MICROWAVE

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2004

Friday, June 11, 2004

Advances and New Directions in Device Modeling and Design Optimization for Microwave CAD

Organizers:
John W. Bandler, Bandler Corporation
Q.J. Zhang, Carleton University

Sponsors:
MTT-1: Computer Aided Design
MTT-15: Microwave Field Theory

WFD

IEEE ★ MTT-S

INTERNATIONAL
MICROWAVE
SYMPOSIUM WORKSHOP
Notes & Short Courses

FORT WORTH, TEXAS

WFD: Advances and New Directions in Device Modeling and Design Optimization for Microwave CAD

Date & Time: Friday, June 11; 8:00 AM to 5:00 PM

Location: Fort Worth Convention Center, Room 204B

Topics & Speakers:

- Device and Component Modeling and Optimization Exploiting Space Mapping and Space-Mapping Based Surrogate Modeling, **John Bandler**, Bandler Corporation
- Taking Optimization to the Next Level – The Complete Integration of Electromagnetic /Circuit/System Design, **Zoltan J. Cendes**, Ansoft Corp.
- Advances in Compact Models and Measurement-Based Models, **Steve Chen** and **David Root**, Agilent Techologies
- Introduction and Application of Cortical Theory to RF Amplifier Modeling and Compensation, **Paul Draxler**, Qualcomm
- Measurement-Based Frequency-Domain Nonlinear Component Modeling, **Jeffery A. Jargon**, **K.C. Gupta**, and **Donald C. DeGroot**, NIST and University of Colorado
- New Directions in Addressing Complexity of High-Speed Circuits and Interconnects, **Michel Nakhla**, Carleton University
- Accurate and Efficient Analysis of Large Spiral Inductors with Thick Metal and Narrow Gaps Using Space Mapping, **James C. Rautio**, Sonnet Software
- Electromagnetics-based Design Through Inverse Space Mapping Techniques, **José E. Rayas-Sánchez**, ITESO
- Coarse EM Modeling of LTCC RF Circuits and Its Applications to Optimization Design, **Ke-Li Wu** and **Jie Wang**, The Chinese University of Hong Kong
- Neural Network-Based Device Modeling and Design Optimization, **Q. J. Zhang**, Carleton University

Organizers: **Q.J. Zhang**, Carleton University
John W. Bandler, Bandler Corporation

Sponsors: MTT-1: Computer Aided Design
MTT-15: Microwave Field Theory

Extraordinary advances continue in modeling, optimization and statistical design for microwave CAD. Developments in EM-based modeling capabilities, mixed linear /nonlinear, EM field/circuit simulation, space mapping and knowledge-based artificial neural network (ANN) technology are creating new opportunities for microwave CAD with higher levels of design automation. These technologies address increased complexity of VLSI, RF and microwave circuits to fulfill the industrial demand for faster design cycle and reducing time to market for electronic products. Competition within the fast growing EDA market lead to incorporation of such technologies in future releases of commercial software.

An objective of this workshop is a tutorial review of the state of the art and a presentation of implementable methodologies and software. It will be substantially physically and electromagnetically oriented. It will highlight recent advances in ANN and space mapping for modeling and design tasks in RF and microwave CAD. Initiatives in integration of ANN capabilities into statistical design, behavioral modeling, measurement standards and computational electromagnetics are being made. Space mapping optimizations are being performed, for example, by linking full-wave EM simulations with empirical circuit-theory based simulations, or devices under test with suitable simulation surrogates. Exploitation of properly managed space-mapped (surrogate) models promises high efficiency in engineering design optimization practice, as well as in enhancing device models for yield-driven design.

We blend methodological aspects of wide applicability, design procedures currently applied in research and development centers, and well-known and widely available full-wave tools for design purposes. The workshop will bring together the foremost practitioners in these fields including microwave component designers, software developers and academic innovators. They will address designers' needs for effective tools for optimal designs, including yield optimization, exploiting accurate, physically-based device and component models, and consider the challenge of real life optimization, i.e., to produce in a short time, with limited resources, a design both competitive and innovative.

Device and Component Modeling and Optimization Exploiting Space Mapping and Space-Mapping Based Surrogate Modeling

J.W. Bandler, Q.S. Cheng and D.M. Hailu

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McMaster University

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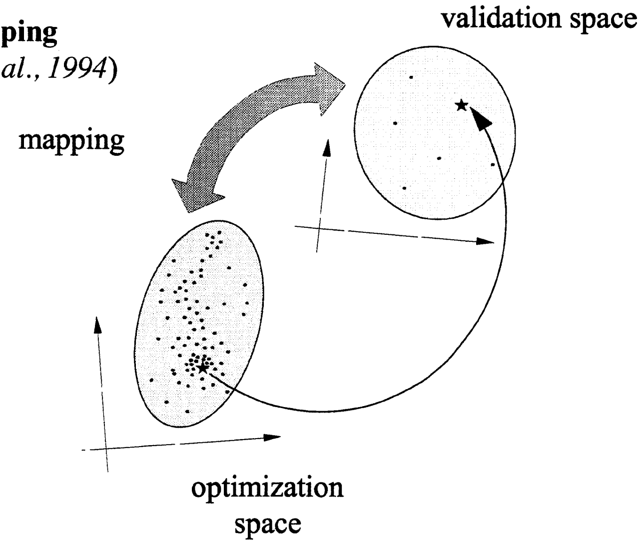
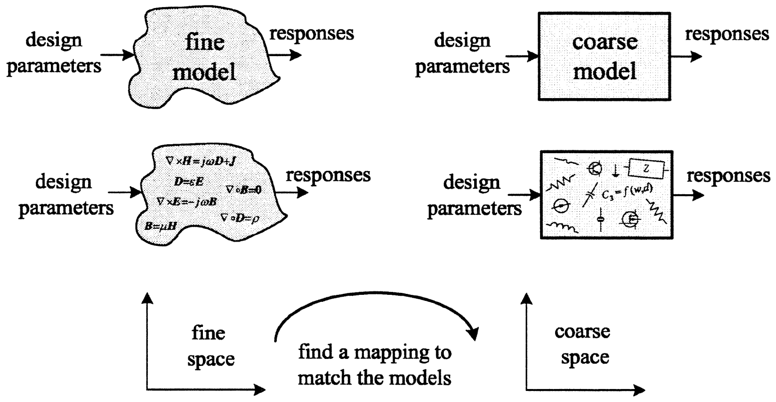
presented at

WFD WORKSHOP ON ADVANCES AND NEW DIRECTIONS IN DEVICE MODELING AND DESIGN OPTIMIZATION USING MICROWAVE CAD
2004 IEEE MTT-S International Microwave Symposium, Fort Worth, TX, June 11, 2004

Space Mapping
(*Bandler et al., 1994*)



Linking Companion Coarse (Empirical) and Fine (EM) Models



design parameters → fine model → responses

design parameters → coarse model → responses

design parameters → responses

design parameters → responses

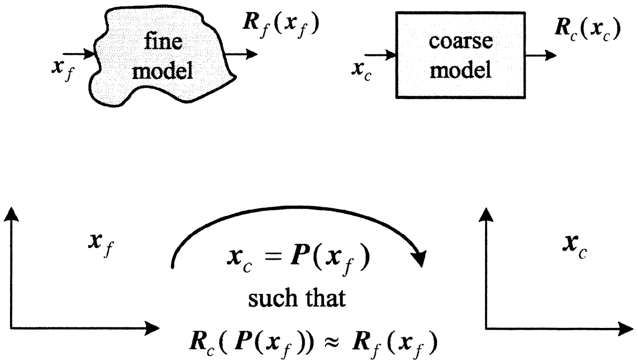
fine space

find a mapping to match the models

coarse space

The Space Mapping Concept

(*Bandler et al., 1994-*)



x_f → fine model → $R_f(x_f)$

x_c → coarse model → $R_c(x_c)$

x_f

$x_c = P(x_f)$

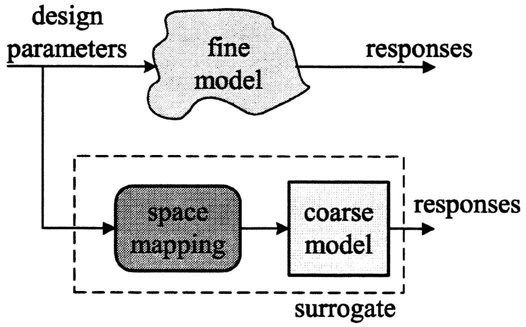
such that

$R_c(P(x_f)) \approx R_f(x_f)$

x_c

Explicit Space Mapping Concept

(*Bandler et al., 1994-*)



Space Mapping: a Glossary of Terms

Space Mapping	transformation, link, adjustment, correction, shift (in parameters or responses)
Coarse Model	simplification or convenient representation, companion to the fine model, auxiliary representation, cheap model
Fine Model	accurate representation of system considered, device under test, component to be optimized, expensive model

Space Mapping: a Glossary of Terms

Surrogate	model, approximation or representation to be used, or to act, in place of, or as a substitute for, the system under consideration mapped or enhanced coarse model
Surrogate Model	alternative expression for Surrogate
Target Response	response the fine model should achieve, (usually) optimal response of a coarse model, enhanced coarse model, or surrogate

Space Mapping: a Glossary of Terms

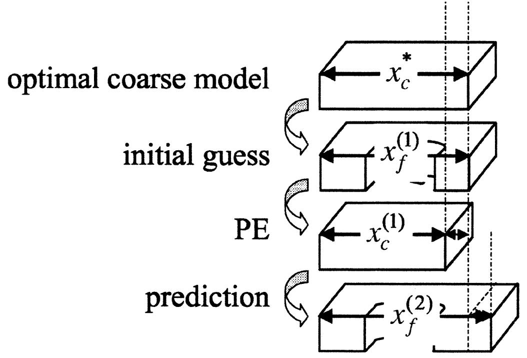
Companion	coarse
Low Fidelity/ Resolution	coarse
High Fidelity/ Resolution	fine
Empirical	coarse
Simplified Physics	coarse
Physics-based	coarse or fine
Device under Test	fine
Electromagnetic	fine or coarse
Simulation	fine or coarse
Computational	fine or coarse

Space Mapping: a Glossary of Terms

Neuro	implies use of artificial neural networks
Implicit Space Mapping	space mapping when the mapping is not obvious
Not Space Mapping	(usually) space mapping when not acknowledged
Parameter Transformation	space mapping
Predistortion	?

Aggressive Space Mapping Practice—Cheese Cutting Problem

(*Bandler et al., 2002*)



$$x_f^{(1)} = x_c^*$$

$$x_f^{(2)} = x_f^{(1)} + P^{-1}(x_c^* - x_c^{(1)})$$

$$x_f^{(2)} = x_f^{(1)} + x_c^* - x_c^{(1)}$$

Interpretation of Space Mapping Optimization

the original optimization problem

$$\boldsymbol{x}^* \triangleq \arg \min_{\boldsymbol{x}} U(\boldsymbol{R}(\boldsymbol{x}))$$

consider $\boldsymbol{R}_c(\boldsymbol{P}(\boldsymbol{x}_f))$ as an "enhanced" coarse model or "surrogate," then

$$\overline{\boldsymbol{x}}_f = \arg\min_{\boldsymbol{x}_f} U(\boldsymbol{R}_c(\boldsymbol{P}(\boldsymbol{x}_f))$$

is equivalent to

$$\boldsymbol{f}(\boldsymbol{x}_f) \triangleq \boldsymbol{P}(\boldsymbol{x}_f) - \boldsymbol{x}_c^*, \quad \boldsymbol{f} \rightarrow \boldsymbol{0}$$

Aggressive Space Mapping Approach

(*Bandler et al., 1995*)

iteratively solves the nonlinear system

$$\boldsymbol{f}(\boldsymbol{x}_f) = \boldsymbol{0}$$

the quasi-Newton step $\boldsymbol{h}^{(j)}$ in the fine space is given by

$$\boldsymbol{B}^{(j)}\boldsymbol{h}^{(j)} = -\boldsymbol{f}^{(j)}$$

the next iterate

$$\boldsymbol{x}_f^{(j+1)} = \boldsymbol{x}_f^{(j)} + \boldsymbol{h}^{(j)}$$

Aggressive Space Mapping Approach (continued)

Broyden-like updates (*Bandler et al., 1995*)

$$\boldsymbol{B}^{(j+1)} = \boldsymbol{B}^{(j)} + \frac{\boldsymbol{f}^{(j+1)} - \boldsymbol{f}^{(j)} - \boldsymbol{B}^{(j)}\boldsymbol{h}^{(j)}}{\boldsymbol{h}^{(j)^T}\boldsymbol{h}^{(j)}}\boldsymbol{h}^{(j)^T}$$

