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Presentation on FET Modeling for Power Amplifier Design

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Seminar: Gain Without Pain November 2000

Principles and Procedures for Successful Large-Signal Measurement-Based FET Modeling for Power Amplifier Design

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Abstract

Measurement-based active device modeling is an attractive approach to simulating circuits. This methodology is based on the comprehensive characterization of devices from a diverse range of processes and technologies. This section reviews the foundations of measurement-based modeling as well as highlighting some future trends, with respect to large-signal FETs. Specific topics include a discussion of different model types and their advantages. Also included are requirements for modeling I-V and Q-V constitutive relations for accurate intermodulation and ACPR simulations, as well as device characterization modeling issues. Recently-developed characterization methods, including pulsed I-V and pulsed S-parameter, are highlighted as well, including a discussion of their implications for dynamic electrothermal models with trapping effects.

Biography

Dr. David E. Root is presently the research and development project manager for computer aided engineering, modeling, and advanced characterization at the Agilent Microwave Technology Center in Santa Rosa, California. David has 14 years of experience in device modeling and has worked in the area of statistical large-signal modeling for circuit design with implications for IC manufacturability. Recent efforts include pulsed-bias and pulsed S-parameter device characterization and modeling.



This section of the workshop includes an overview of the major concepts and modern trends of measurement-based large-signal FET modeling as it applies to simulating power amplifiers. A key goal is to correlate modeling concepts and methods with the requirements for accurate simulation of power amplifiers

This section provides a framework for much of the information in the remaining sections of the workshop.



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The principles of semiconductor device physics govern the workings of an active device, such as the generic FET shown above. Semiconductor physics are based on the quantum theory of matter, electrodynamics, and non-equilibrium thermodynamics. These physical principles can be described by partial differential equations that require specific boundary conditions for their solution. Other relevant ingredients required for a first-principles quantitative understanding of the device include details about its physical geometry, material properties, surface chemistry, and manufacturing details.

About 50 years ago, William Shockley described the time-dependent terminal characteristics for a FET using partial differential equations. Shockley accomplished this based on simple physical assumptions including the field-independent mobility of free carriers in the semiconductor material and the gradual channel approximation, among others. This work yielded a set of dynamical equations (ordinary differential equations) and constitutive relations (explicit algebraic transfer functions) for the terminal current and the capacitances, as functions of the applied terminal voltages. These dynamical equations and constitutive relations take the form of the simple large-signal equivalent circuit on the right side of this slide. The equivalent circuit has a voltage dependent current source and a pair of two-terminal nonlinear capacitors.

The coefficients of the constitutive relations include the geometric parameters, such as the length and width of the gate, and the depth of the channel. Other key parameters are the doping density of the semiconductor, the dielectric constant, and the carrier mobility.

I find it remarkable that the complex behavior of a device such as a transistor can be described with ordinary differential equations that can be solved by nonlinear circuit simulators.



The closed-form expressions derived from the Shockley theory are appealing in their simplicity. However, the performance of modern GaAs FETs require additional physics. For example, velocity saturation, is a key phenomenon missing in Shockley's theory. As recently as 1974, the velocity saturation effect in GaAs FETs with gate-lengths of 1mm revealed significant gaps between measured and expected performance when predictions were based on the closed-form, expressions from the Shockley theory.

When physics are simplified to the point of closed-form expressions, the results are generally not accurate enough for circuit simulation.

Predictablility from first-principles physics often doesn't happen until *after the technology has become practical and introduced* to the marketplace.

Even when the relevant physics can be modeled, the physical parameters required by the physics may be unavailable. For example, while the physics for trapping effects may be modeled, surface state density for a particular device might be unknown or impractical to obtain.



Measurement-based active device modeling is an attractive approach to simulating circuits. This methodology is based on the comprehensive characterization of devices from a diverse range of processes and technologies. The most common and practical approach to measurement-based modeling is to characterize the device using a range of high-frequency measurements, and then mathematically transform the data into predictive device models that can then be used for large- and small-signal circuit design and simulation purposes.

However, many mathematical, physical, and measurement considerations must be incorporated to produce a framework that will yield successful measurement-based modeling. With this in mind, we'll now discuss some of the foundations of this subject as well as looking at some future trends.

