercises and may include AM/AM, AM/PM and IMD measurements or more behavioral modelling approaches measuring DUT harmonics (possibly with modulated signals and possibly under load-pull conditions) whether done with a VNA or spectrum analyzer. Obviously system and receiver linearity can play a role in these measurements but often the signal level range can be shifted to minimize those problems (although may exacerbate noise floor issues). Source harmonics can directly impact any harmonic measurements and sometimes in complicated ways as the DUT operating state may be modified by the contamination. As an experiment, consider an amplifier harmonic measurement where a 2nd harmonic of the input was injected with variable power and phase. The DUT in this case had relatively little linear gain at the 2nd harmonic frequency so the effect on the output harmonics will mainly be in how multiple frequency components interact within the DUT. When the injected harmonic levels are -25 dBc, there is relatively little scatter in the harmonic measurements (as the injection phase was varied) aside from analyzer noise floor and repeatability effects as shown in Fig. 6. At -10 dBc injection harmonics, however, there was considerably more variation and that variation was a function of main drive. The amount of variation at very high drive levels actually decreased in this case as main signal compression started to dominate DUT behavior. The central point of this experiment is that source harmonic contamination can have a complicated effect on the harmonic measurement uncertainty and those effects can be rather large. External filtering or drive level manipulation (attenuators versus instrument level control) can change the source harmonic distribution.

Absolute power accuracy becomes important both for establishing the known drive levels in a quasi-linear measurement and in their referencing in the model (power scale for AM/AM, IMD, etc.). Source power accuracy also maps through into quantities such as absolute IMD levels through receiver calibrations. This topic brings in another collection of uncertainty mechanisms including basic calibration accuracy and linearity of the power sensors involved, uncorrected mismatch between sensor and source, etc. Less often considered but more important at mm-wave frequencies is exactly what a power calibration means when harmonic contamination is non-negligible. Since power sensor bandwidths in this range (sensing bandwidth not video bandwidth) can be as high as 110 GHz (as of this writing), all harmonic content is included in the power calibration. During measurement, when only the fundamental is being analyzed by the VNA or spectrum analyzer, the power being reported at that time may be lower (close to 1 dB lower for -20 dBc harmonics). Since it may not be obvious, *a priori*, whether the DUT will be reacting to the full integrated power or some portion of it, understanding the power spectral distribution may be important in interpreting results.