Jacobian based updates (*Bandler et al., 1999, 2002*)

$$\boldsymbol{B} = (\boldsymbol{J}_c^T\boldsymbol{J}_c)^{-1}\boldsymbol{J}_c^T\boldsymbol{J}_f$$

$$\boldsymbol{E} = \boldsymbol{J}_f - \boldsymbol{J}_c\boldsymbol{B}$$

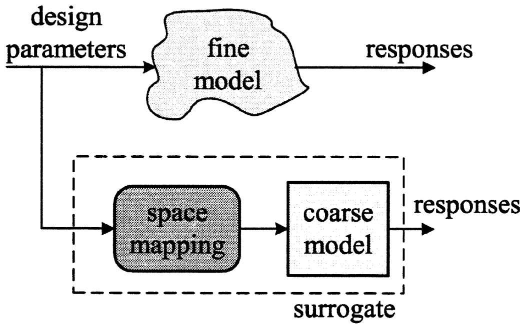
constrained update (*Bakr et al., 2000*)

$$\Delta\boldsymbol{B} = \boldsymbol{B} - \boldsymbol{I}$$

$$\boldsymbol{B} = \arg\min_{\boldsymbol{B}} \left\| [\underline{\boldsymbol{e}_1^T \cdots \boldsymbol{e}_n^T} \quad \eta\underline{\Delta\boldsymbol{b}_1^T \cdots \eta\Delta\boldsymbol{b}_n^T}]^T \right\|_2^2$$

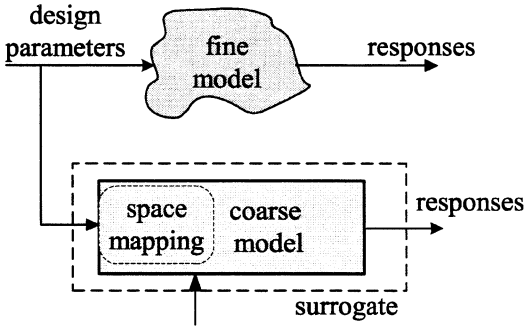
Explicit Space Mapping Concept

(*Bandler et al., 1994-*)



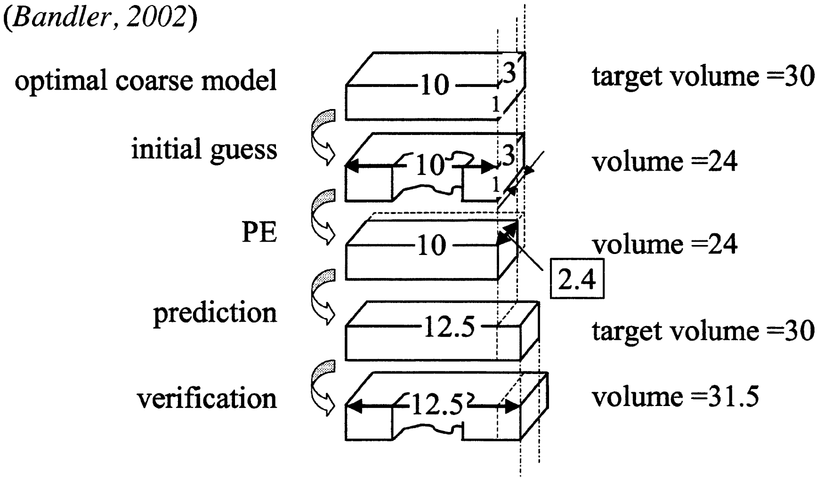
Implicit Space Mapping Concept

(*Bandler et al., 2004*)



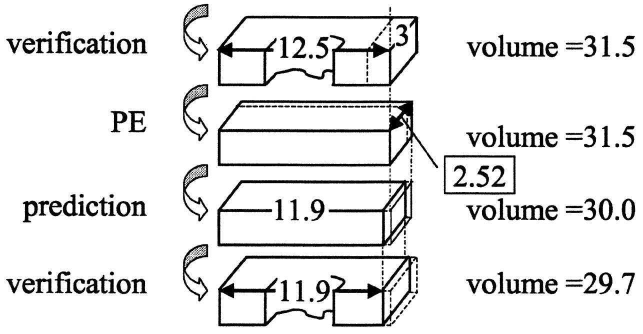
Implicit Space Mapping Practice—Cheese Cutting Problem

(*Bandler, 2002*)



Implicit Space Mapping Practice—Cheese Cutting Problem

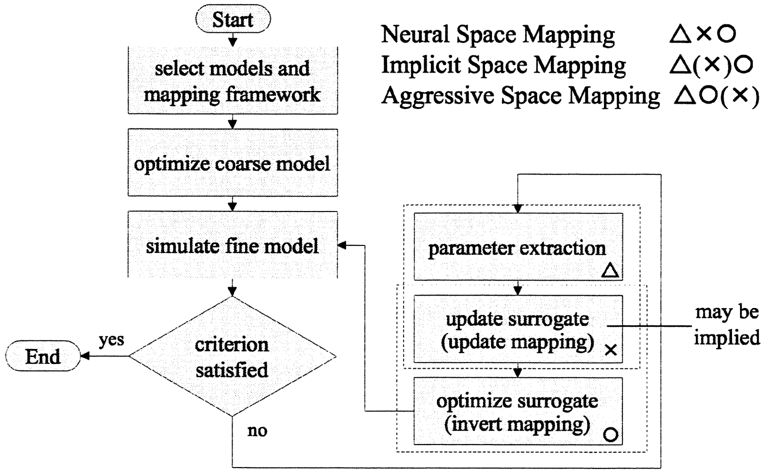
(*Bandler, 2002*)



∴ error = (30–29.7)/30×100%
=1%

Space Mapping Framework

(*Bandler et al., 2004*)

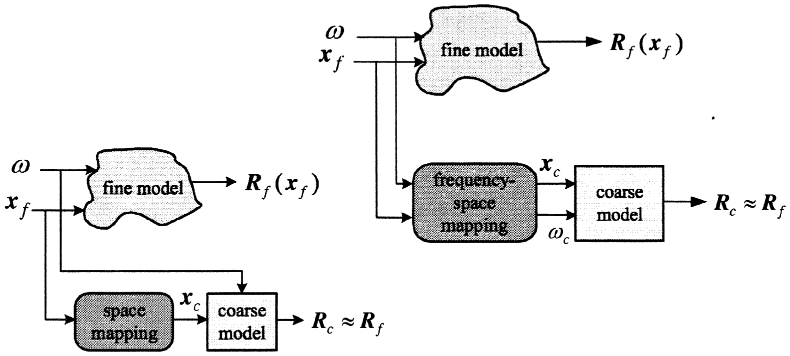


Neural Space Mapping	△×○
Implicit Space Mapping	△(×)○
Aggressive Space Mapping	△○(×)

Model Enhancement—the SM Tableau Approach

(*Bandler et al., 2001*)

already used in the RF industry for new library models (*Snel, 2001*)



Space Mapping Implementation and Applications 1

RF and microwave implementation (*Bandler et al., 1994-2004*)

civil engineering structural design (*Leary et al., 2000*)

automobile crashworthiness design (*Redhe et al., 2001-2002*)

generating microwave neural models (*Devabhaktuni et al., 2002*)

combline filter design (*Swanson and Wenzel, 2001*)

microwave filter design (*Harscher, et al., 2002, 2003*)

CAD of integrated passive elements on PCBs (*Draxler, 2002*)

Space Mapping Implementation and Applications 2

CAD technique for microstrip filter design
(*Ye and Mansour, 1997*)

SM models (model enhancement) for RF components (*Snel, 2001*)

multilayer microwave circuits (LTCC) (*Pavio et al., 2002*)

cellular power amplifier output matching circuit (*Lobeek, 2002*)

multilevel ASM strategy applied to filter optimization
(*Safavi-Naeini et al., 2002*)

coupled resonator filter (*Pelz, 2002*)

Space Mapping Implementation and Applications 3

LTCC RF passive circuit design (*Wu et al., 2002-2004*)

waveguide filter design (*Steyn et al., 2001*)

inductively coupled filters (*Soto et al., 2000*)

magnetic systems (*Choi et al., 2001*)

Implicit Space Mapping optimization with preassigned parameters
(*Bandler et al., 2001-2004*)

Output Space Mapping optimization (*Bandler et al., 2003-2004*)

Space Mapping Implementation and Applications 4

EM-based optimization of microwave oscillators
(*Rizzoli et al., 2003*)

circuit level, neuro-SM modeling of nonlinear devices
(*Zhang et al., 2003-2004*)

optimization of dielectric resonator filters and multiplexers using ASM (*Ismail et al., 2003-2004*)

waveguide filter design (*Morro et al., 2003*)

optimal control of partial differential equations
(*Hintermueller and Vicente, 2003*)

Space Mapping Implementation and Applications 5

modeling and simulation of photonic devices
(*Feng and Huang, 2003*)

design of comb filters using implicit SM (*Gentili et al., 2003*)

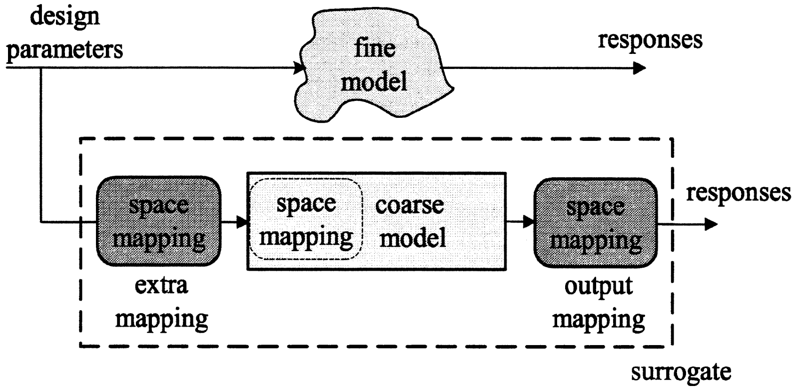
optimization of antireflection coatings in photonic devices
(*Feng et al., 2003*)

time-domain design, CMOS drivers, using linear inverse and neuro inverse SM (*Rayas-Sánchez, 2004*)

Space Mapping Interpolating Surrogates (SMIS) for highly optimized EM-based design (*Bandler et al., 2004*)

Implicit, Extra and Output Space Mappings

(*Bandler et al., 2003*)



Output Space Mapping

Response Residual Space Mapping (RRSM) surrogate definition

$$\boldsymbol{R}_s = \boldsymbol{O}(\boldsymbol{R}_c) \triangleq \boldsymbol{R}_c(\boldsymbol{x}_c, \boldsymbol{x}) + \boldsymbol{diag}\{\lambda_1, \lambda_2, \cdots, \lambda_m\} \Delta \boldsymbol{R}$$

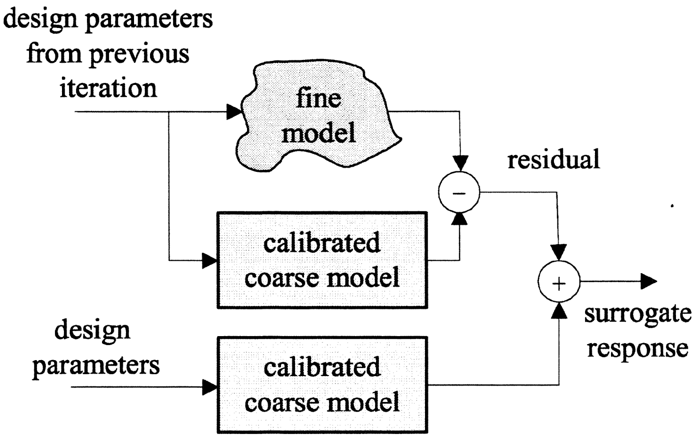
where

$$\Delta \boldsymbol{R} \triangleq \boldsymbol{R}_f(\boldsymbol{x}_f) - \boldsymbol{R}_c(\boldsymbol{x}_c^{*(j)}, \boldsymbol{x})$$

from experience

$$0 \le \lambda_i \le 1$$

Illustration of the RRSM Surrogate



ADS Schematic Design Framework for Space Mapping

Step 1 set up the coarse model in the ADS schematic

Step 2 optimize the coarse model using ADS

Step 3 copy and paste the parameters into the parameterized fine model (Agilent Momentum, HFSS/Empipe3D, or Sonnet's ***em***) *Comment*: the Momentum fine model can be generated using the *Generate/Update Layout* command

Step 4 simulate the fine model and save the responses in Touchstone format (Agilent Momentum, HFSS, or Sonnet's ***em***) or Dataset (Momentum)

ADS Schematic Design Framework for Space Mapping

Step 5 if stopping criteria are satisfied, stop

Step 6 parameter extraction
- (a) import the responses to the ADS schematic using S*n*P component under *Data Items*
- (b) set up ADS (calibrated) coarse model or output residual SM surrogate to match the S*n*P component
- (c) run ADS optimization to perform parameter extraction

Comment extract the coarse model design parameters to implement explicit SM (original or aggressive) or the preassigned parameters for implicit SM

ADS Schematic Design Framework for Space Mapping

Step 7 predict the next fine model solution by
- (a) explicit SM: transfer extracted parameters to Matlab (or other tool); predict through the original or aggressive SM algorithm, or,
- (b) implicit SM: reoptimize w.r.t. design parameters the calibrated coarse model (surrogate), and/or,
- (c) RRSM: reoptimize surrogate w.r.t. design parameters (calibrated coarse model plus response residual)