Technological innovations continue to yield ever-faster devices that use less power and provide better performance. However, the rate of developments in device technology can easily outpace the availability of accurate device models. Even with advances like time-domain, harmonic balance, and envelope simulators, circuit designers may face complex, imperfect, and unfamiliar device behavior, long before the effects are precisely understood and modeled in terms of first-principles physics.



The dynamics of empirical models are assumed to be the same as the physically-based Shockley model. This means that the form of the differential equations as well as the topology of the equivalent circuit are the same for both models. However, in the empirical models, the functional form of the constitutive relations, I-V and C-V, are replaced by empirical, closed-form expressions. Lacking any direct physical significance, these empirical forms are chosen merely to fit the measured data, while preserving mathematical convenience. Also note that the form of Cgs remains consistent with simple physical considerations.

The Curtice Cubic polynomial model is one example of a simple, familiar, and successful empirical model. The polynomial is only defined for a finite range of V1, Vp \leq V1 \leq Vmax. This is indicated by the solid blue curve in the figure. But, if the cubic polynomial is defined for the entire bias range, as indicated by the dashed blue lines, it diverges at the negative and positive extremes for V1. This divergence is a consequence of using a polynomial that is unbounded at its argument's extremes.

A model's constitutive relation should be well-defined over its entire range. In this example, the solid blue curve can be extended, as indicated by the red lines, with continuous derivatives, as a constant (Id=0) for V1<Vp, and as a linear function of V for V>Vmax. Therefore, this constitutive relation is a piece-wise function and does not have continuous second derivatives at Vp or at Vmax.

The parameters of the cubic equation cannot be arbitrary. Rather they are constrained according to the need to have both Id, and its first derivative with respect to V1, equal to zero at V=Vp, as well as having fixed values for Imax and gm, respectively, at Vmax. Since the nonlinear transfer function is essentially determined by Vp and Imax, simply extracting coefficients from best fit data over a specified range can lead to a model that doesn't pinch off.

Also note that the forms for Cgs and Cgd are different, unlike the Shockley Model.



This slide illustrates the conventional method for extracting model parameters for simulation and optimization. This procedure is used to determine the unknown coefficients of the nonlinear constitutive relations of empirical models.

First, an initial guess is made for the parameter values for the circuit, which may be just a single device with a biasing network. Using these values, the circuit is simulated, and the results are compared with measurements. If the fit is good, the process ends. Otherwise, the model coefficient values are updated and simulation is repeated. The process continues until an acceptable level of agreement is achieved between measured and simulated results.

Ultimately, a precise fit may never be achieved. The fixed constitutive relations may not be flexible enough to be represented well by detailed measured characteristics of the device.

Also, parameter extraction is slow, since the circuit must be simulated many times. Final parameter values may be highly sensitive to the initial values chosen. The effort can get bogged down and never return a good set of parameter values. And optimization can stuck at a local minimum of the cost function.

Finally, any time a constitutive relation is modified, say to improve fit with the data, the corresponding extraction routine must to be updated as well.



The figure shows measured (X) and simulated I-V characteristics (lines) for a three-terminal vertical, power MOSFET, and a GaAs PHEMT. The same model simulates both very different devices extremely well. This is because of the procedure used to develop the model, which is essentially to measure the device, transform the data, tabulate the data, interpolate the discrete data (to define the constitutive relations as a continuous, smooth function), and then to scale the results. The constitutive relations of table-based models *are the interpolated data*.

If this were not the case, distinct physics and different closed-form constitutive relations would be needed to describe a MOSFET and PHEMT. In particular, the PHEMT exhibits a drain-voltage dependent pinch-off voltage, and a subtle "kink" in its I-V characteristics.

Table models can be both *accurate* and *general*—accurate, because the device's own data is used directly in the nonlinear model, and general, because the models are technology and process independent...to a great degree. Splines faithfully interpret the measured characteristics, however different they may be from one technology to the next.

Many table-based models can be *directly constructed* from data, without the need for optimization-based parameter extraction. The generation of a table-based model, once the data is collected, can be carried out several orders of magnitude faster than extracting the parameters of an empirical model.