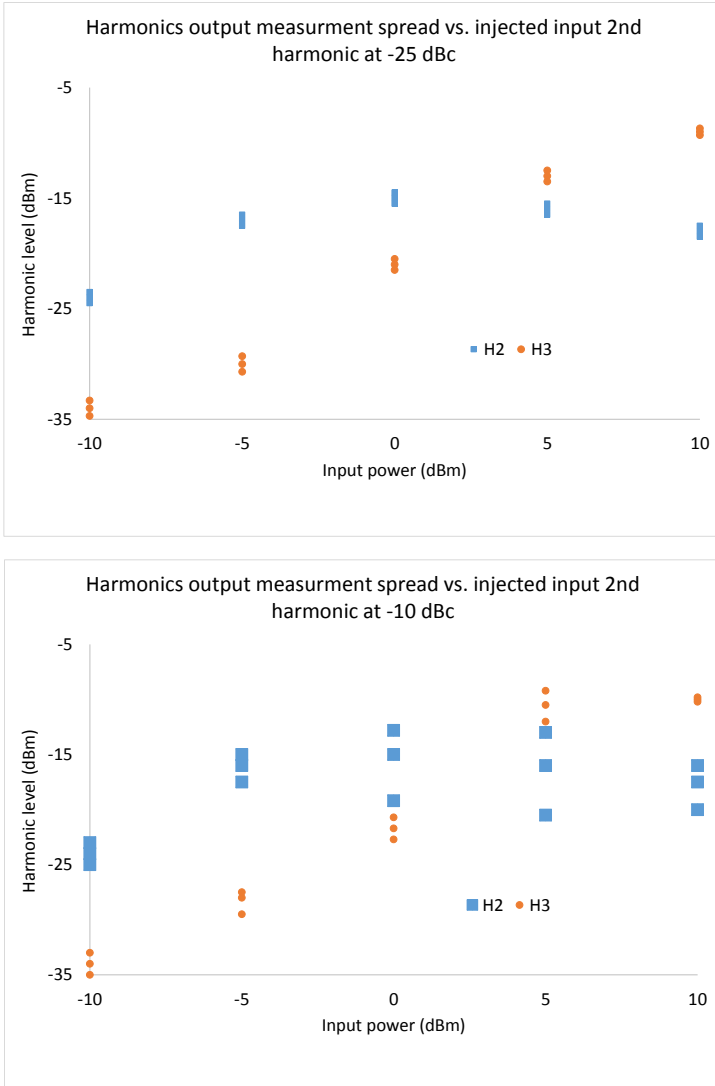


Figure 6. Harmonic measurements at the output of a DUT can be affected by source harmonics in complicated ways since those input harmonics may alter the DUT state. Here a 2nd input harmonic is intentionally injected at -10 and -25 dBc levels and its phase varied relative to the fundamental. The spreads in output harmonic measurements (min, max, median) are plotted here for a DUT with little linear gain at the 2nd harmonic. The number and state of the harmonic generation mechanisms in the DUT play a role.

Another concept that can be important is that of correlation of uncertainties. A number of model extraction processes rely on multiple data inputs (e.g., extraction of series inductor Q relies on both S_{11} and S_{21} data, extraction of a nonlinear transconductance model requires data at multiple harmonic frequencies) and hence it can matter if the uncertainties in the input variables move in the same or different directions in reaction to some issue. Using the inductor Q example, if the calibration happened to be in error (due to a problem with the definition of a standard for example), it is possible for the S_{11} error to move Q in the same direction as the S_{21} error or in a different direction (resulting in cancellation). An example is shown below where different sets of calibration standards were used (for an SOLR calibration in this case) with different errors in standards definitions. The resulting calibrations were then used to measure the same DUT (and inductor in this case). While only magnitude deviations in S_{21} and S_{11} are shown here (from the result with a presumed correct calibration kit), the important observation is that the variations were correlated in all cases with the standards errors but the relative directions of variation were dependent on the details of the calibration problem.

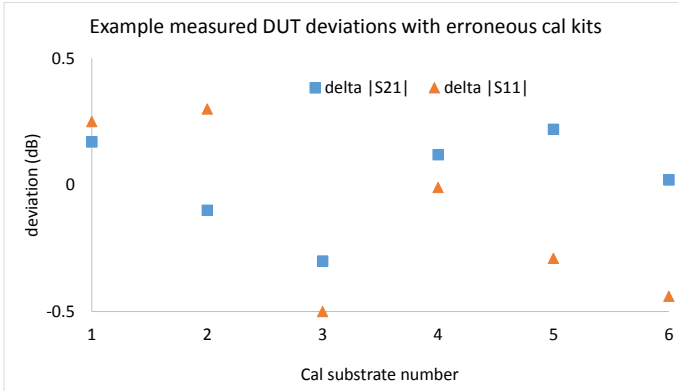


Figure 7. Both reflection and transmission uncertainties can be affected by calibration issues but the relationship between those changes depends on the details of the calibration problem. Errors created by six different problematic calibration substrates are plotted here. Since some model extractions are based on multiple S-parameters, this correlation of effects can present complicated issues at the model level.

One can sometimes mitigate correlation issues with hardware choices (avoiding linearity issues is one) or with a choice of calibration method (e.g., using the unknown thru/SOLR technique causes reflection and transmission uncertainties to be more highly correlated than a method such as SOLT). It is, however, not obvious if correlation is going to help or hurt the overall uncertainty picture.