Step 8 update the fine model design and go to Step 4

ADS Schematic Design Framework for Space Mapping

the ADS schematic framework works in an interactive way

three ADS schematic designs:

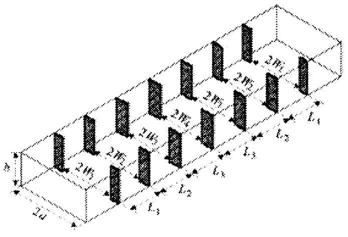
- (a) coarse model optimization design
- (b) parameter extraction design
- (c) surrogate (re)optimization design

only parameter values are updated in each iteration

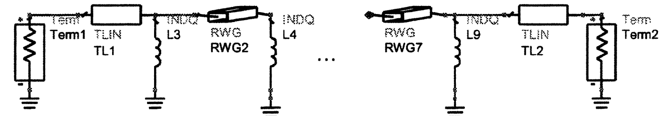
currently, the fine model is Agilent Momentum, HFSS, or Sonnet's ***em***

H-plane Filter Design

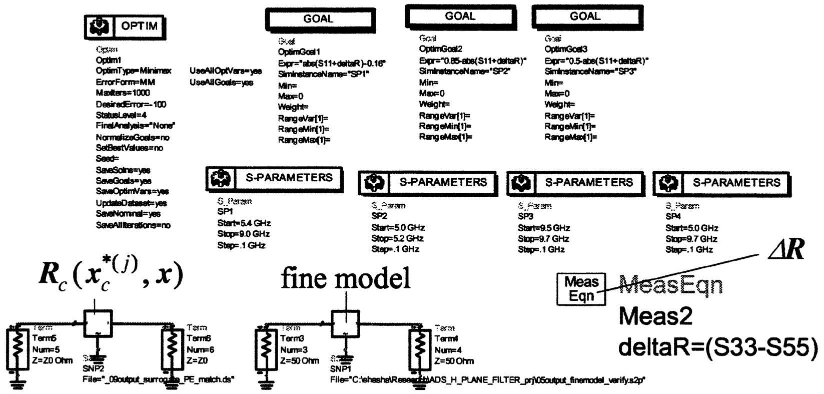
(*Young et. al., 1963, Bakr et al., 1999, Bandler et al., 2004*)



design specifications
$|S_{11}| \leq 0.16,\ 5.4 \leq \omega \leq 9.0\text{GHz}$
$|S_{11}| \geq 0.85,\ \omega \leq 5.2\text{GHz}$
$|S_{11}| \geq 0.5,\ \omega \geq 9.5\text{GHz}$



H-plane Filter: ADS Setup For RRSM Surrogate Optimization



Optim
Optim1
OptimType=Minimax
ErrorForm=MM
MaxIters=1000
DesiredError=-100
StatusLevel=4
FinalAnalysis="None"
NormalizeGoals=no
SetBestValues=no
Seed=
SaveSolns=yes
SaveGoals=yes
SaveOptimVars=yes
UpdateDataset=yes
SaveNominal=yes
SaveAllIterations=no
UseAllOptVars=yes
UseAllGoals=yes

Goal
OptimGoal1
Expr="abs(S11+deltaR)-0.16"
SimInstanceName="SP1"
Min=
Max=0
Weight=
RangeVar[1]=
RangeMin[1]=
RangeMax[1]=

Goal
OptimGoal2
Expr="0.85-abs(S11+deltaR)"
SimInstanceName="SP2"
Min=
Max=0
Weight=
RangeVar[1]=
RangeMin[1]=
RangeMax[1]=

Goal
OptimGoal3
Expr="0.5-abs(S11+deltaR)"
SimInstanceName="SP3"
Min=
Max=0
Weight=
RangeVar[1]=
RangeMin[1]=
RangeMax[1]=

S_Param
SP1
Start=5.4 GHz
Stop=9.0 GHz
Step=.1 GHz

S_Param
SP2
Start=5.0 GHz
Stop=5.2 GHz
Step=.1 GHz

S_Param
SP3
Start=9.5 GHz
Stop=9.7 GHz
Step=.1 GHz

S_Param
SP4
Start=5.0 GHz
Stop=9.7 GHz
Step=.1 GHz

MeasEqn
Meas2
deltaR=(S33-S55)

Term5
Num=5
Z=Z0 Ohm

Term6
Num=6
Z=Z0

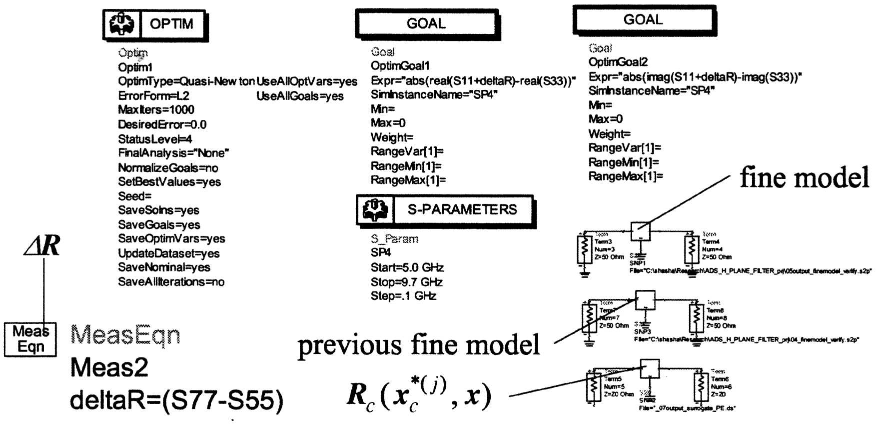
SNP2
File="_09output_surrogate_PE_match.ds"

Term3
Num=3
Z=50 Ohm

Term4
Num=4
Z=50 Ohm

SNP1
File="C:\shasha\Research\ADS_H_PLANE_FILTER_prj\05output_finemodel_verify.s2p"

H-plane Filter: ADS Setup For RRSM Parameter Extraction



Optim
Optim1
OptimType=Quasi-Newton
ErrorForm=L2
MaxIters=1000
DesiredError=0.0
StatusLevel=4
FinalAnalysis="None"
NormalizeGoals=no
SetBestValues=yes
Seed=
SaveSolns=yes
SaveGoals=yes
SaveOptimVars=yes
UpdateDataset=yes
SaveNominal=yes
SaveAllIterations=no
UseAllOptVars=yes
UseAllGoals=yes

Goal
OptimGoal1
Expr="abs(real(S11+deltaR)-real(S33))"
SimInstanceName="SP4"
Min=
Max=0
Weight=
RangeVar[1]=
RangeMin[1]=
RangeMax[1]=

Goal
OptimGoal2
Expr="abs(imag(S11+deltaR)-imag(S33))"
SimInstanceName="SP4"
Min=
Max=0
Weight=
RangeVar[1]=
RangeMin[1]=
RangeMax[1]=

S_Param
SP4
Start=5.0 GHz
Stop=9.7 GHz
Step=.1 GHz

MeasEqn
Meas2
deltaR=(S77-S55)

Term3
Num=3
Z=50 Ohm

Term4
Num=4
Z=50 Ohm

SNP1
File="C:\shasha\Research\ADS_H_PLANE_FILTER_prj\05output_finemodel_verify.s2p"

Term7
Num=7
Z=50 Ohm

Term8
Num=8
Z=50 Ohm

SNP3
File="C:\shasha\Research\ADS_H_PLANE_FILTER_prj\04_finemodel_verify.s2p"

Term5
Num=5
Z=Z0 Ohm

Term6
Num=6
Z=Z0

SNP2
File="_07output_surrogate_PE.ds"

H-plane Filter: Optimization Steps

two iterations of implicit SM drive the design close to optimal

then one implicit SM plus RRSM iteration using weighting parameters $\lambda_i = 0.5$, $i = 1\ldots m$

then, with the full residual added, a second implicit SM plus RRSM iteration

Comment $\lambda_i \leq 1$ because ADS has difficulty reoptimizing the surrogate with the full residual added

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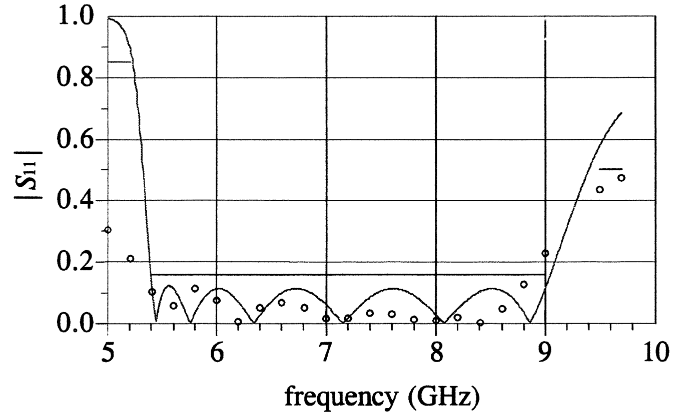
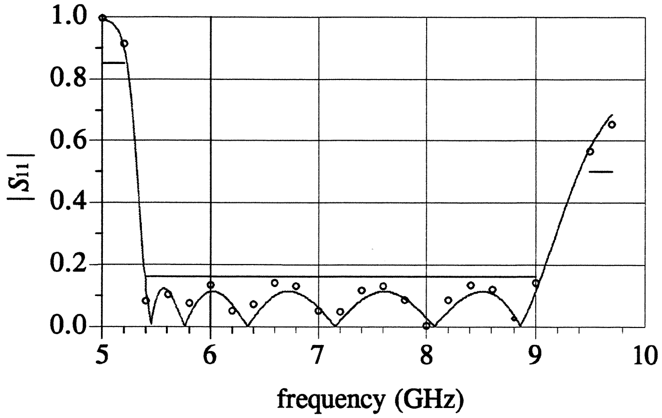
H-plane Filter Optimization

optimal coarse model response (—)
fine model response at initial solution (○)



H-plane Filter Optimization

optimal coarse model response (—)
fine model response (○) reached via RRSM after 4 iterations



Conclusions

glossary of terms for Space Mapping

Space Mapping optimization: original (1993)

Aggressive Space Mapping optimization: Broyden-based (1995-), Implicit Space Mapping (2001-), Output Space Mapping (2003-)

Response Residual Space Mapping (RRSM) modeling technique

RRSM implemented entirely in the ADS framework

interesting Space Mapping implementations and applications

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J.W. Bandler, Q.S. Cheng, N.K. Nikolova and M.A. Ismail, “Implicit space mapping EM-based modeling and design using preassigned parameters,” *IEEE Trans. Microwave Theory Tech.*, vol. 52, 2004, pp. 378–385.