The table in this slide shows a typical, but simplified, model datafile. The model datafile replaces the parameters of an empirical model.

Independent variables Vgs and Vds, shown in black, are discrete operating points that were measured for the device. For each operating point, the measured value of the terminal currents was tabulated. Note that the data is not uniformly distributed. The table also provides geometrical information about the device—here, the gate width is given.

The model's constitutive relations are scaled versions of the interpolated values from the measured data. The scaling rules follow the same basic geometrical rules that standard empirical and physical models use.

The interpolation method for large-signal models must be smooth. For example, piece-wise cubic splines can be used, but as we'll find out later, these can cause problems. The interpolation must be smooth to facilitate processing by nonlinear equation solvers in most circuit simulators, which are usually based on Newton's Method.

Effects such as parasitic resistances, intrinsic versus extrinsic voltage difference, and other details are not treated here.



The upper left-hand plot shows each data point specified by the operating point in Vgs-Vds voltage space, as measured by an automated, adaptive, data acquisition system [22] used to characterize the device. The boundary of this domain is determined by a set of compliances, or limitations imposed for device safety while it is characterized.

Limitations for maximum forward DC gate current and reverse breakdown current are indicated by black and blue lines. Note that the breakdown locus is characterized by Vgd=constant, for Vgs < -2V. This is conventional breakdown. At Vgs > -2V, the breakdown boundary changes slope, indicative of another breakdown mechanism becoming dominant.

The next compliance is determined by the maximum DC power that the device can safely dissipate. This locus is the red curve. It is important to note that the actual shape of the boundary depends both on the compliances and the device-specific I-V characteristics. Another device with different characteristics would exhibit a data domain with a different shape. The adaptive data acquisition system interacts with the device for each data point, to achieve measurements within the safe range for characterization.

The remaining three plots show the DC drain current, gate current, and the magnitude of S21 at a single frequency, measured at each operating point.

Since the model is based on data and interpolation, it is necessary to have access to lots of good data. Given the volume of data that must be collected, an automated acquisition system is required.



This slide shows an example of some issues to consider with spline interpolation for tablebased models. The graphs indicate measured I-V points for a diode (triangles), spline-fit interpolation (wavy lines on lower graph), and well-behaved model lines. While at full scale in the upper graph, they all look consistent. At a finer scale (near I=0), we see that cubic splines do in fact go through each data point, but oscillate in between points!

One conclusion we can reach here is that interpolation of noisy data leads to a poor model. It also points to the need for constraints on splines (e.g. monotonicity). In practice, it is hard to achieve this in more than one dimension (e.g. FETs). The diode model used in this example is based on an analytic function where data is noisy.



Charge storage modeling is critical for simulating S-parameters versus frequency. The charge model also contributes significantly to device distortion under moderate to high amplitude drive conditions, even more so as signal frequencies increase. Charge also plays a significant role in ACPR simulations, something we'll look at later in this presentation.

Several issues make charge modeling more difficult and subtle than modeling I-V relations. Unlike terminal currents, charge is not directly measurable with conventional DC and S-parameter data. What we can measure are admittance parameters (a simple transformation from S-parameters). Also, using simple linear algebra, we obtain the "measured" small-signal Equivalent Circuit Parameters (ECP) or elements, parameterized by the operating point. These measurements can be directly compared to the large-signal model by simulating in small-signal conditions.

The blue sequence shows how to go from large-signal model to parameterized small-signal admittances or small-signal model capacitances. There red sequence shows how data can be converted into "measured" admittances or "measured" ECPs. At these points, the model can be compared to data (black double arrows).

The measurement-based approach starts with data at the beginning of the red sequence. The red sequence is completed and then the magenta arrows are followed to recover, if possible, the large-signal model nonlinear ODEs. This is not possible, in general. There are many nonlinear ODEs which, when linearized, will yield the exact same bias-dependent linear equivalent circuit elements [1],[3], yet give different simulation results under large-signal conditions. Moreover, even if the large-signal model ODEs (or equivalent circuit model) are formulated in terms of Q-V relations, it is an ill-posed mathematical problem to recover this charge from the parameterized measured capacitances, unless the latter respect, within measurement error, the constraints of terminal charge conservation (see next few overheads).