Extraction

One simple way of illustrating the nonlinear transfer of uncertainty is the extraction of a shunt input capacitance. Assuming one capacitance is dominant and later components are not relevant, one can look at y_{11} which will be proportional to the capacitive susceptance. The full y_{11} is given by

$$y_{11} = \frac{(1 - S_{11})(1 + S_{22}) + 2S_{21}S_{12}}{(1 + S_{11})(1 + S_{22}) - 2S_{21}S_{12}}$$

And the fractional sensitivity to S_{11} deviations is given by

$$\frac{(\partial y_{11}/y_{11})}{(\partial S_{11}/S_{11})} = \frac{-2S_{11}}{1 - S_{11}^2}$$

At low frequency, S_{11} is going to unity so the sensitivity diverges. Further complicating the situation, the capacitance goes as y_{11}/frequency making the sensitivity of capacitance diverge even faster at low frequencies. Similar issues can occur for asymptotic extractions at high frequency (e.g., a substrate capacitance or resistance). Thus even if the uncertainty in S_{11} is constant in absolute terms (and actually decreasing slightly in relative terms), the uncertainty in capacitance can grow dramatically as suggested in Fig. 8. Here the maximum error of S_{11} was constrained and a simulation was run where the uncertainty varied within those bounds. The sensitivity function (2) starts increasing again at higher frequencies as the rate of decline of $|S_{11}|$ slows (this model had a resistive component).

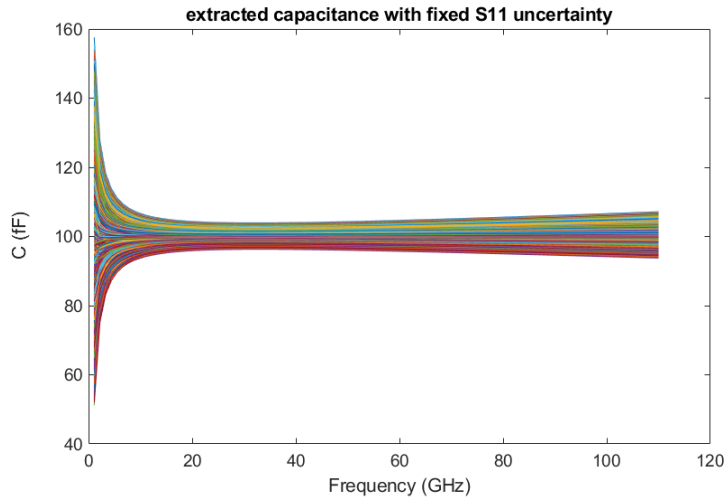
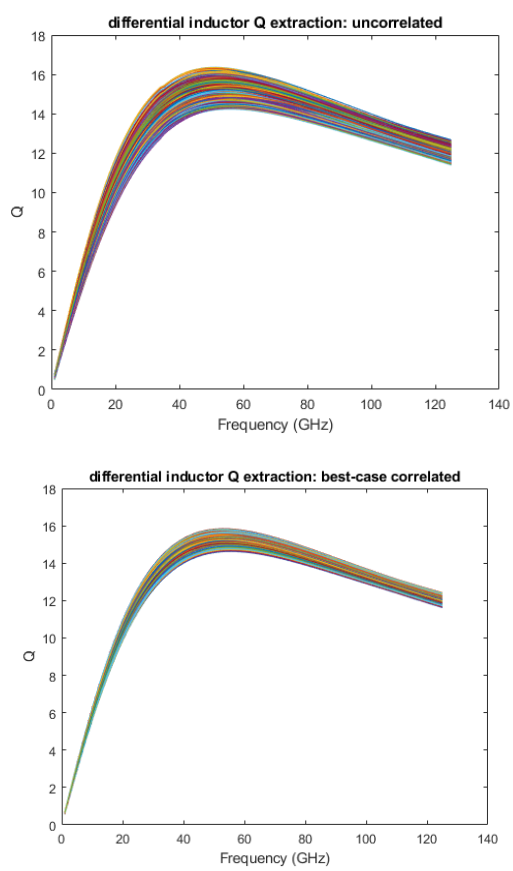


Figure 8. In the simple problem of extracting a shunt capacitance from a S_{11} measurement, the S-parameter uncertainties may be relatively constant with frequency but the uncertainty in capacitance is not. At low frequencies, not only is susceptance divided by frequency to get capacitance but the susceptance changes very rapidly with frequency as $|S_{11}|$ approaches unity.