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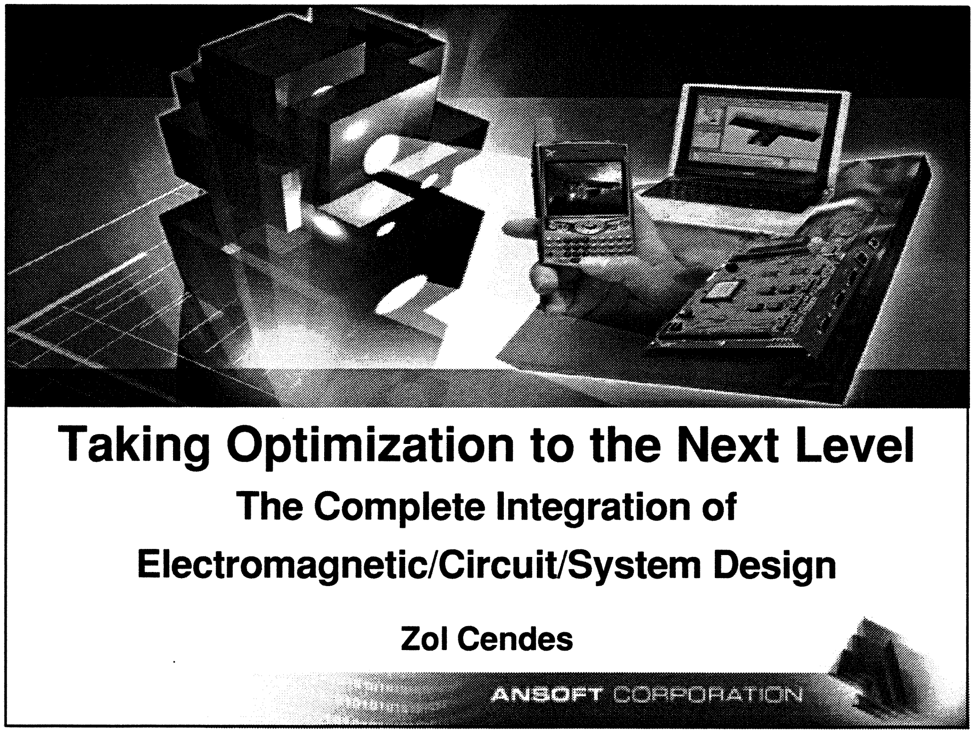
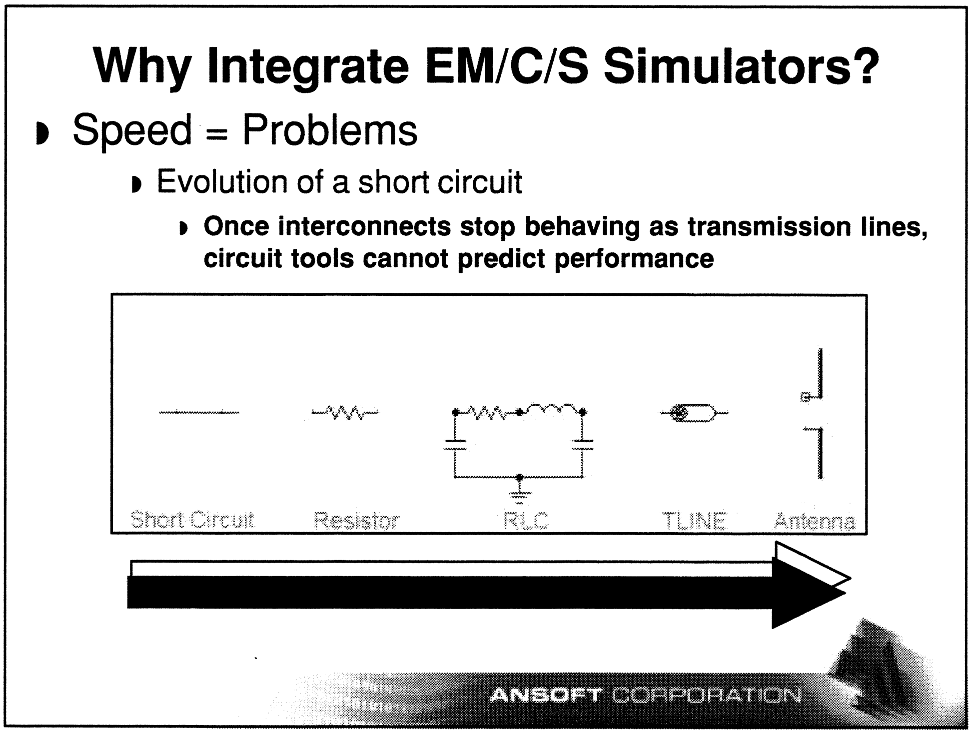
J.W. Bandler, Q.S. Cheng, S.A. Dakroury, D.M. Hailu, K. Madsen, A.S. Mohamed and F. Pedersen, “Space mapping interpolating surrogates for highly optimized EM-based design of microwave devices,” *IEEE MTT-S Int. Microwave Symp. Digest* (Fort Worth, TX, June 2004).



Taking Optimization to the Next Level

The Complete Integration of Electromagnetic/Circuit/System Design

Zol Cendes



Why Integrate EM/C/S Simulators?

- Speed = Problems
  - Evolution of a short circuit
    - **Once interconnects stop behaving as transmission lines, circuit tools cannot predict performance**

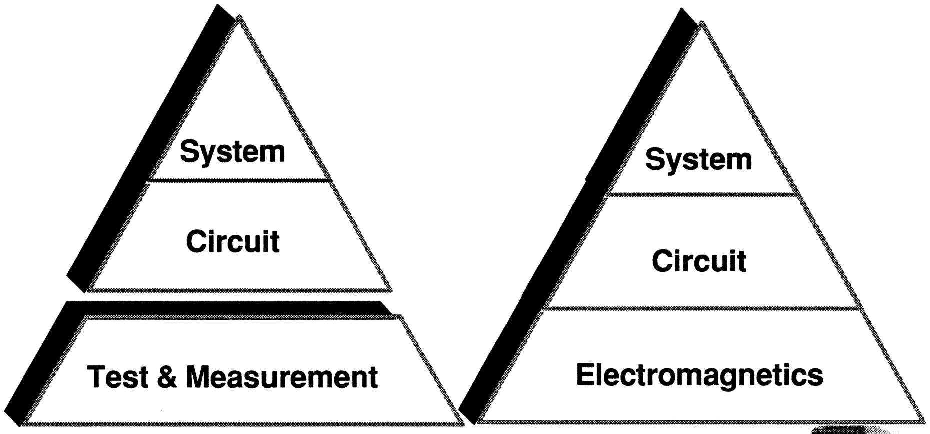
The Electromagnetic Opportunity

20th Century

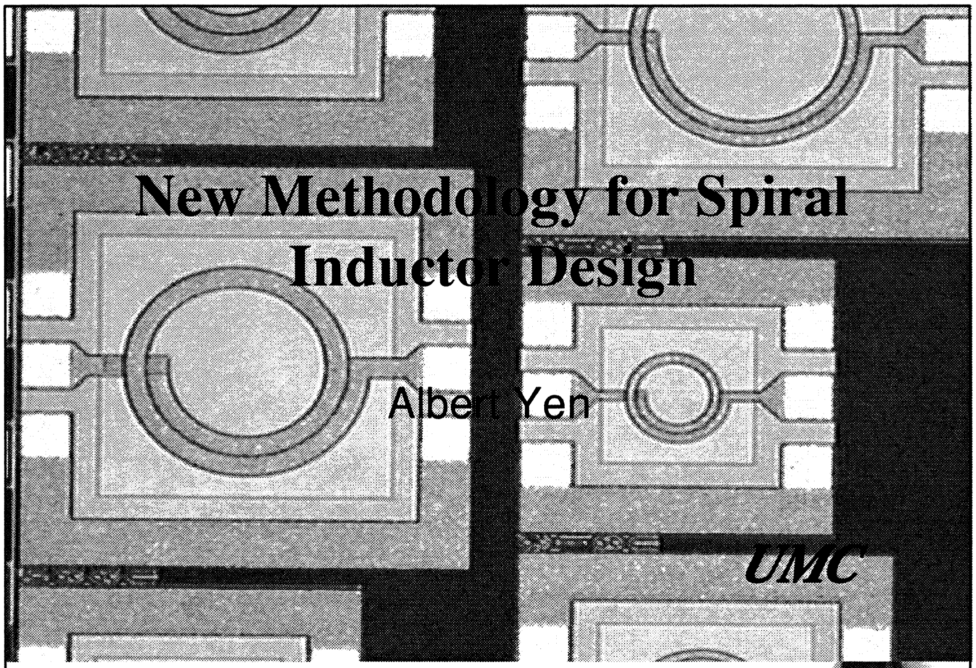
Use Software and T&M

21st Century

Full Software Solution


System
Circuit
Test & Measurement

System
Circuit
Electromagnetics



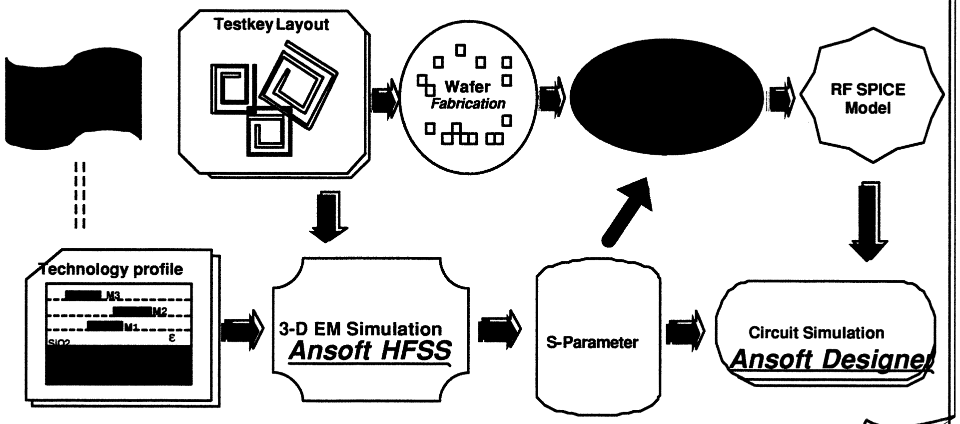
New Methodology for Spiral Inductor Design

Albert Yen

UMC

UMC's New Approach

Work Flow



Testkey Layout

Wafer *Fabrication*

RF SPICE Model

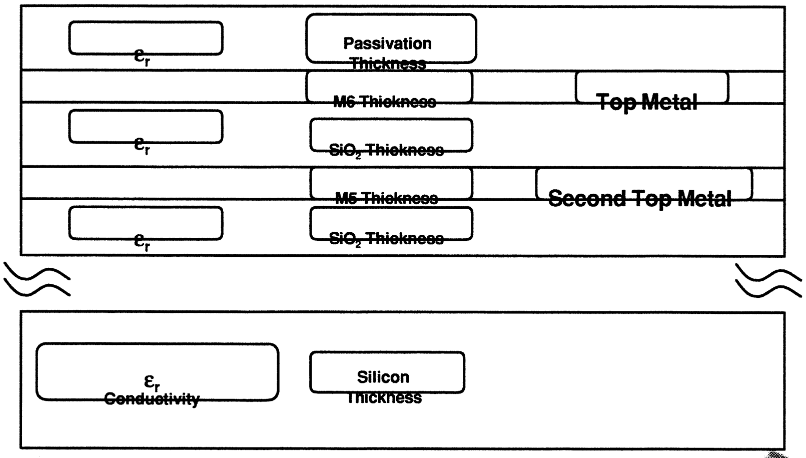
Technology profile

3-D EM Simulation
Ansoft HFSS

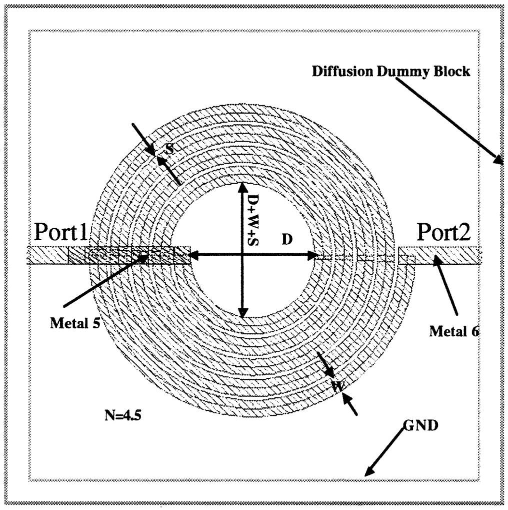
S-Parameter

Circuit Simulation
Ansoft Designer

UMC's New Approach
The Technology File



Ideal Circular Inductor



Geometrical Parameters

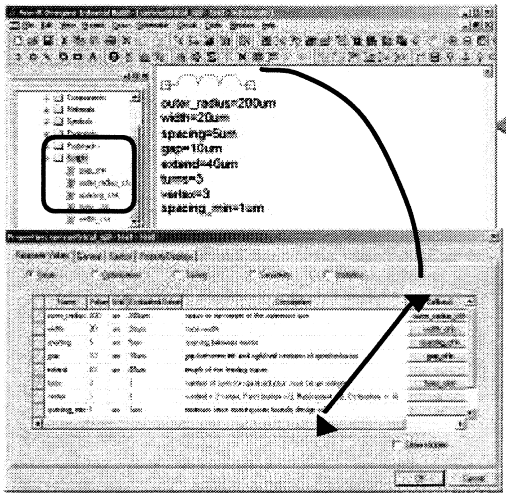
Diameter

Number of Turn

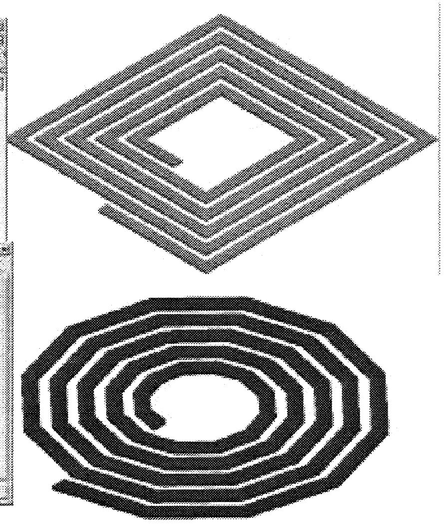
Width of Trace

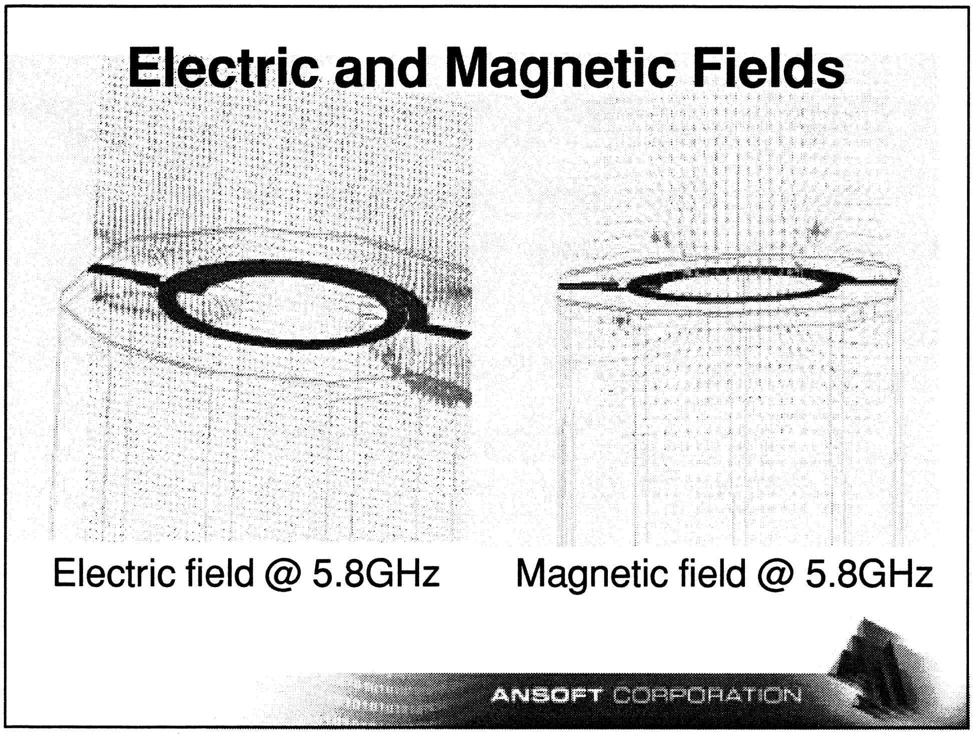
Space between Trace

User Defined Primitives



outer_radius=200um
width=20um
spacing=5um
gap=10um
extend=40um
turns=3
vertex=3
spacing_min=1um