Any C(V) relation (1-D) for a two-terminal capacitance function can be integrated to calculate Q(V) (even without a constant doping profile).

The model admittance (common source) is calculated by linearizing the gate current equation at each operating point, and then transforming to the frequency domain.

Monotonic C(V) relations (1-D physics) are assumed.

- VDS=VGS-VGD
- Plot CGS versus VGS and VDS
- Plot CGD versus VGS and VDS



This slide shows measured capacitance data versus bias. First we can see that Cgs and Cgd are both dependent on the values for both Vgs and Vds. Cgs and Cgd do not behave like two-terminal capacitors, which depend only on the voltage across them. Therefore, the data requires that we reinterpret the equivalent circuit

It is not possible for the pair of two-terminal capacitances to ever reproduce measurements, independent of their constitutive relations.

What we must do is find large-signal equations which, when linearized reproduce the measured linear equivalent circuit elements, parameterized by operating point.



The spectrum of Ig given by equation A contains a DC component that is frequency (fundamental) dependent (for the proof see [1], [2]).

Charge-based constitutive relations can never cause this problem.

But unlike the 1-D case, multi-terminal capacitance functions need not admit the existence of a charge.

The terminal charge conservation is NOT KCL. Terminal charge conservation relates the bias-dependence of capacitance functions attached to a given node. Terminal charge conservation is a modeling constraint that may or may not be imposed for simulation. On the other hand, KCL must always be satisfied for circuit equations.



This slide looks at the charge calculation from the data. If and only if the line-integral of the data is path-independent [21], then the data is consistent with the model constraint of terminal charge conservation, and a unique model charge function (up to a constant) will fit both measured Cgs and Cgd over bias. The model also exactly fits the measured ImY11 and ImY12 over bias.

Even a path-dependent charge integral of data will produce a large-signal model with capacitances that conserve terminal charge. However, these capacitances do not fit the bias-dependence of the measured capacitance exactly.



Most fail to consider drain charge/capacitance issues.

In the model, Cm represents a transcapacitance.

At least one transcapacitance must appear in a three-node lumped equivalent circuit. The transcapacitance can be placed in any branch, with simple changes in the functional form of the capacitance functions and transcapacitance function with bias.

Cm can cause excessive gain at very high frequencies.



Four "measurable" capacitances are modeled by two independent charge functions, Qg and Qd.

FET capacitance data is consistent with the modeling constraint of charge conservation at the gate. There is some discrepancy between FET linear data and the terminal charge conservation constraint at the drain.

Energy conservation constrain the model terminal charge constitutive relations in precisely the same way (mathematically) that terminal charge conservation constrains the capacitance functions attached to a node.

It is not always possible to recover a conserved energy by two path-independent lineintegrations of measured Cij matrix elements.

There are consequences of not enforcing conservation laws and there are trade-offs made when implementing them.



These figures [8] compare load-pull RF measurements with model predictions. Gain and insertion phase are illustrated on the left, while power added efficiency (PAE) is shown on the left.

Model A is a simple junction capacitance model (one-dimensional) equivalent to that for Cgs of the Curtice Model.

Model B is an analytic charge model that includes the Vds and Vgs bias-dependence (twodimensional model) of measured Cgs and Cgd, fitted to the data.

Model C is a table-based charge model with splines, which directly constructs the charge functions by integrating the measured parameterized capacitance functions.

All three models have the same closed-form analytical current models. Any differences between simulations can be attributed entirely to the charge-model.

It is clear that the table-based charge model is superior to the other two.



These figures [8] compare load-pull RF measurements with model predictions. Third order intermodulation products are shown at left, while adjacent channel power ratio (ACPR) for an NADC compliant Pi/4 DQPSK stimulus is illustrated on the right.

Model A is a simple junction capacitance model (one-dimensional) equivalent to that for Cgs of the Curtice Model.

Model B is a analytic charge model which includes the Vds and Vgs bias-dependence (twodimensional model) of measured Cgs and Cgd, fitted to the data.