A next step might be consider Q of a differential inductor which is sometimes evaluated as $Q=Im(1/Y_{12})/Re(1/Y_{12})$. Which brings up the concept of correlation between S-parameter uncertainties discussed in the last section. One can take estimates of the underlying S-parameter uncertainties but if the uncertainties are assumed uncorrelated, the results are different from those with different phases of correlation. Some results are shown in Fig. 9 and the spreads can range from about 1 at 60 GHz to about 2 relative to a mean of 15.2.



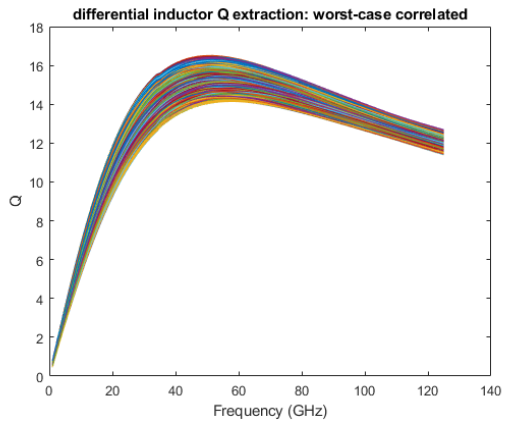


Figure 9. In extracting differential inductor Q, the uncertainty varies greatly if the underlying S-parameters have uncorrelated or correlated uncertainties (and phase of correlation in the latter case). Potential Q errors are thus different for systems limited by, for example, trace noise versus those limited by calibration or linearity even if the S-parameter uncertainties are similar.

Often extracted parameters are dependent on other parameters extracted earlier (e.g., a substrate resistance extraction often uses knowledge of the drain-source resistance) which may be based on data in different frequency ranges (high vs. low in the resistance example) or from data from completely different instruments. Analyzing the correlation of uncertainties in these cases can get complicated.

Transistor transconductance is often found from low frequency measurements with parasitics evaluated using passive extractions like those discussed above but sometimes direct high frequency transconductance extractions are desirable (e.g., when there is a wide distribution in trapping behavior which may arise in some of the newer technologies). The real part of transconductance is often extracted from the real part of Y_{21} (ignoring some correcting terms related to gate-drain capacitance) whose normalized form is

$$y_{21} = \frac{-2S_{21}}{(1+S_{11})(1+S_{22}) - 2S_{21}S_{12}}$$

The correlation topics discussed before can appear here and there are cases where the denominator approaches 0 bringing in heightened sensitivities. In addition, most physical uncertainty mechanisms tend to operate on magnitude or phase parameters but taking the real part of this composite quantity adds a new level of uncertainty obfuscation. An example, albeit a pathological one, is shown in Fig. 10. In the 20 GHz range, because of the frequency behavior of S_{12}, the denominator gets quite small and the sensitivity to noise and de-embedding issues increased resulting in a larger spread. S_{21} continued to increase and the phasing of S_{12} changed at higher frequencies resulting in a less sensitive equation even though the underlying S-parameter uncertainties continued to increase. The cancelling correlation of those S-parameter uncertainties led to a transconductance uncertainty reduction in the 70 GHz range. Finally the spread started increasing again as the S-parameter uncertainty continued to increase (including the increase in a linearity term). While this is an extreme example, it does illustrate the impact the relationship of the absolute values of the sub-terms can have on the extracted uncertainty.