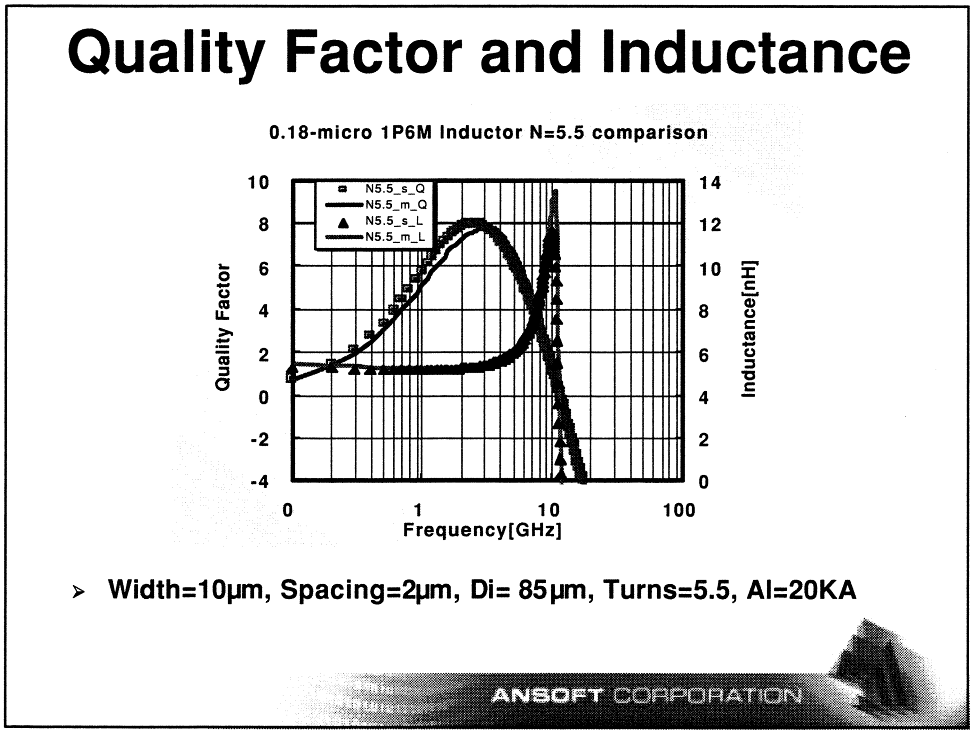


Electric and Magnetic Fields

Electric field @ 5.8GHz

Magnetic field @ 5.8GHz

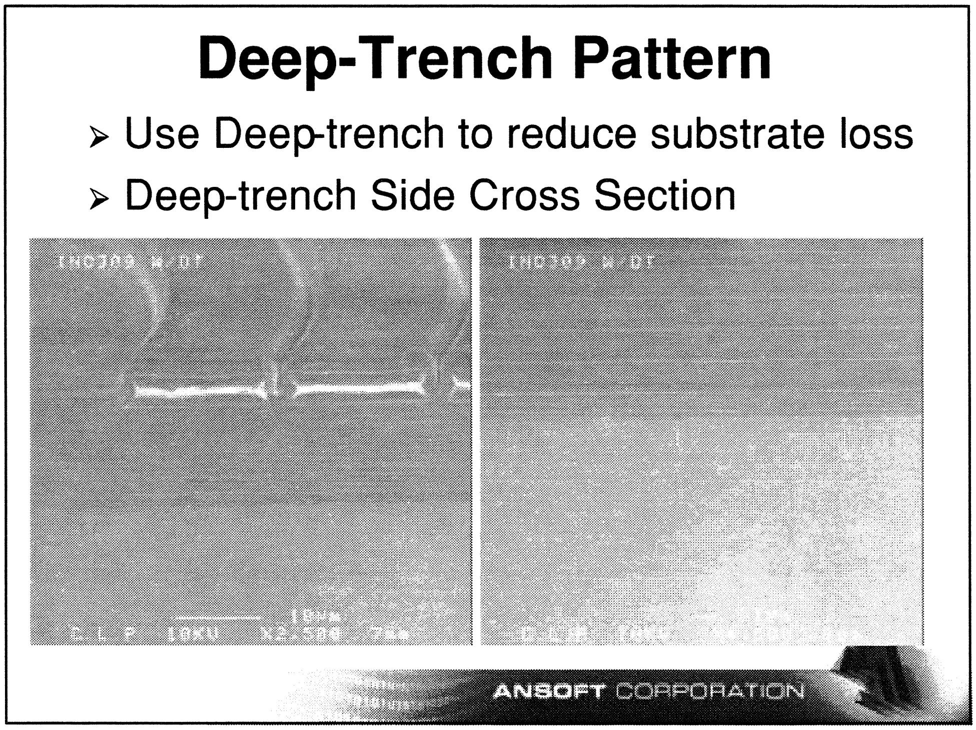
ANSOFT CORPORATION



Quality Factor and Inductance

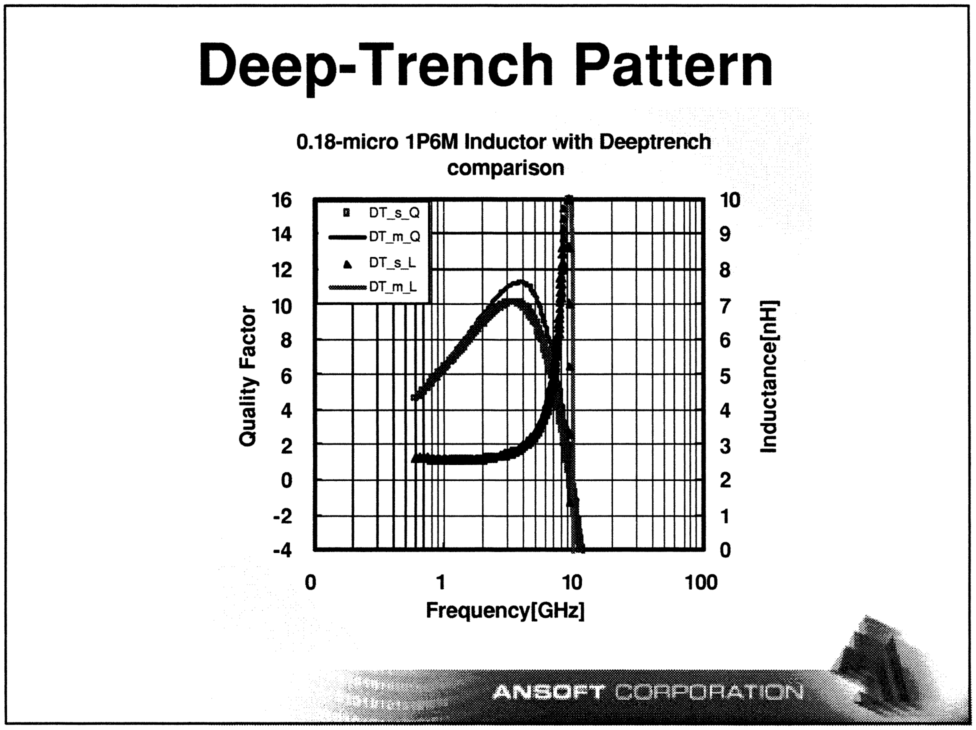
- Width=10μm, Spacing=2μm, Di= 85μm, Turns=5.5, Al=20KA

ANSOFT CORPORATION



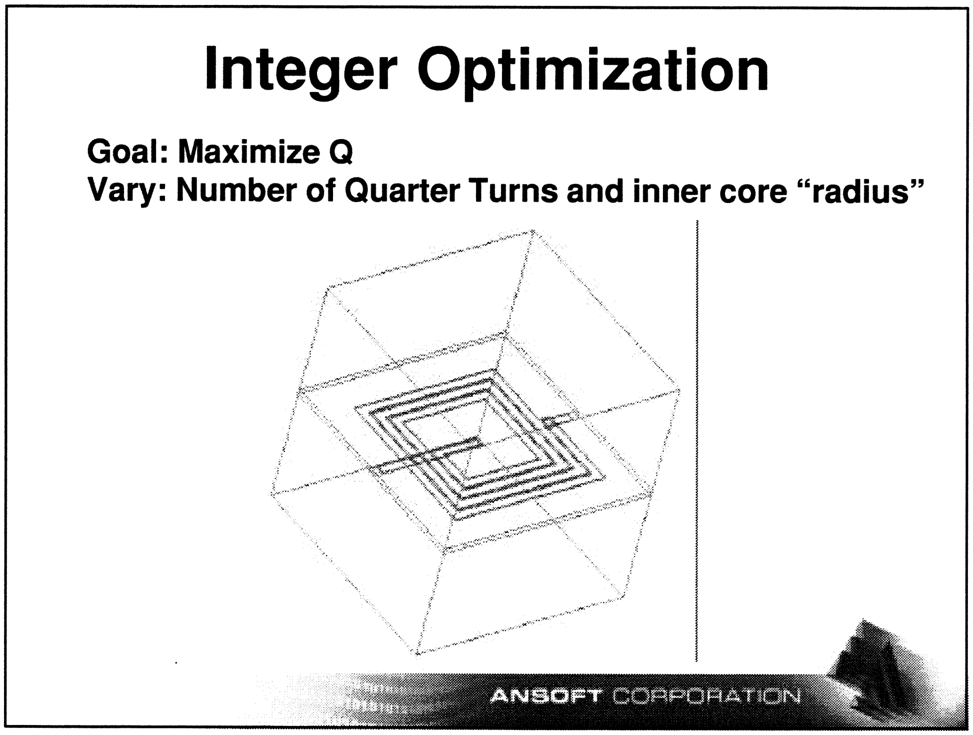
Deep-Trench Pattern

- Use Deep-trench to reduce substrate loss
- Deep-trench Side Cross Section



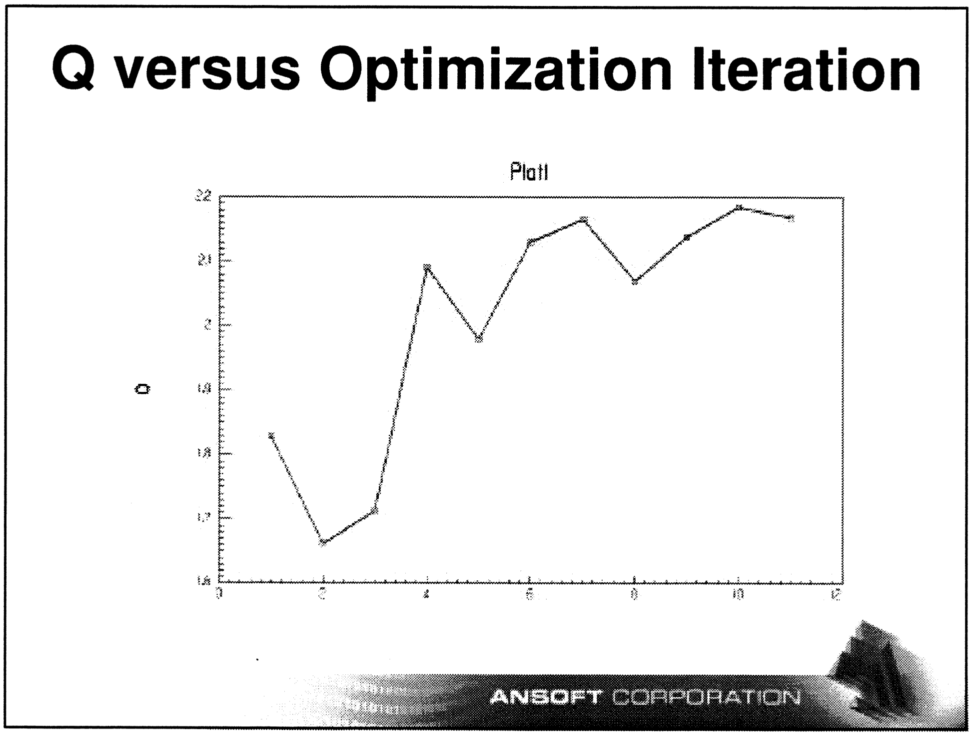
Deep-Trench Pattern

0.18-micro 1P6M Inductor with Deeptrench comparison



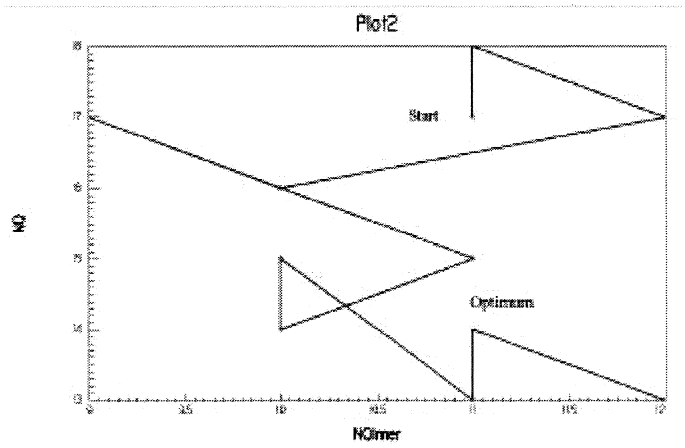
Integer Optimization

Goal: Maximize Q
Vary: Number of Quarter Turns and inner core "radius"