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It is clear that the table-based charge model is superior to the other two.

It is also clear that charge modeling is very important for simulating, accurately, quantities of interest for Power Amplifier Design.





Convergence depends on model equations & simulator algorithms

- Correct evaluation of model equation partial derivatives critical
- Choose constitutive relations with bounded second partial derivatives (for Newton solvers) *Infinitely differentiable*, global analytic functions are best
- Correct transfer function asymptotes help bring simulator to reality
- Choose charge functions such that the capacitances are bounded away from 0 (keeps minimum step size finite in transient)
- Limit exponential and other highly nonlinear functions (to limit overflow and reduce iterations)
- Use minimum conductances to avoid singular equations



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DC characteristics are not equivalent to RF performance over bias! FETs are "non quasi-static".

Ids in the equivalent circuit has no explicit frequency dependence.

Non quasi-static effects on currents (dispersion) should be modeled by dynamic thermal effects and trapping effects [17].



In this slide, the scales on data graphs have been removed to maintain customer confidentiality.

At the top and middle, examples A and B are simulations at an operating point symmetrically located with respect to neighboring data, in Vgs space. A cubic spline has a zero of its third derivative there!

Small-amplitude simulations are strongly related to derivatives at the operating point.

Example C a the bottom shows a simulation done slightly away from the above point.

The results from Spline-based models can depend on interpolant rather than data for smallamplitude signals (for some spline schemes, at some biases).

When signal voltage swings become comparable to the distance between data points, model accuracy improves.



This slide compares analytic transfer functions versus splines for a PHEMT common source amplifier, biased at Vds=2V. The single points represent measured data, solid lines represent simulated data using an analytic model, while the dashed lines show simulated data using a spline-based model.

The model is based on infinitely differentiable I-V relations, extracted from pulsed bias measurements with built-in dynamic self-heating.

The analytic model is generally superior to simple spline-based models, at low signal levels.

High-order derivatives of analytic functions are a necessary, but insufficient, condition for good intermodulation simulation.



This slide compares analytic transfer functions versus splines for a PHEMT common source amplifier, with a different bias at Vds=3V. Again, the single points represent measured data, solid lines represent simulated data using an analytic model, while the dashed lines show simulated data using a spline-based model.



Electro-thermal measurement models are based on the assumption that the electro-thermal constitutive relation is linear for temperature deviation about ambient. This assumption allows for 2-D interpolation of voltage-dependent gamma, and requires only two different temperatures to extract gamma.

We could use a 3-D interpolation, or fitting of I(Vds,Vgs,T), which would be slower to simulate and require more data.

We can also use isothermal or non-isothermal data.

Thermal circuit (ODE) can be generalized to more closely represent distributed thermal effects (diffusion) and to couple, thermally, with the environment and other devices. [20]

It should be noted that careful implementation is required for robust convergence.



This slide shows the dynamic trajectory of a table-based model constructed using DC and Sparameter data, for a Class A amplifier.

By using pulsed measurements, the device characterization is extended into regions where the device operates, extrapolating beyond the domain where DC and small-signal measurements can be made.

Pulsed measurements also allow separation of thermal and trap effects. Roughly speaking, these measurements can be considered "iso-dynamic".

An arbitrary pulsed semiconductor parameter analyzer (APSPA) [13] can experiment over all regions of quiescent bias.

The procedure consists of holding the device under test at a constant bias for 10 ms, and then measuring I-Vs with <500 ns pulses is pseudo-random order.



The graph in this slide shows pulsed data versus DC data. Pulsed characteristics, represented as solid lines, are shown for two different quiescent bias points (indicated by circles). DC characteristics are shown as triangles for a PHEMT.

Pulsed data covers a much greater range of bias-space than DC data.

Pulsed data does not reveal any negative differential conductance at high power (DC) dissipation I-V points.

The shape of the device's pulse response is highly dependent on the quiescent point. This remains true even after temperature normalization. The remaining difference is attributed to "trapping" effects.

Models built from this data, with dynamic self-heating and gate-lag, have been shown to fit the rich time-dependent large-signal device phenomena well [14] [6].











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Revised: March 27, 2008		

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