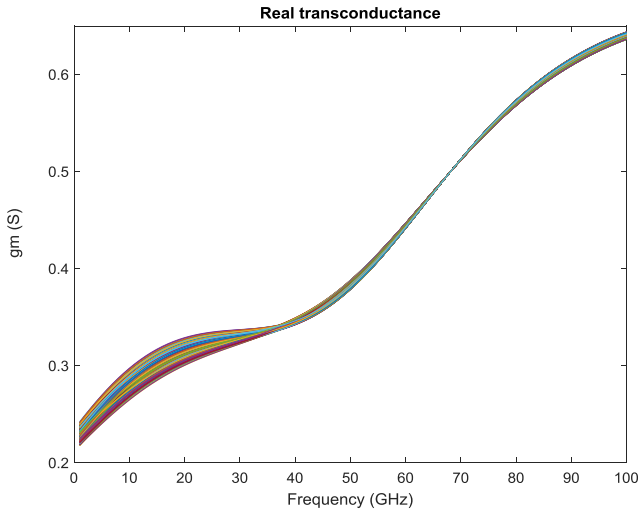


Figure 10. An example transconductance uncertainty spread is shown here for a somewhat pathological (but not unphysical) device. The interaction between the S-parameter values, their uncertainties and the extraction equation led to an unusual frequency dependence of uncertainty.

A very different approach to modeling is behavioral where the DUT is treated more as a black-box. This approach has become for analyzing non-linearities as the physics may be difficult to reduce in a timely fashion for a new device structure or technology. One approach is to acquire data over a range of input parameters (frequency, power, bias, load impedance…) and perhaps interpolate/extrapolate to other values when using the model. Much work has been done in this area (e.g., [11]-[13], [21]) but as in interpolating any multi-dimensional surface, there can be pathological cases. One example is in harmonic generation where there a multiple physical mechanisms generating the product. As power or bias (or some other variable) change, these various contributors can phase in and out making the sum a rather complicated surface. An example 3rd harmonic output surface is shown in Fig. 11. The measurement uncertainty is also shown in the figure and it has a component due to noise floor (that increase in dB terms as the intended harmonic drops in amplitude) and one due to instrument linearity (assuming the fundamental is not being filtered). Depending on which parts of the surface are more important (e.g., the higher drive slope section), one could shift the uncertainty surface by altering receiver padding or filtering (or bias in some cases) into a more favorable state.

Improvement strategies

Regions of heightened sensitivity in extractions are somewhat inevitable so aside from minimizing measurement uncertainties globally, one strategy is to at least try to avoid uncertainty expansion happening at the same time as sensitivity increases. Some observations:

- Parasitic extractions often get more sensitive as $|S_{11}|$ or $|S_{21}|$ approach unity. Avoiding linearity problems with the measurement becomes a high priority (whether through drive level choice, receiver attenuation or measurement hardware choice). For certain calibration and de-embedding choices (particularly defined standards methods), thru line and high-reflect standard characterization accuracy will be more important. Repeatability and noise issues can be globally reduced as discussed earlier.
- For active devices, RF power budgeting can sometimes be important to put the DUT in the desired linearity state while not exciting measurement linearity issues.
- Understanding the correlations between terms can be important, particularly for some extractions. Hardware choices can influence levels of correlation (particularly in linearity and frequency-localized drift behaviors that might be associated with the architecture) as can calibration/de-embedding choices.
- Many extractions use asymptotic frequency assumptions (both low and high frequencies) so having more measurement bandwidth obviously helps. Sometimes, uncertainties degrade rapidly at instrument frequency

extremes so one may need to compute the trade-off between model issues (using less range) and measurement issues (using more range). As an example, a transistor substrate capacitance is sometimes extracted using a high frequency asymptote of a Y-parameter sum. A plot in Figure 12 shows the potential error in capacitance in using only frequency data up to a certain point and projecting the asymptote (the solid curve) and the uncertainties at those frequency for two different measurement systems. The two systems are specified for different frequencies of operation (up to 145 GHz for the second system) but one may want to base the extraction on lower frequency data (perhaps 115 and 130 GHz for system 1 and 2, respectively) for this particular DUT. The measurement uncertainty increase with frequency is partially from increasing trace noise and partially from degrading assumptions about the calibration standards.