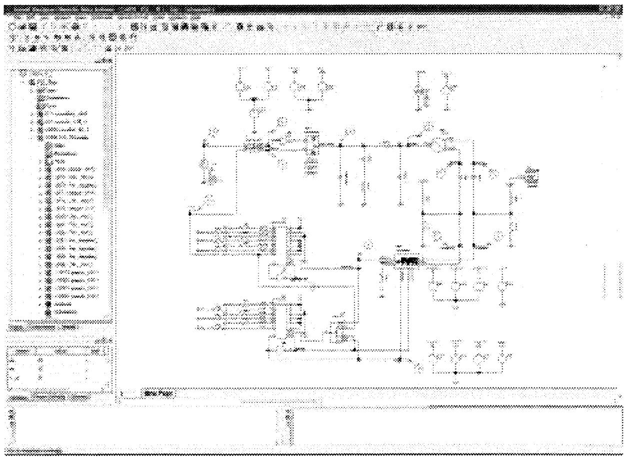


Q versus Optimization Iteration

Integer Optimization

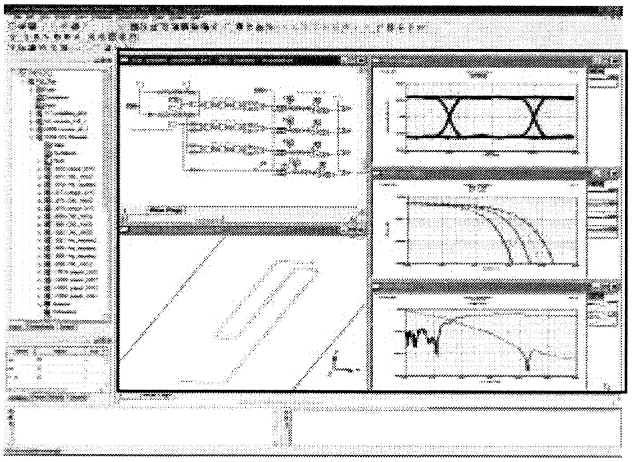


Complete EM/C/S Integration



Circuits

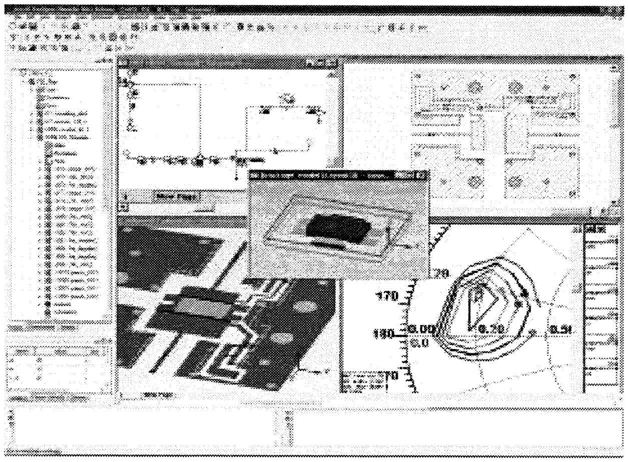
Complete EM/C/S Integration



Circuits

Systems

Complete EM/C/S Integration

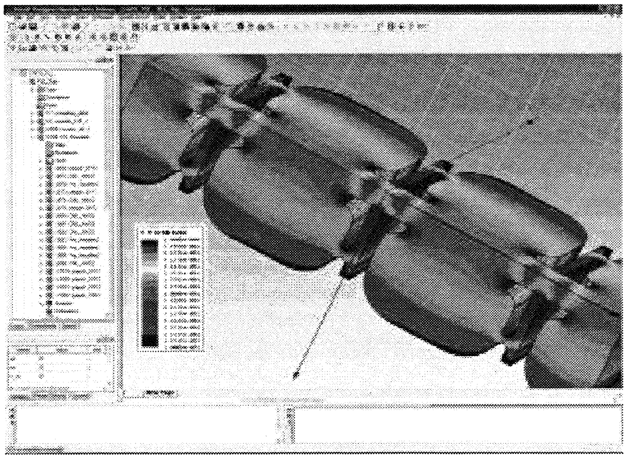


Circuits

Systems

Planar EM

Complete EM/C/S Integration



Circuits

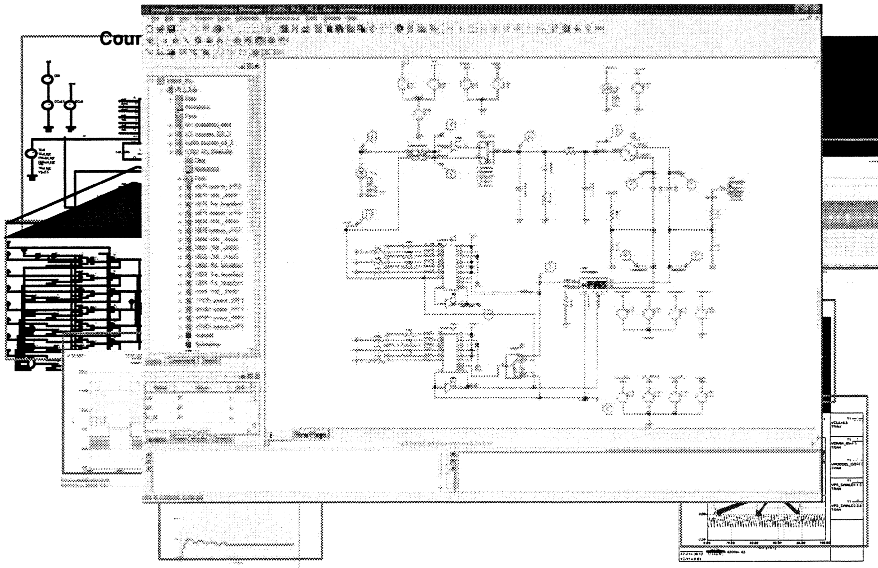
Systems

Planar EM

3D EM

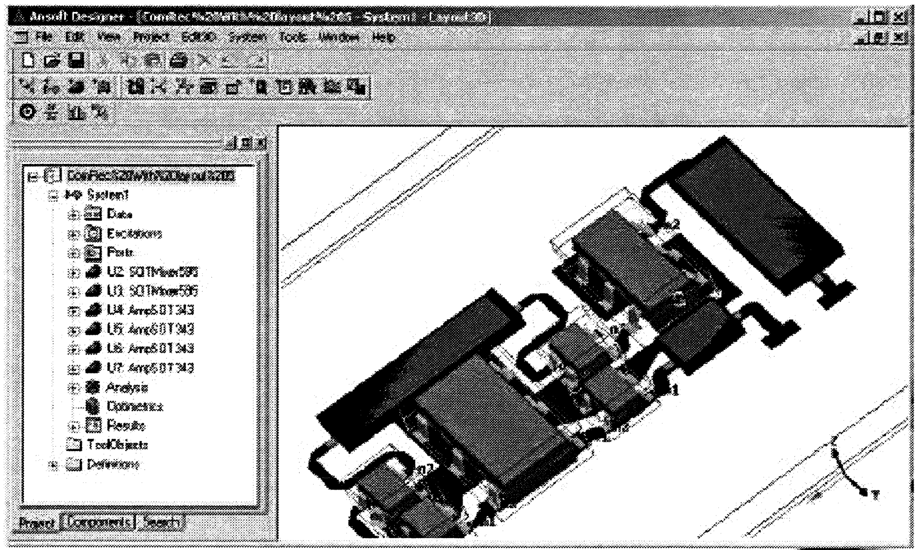
NEXXIM™

1.8 GHz CMOS PLL

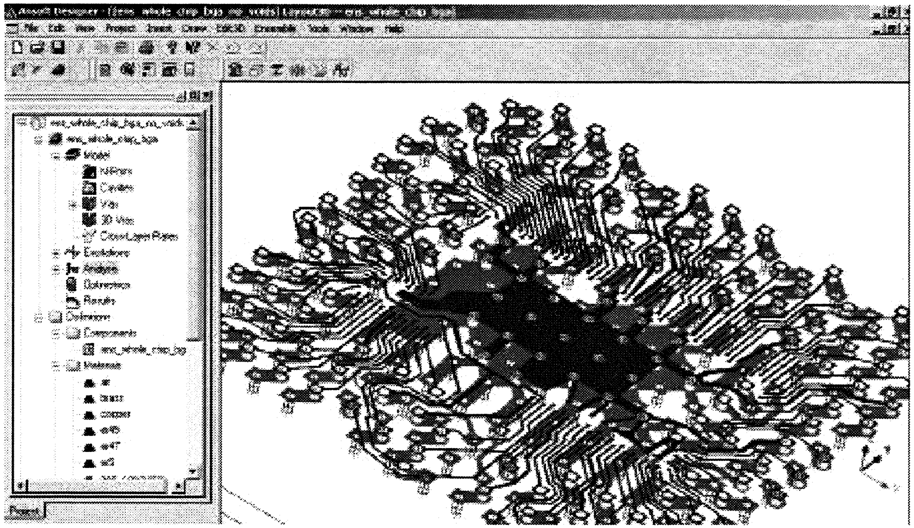


[redacted]

“Planar” or 2.5D EM

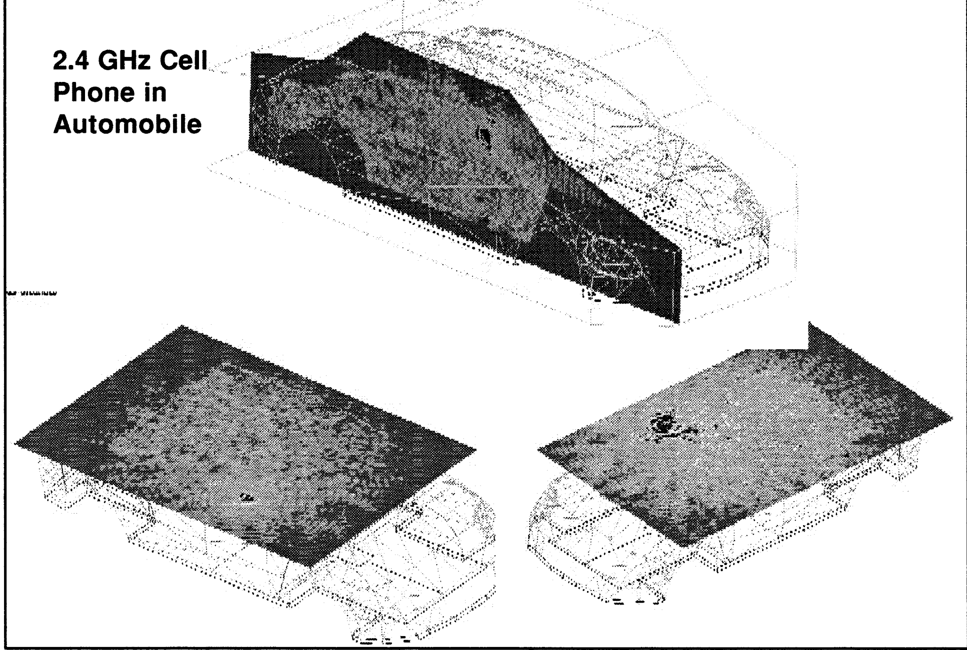


Planar EM SVD Fast Solver

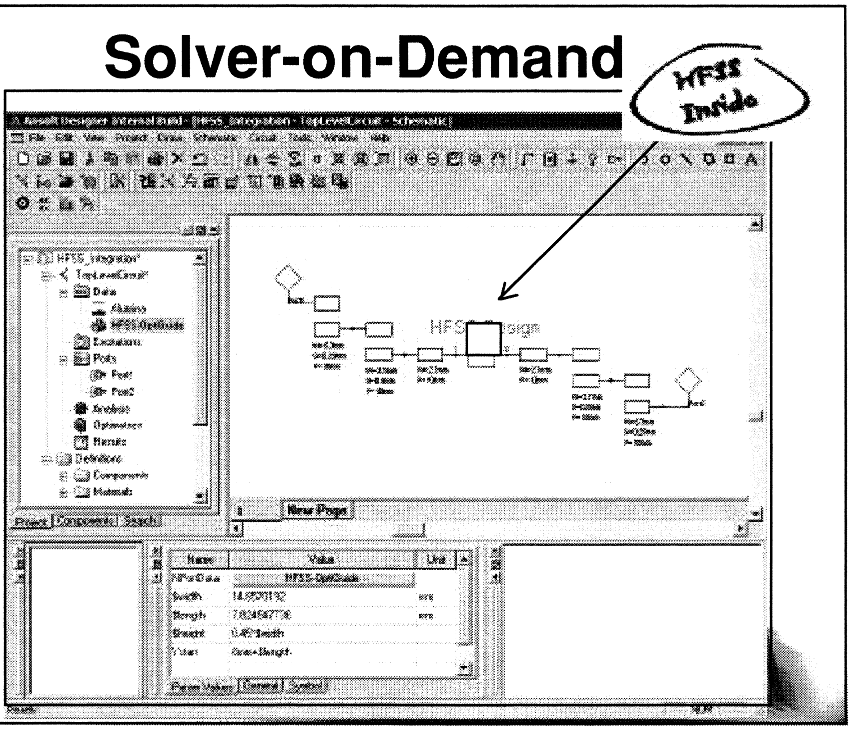


3D EM FEM Fast Solver

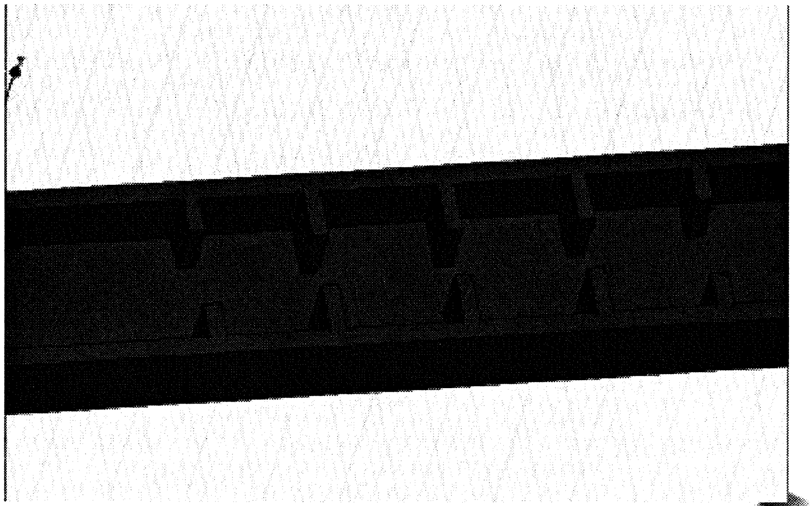
2.4 GHz Cell Phone in Automobile



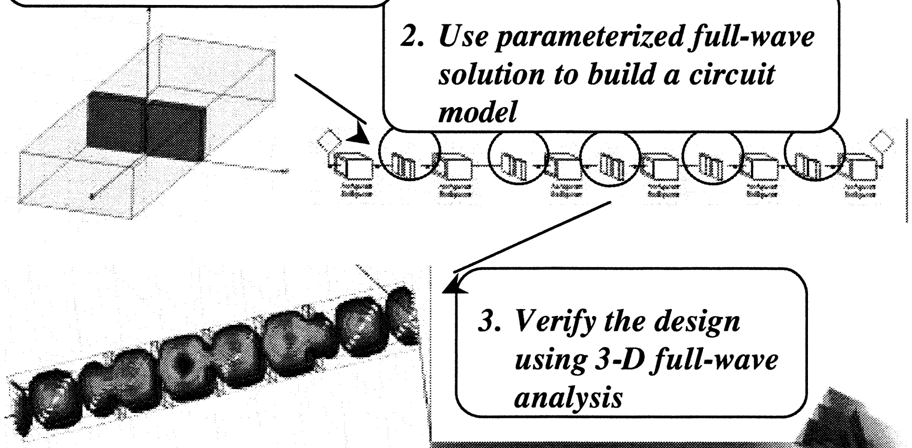
Solver-on-Demand



4-Pole WG Iris Filter



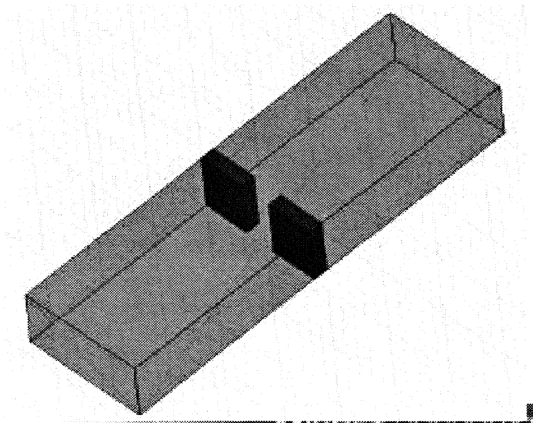
Design Flow



1. *Solve a parameterized model in HFSS for the iris*
2. *Use parameterized full-wave solution to build a circuit model*
3. *Verify the design using 3-D full-wave analysis*

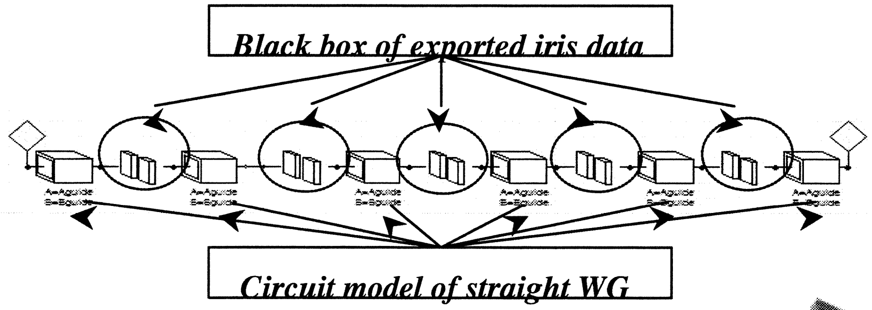
Single Iris Field Solution

- Solve frequency sweep for several window widths
- De-embed to iris and use *DYNAMIC LINK* between HFSS and Ansoft Designer



Circuit Model of Complete Filter

- 11 Variables:
  - Window of each iris; lengths of straight WG sections
- Optimize filter response