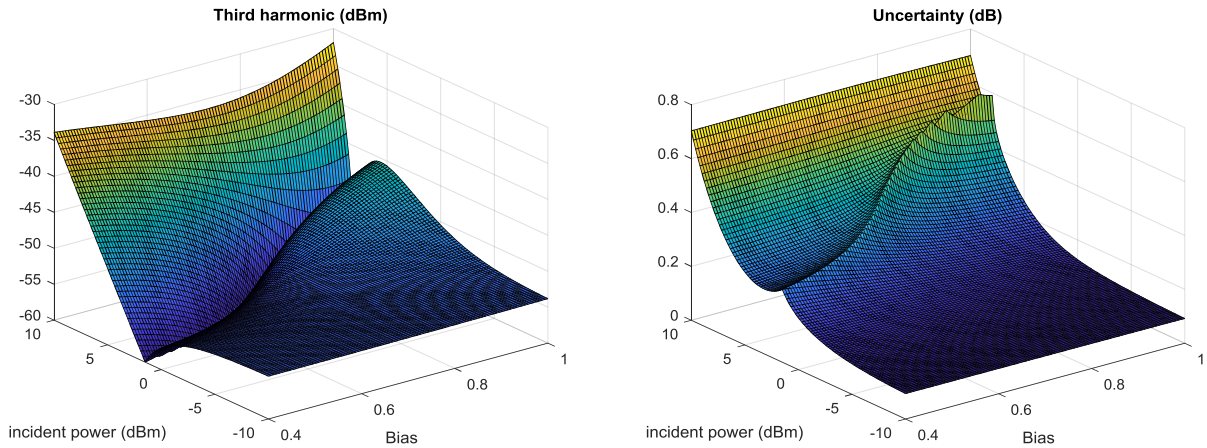
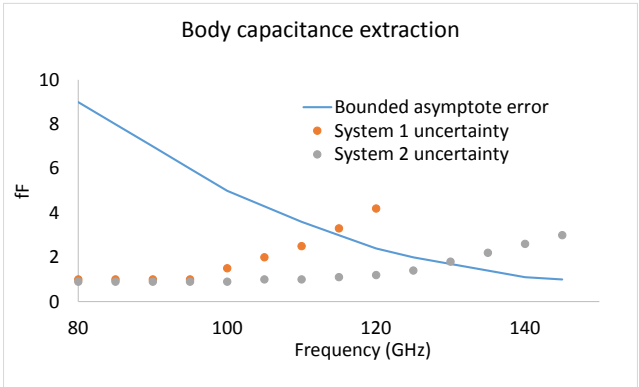


Figure 11. Multi-variable response surfaces can be complicated, particularly for nonlinear responses. This presents some obvious challenges for interpolation and extrapolation often associated with behavioral modeling processes. The measurement uncertainty can also change over the same parameter space but it can potentially be shifted by changing the setup.

- In behavioral model measurements, understanding the response surface is of course important as is the uncertainty profile superimposed on the device surface. At least in terms of linearity and noise issues, one might be able to shift the overlap to reduce the size of interpolation/extrapolation errors.



- *Figure 12. When measurement asymptotes are used for model extraction, there may be a trade-off between extrapolation uncertainty (to the asymptote) and measurement uncertainty.*

Conclusions

Model extraction is an increasingly important task at higher frequencies and measurement uncertainties generally do not decrease. A number of possibly significant mechanisms have been surveyed in this paper including some the parametric dependencies. Individual extraction methodologies typically will be more sensitive to specific parts of the underlying

measurement parameter space and some examples were discussed including those where correlation of uncertainties between the underlying parameters can have a significant effect. On strategy to minimize issues is to offset and misalign, where possible, the elevated uncertainty regions from the sensitive regions for extraction but this does require looking at more of the process.

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