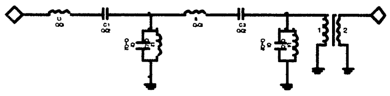


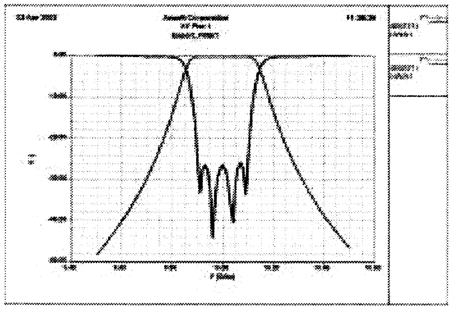
Black box of exported iris data

Circuit model of straight WG

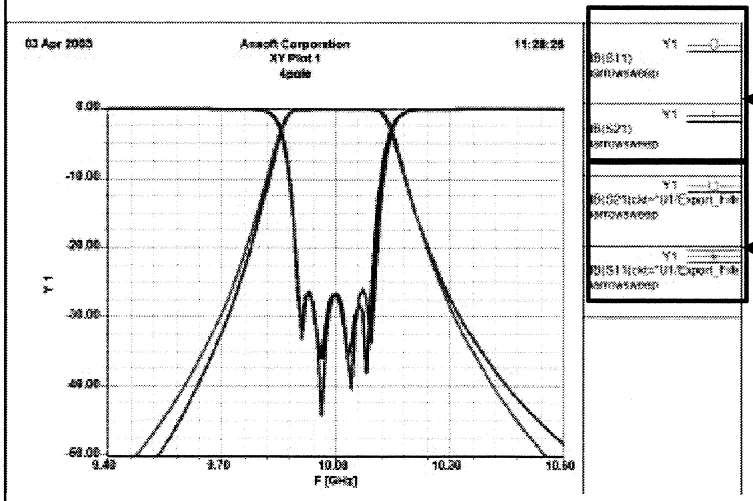
Circuit Model of Desired Filter

4-pole lumped element filter with desired center frequency and BW





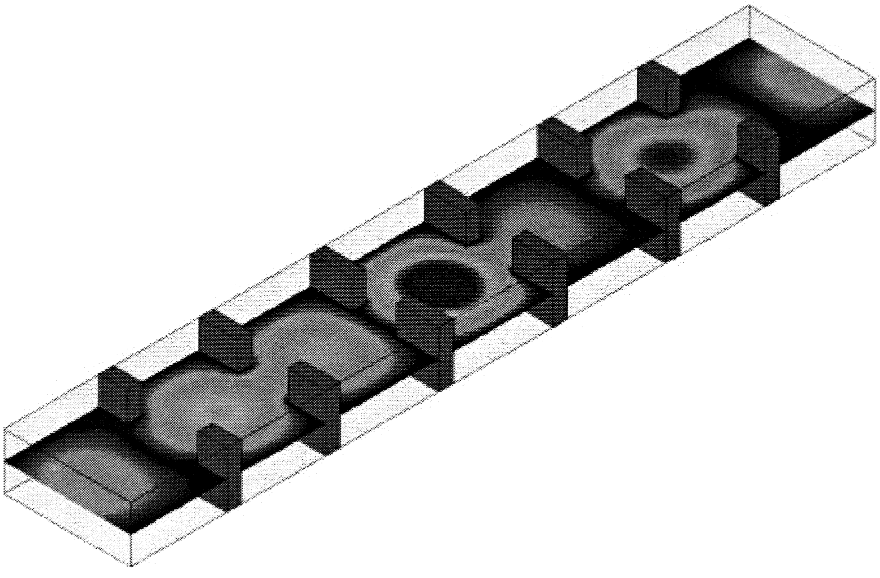
Optimized EM-Circuit Response



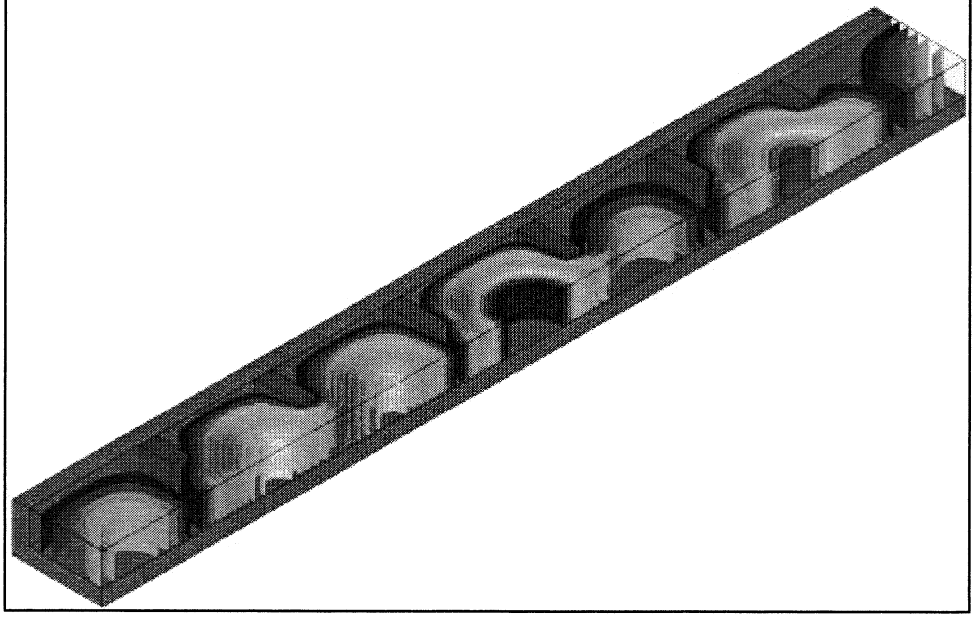
Optimized EM/circuit filter

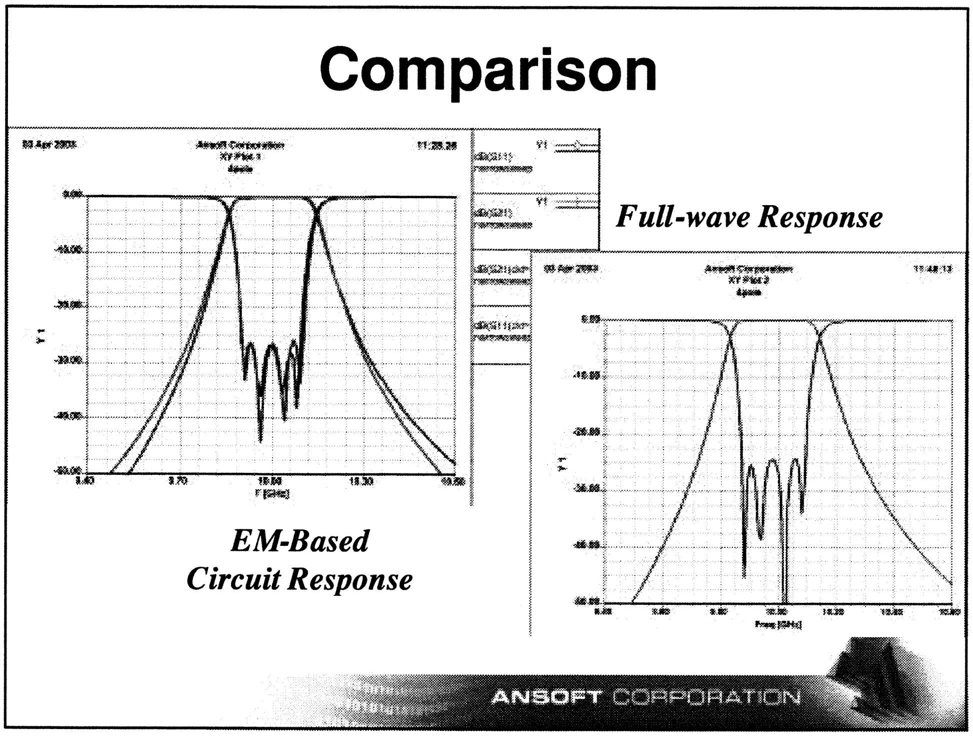
Lumped element target circuit

Optimized Full-wave Response



Optimized Full-wave Response





Comparison

Full-wave Response

EM-Based Circuit Response

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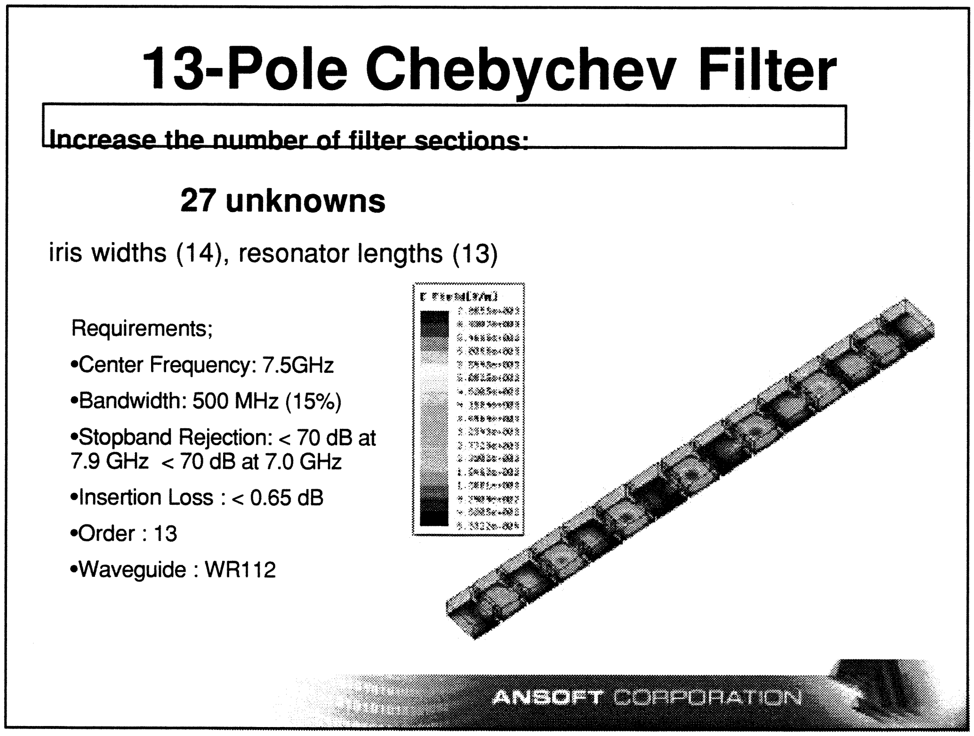

13-Pole Chebychev Filter

Increase the number of filter sections:

27 unknowns

iris widths (14), resonator lengths (13)

Requirements;

- Center Frequency: 7.5GHz
- Bandwidth: 500 MHz (15%)
- Stopband Rejection: < 70 dB at 7.9 GHz < 70 dB at 7.0 GHz
- Insertion Loss : < 0.65 dB
- Order : 13
- Waveguide : WR112

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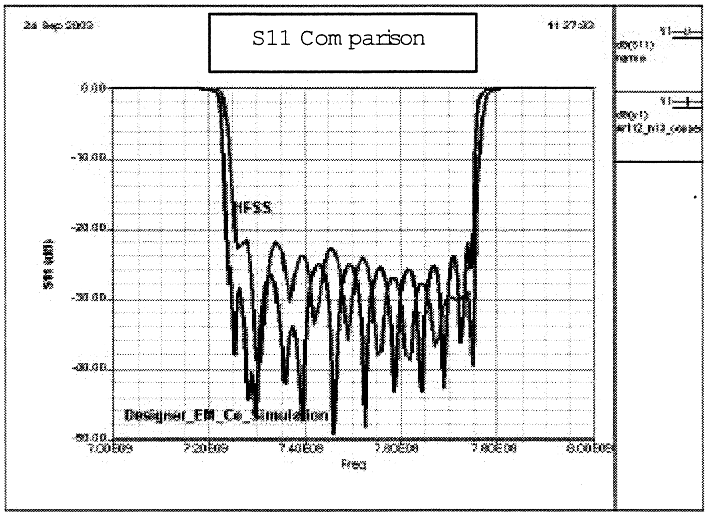
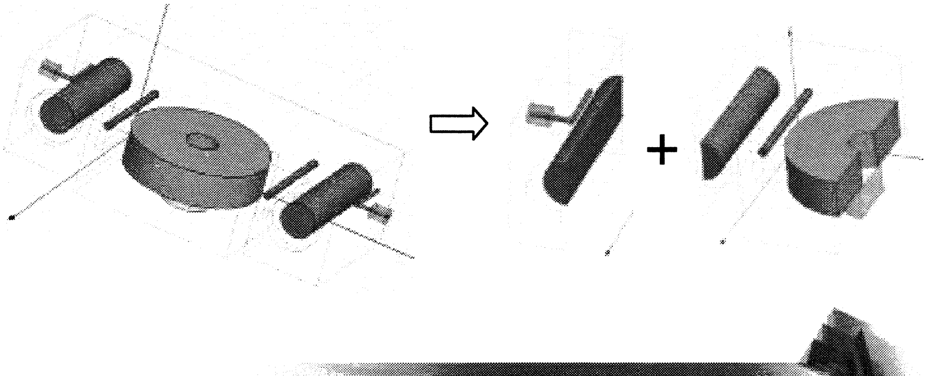
Comparison: Full EM vrs. Cosimulation

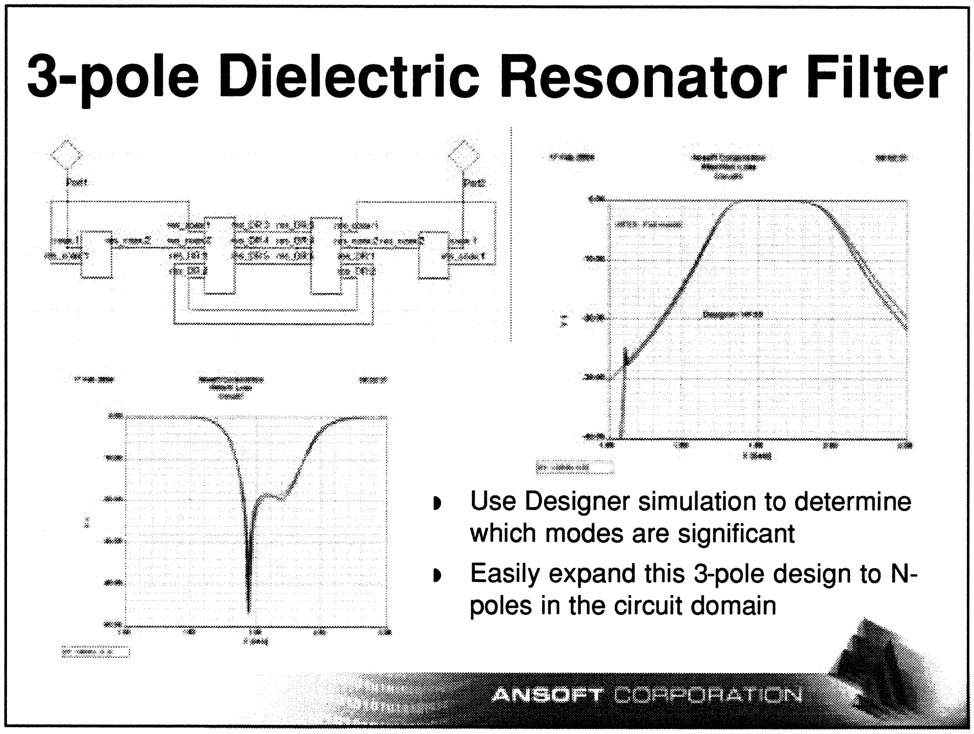


M odeland resu lts courtesy of
RS Microwave

3-pole Dielectric Resonator Filter

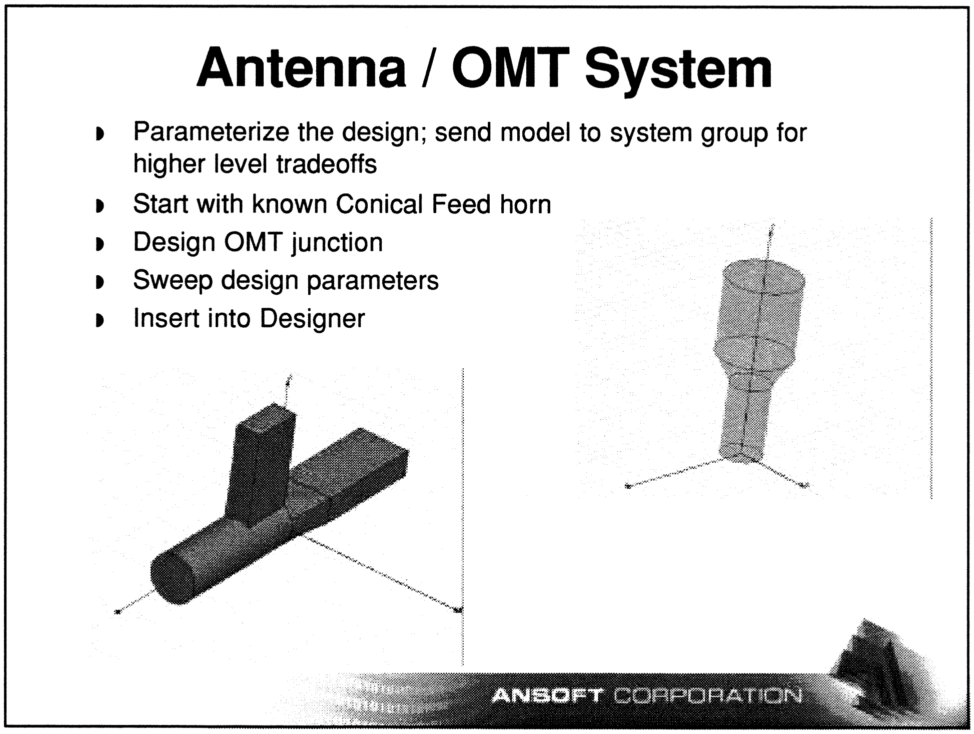
- Break model on geometric planes of symmetry
  - Higher order modes will definitely be generated at ports
  - Have HFSS solve for these higher order modes, and connect them together in circuit simulator





3-pole Dielectric Resonator Filter

- Use Designer simulation to determine which modes are significant
- Easily expand this 3-pole design to N-poles in the circuit domain

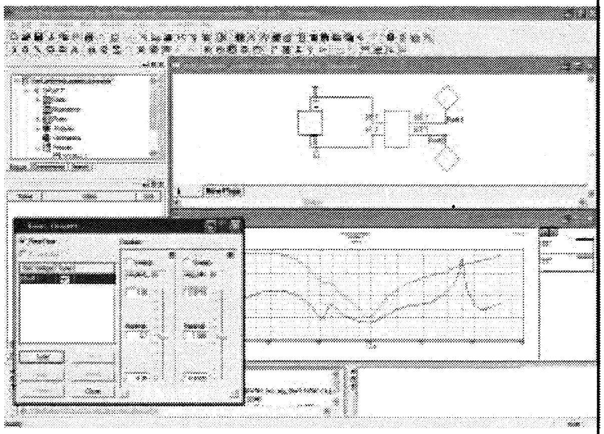


Antenna / OMT System

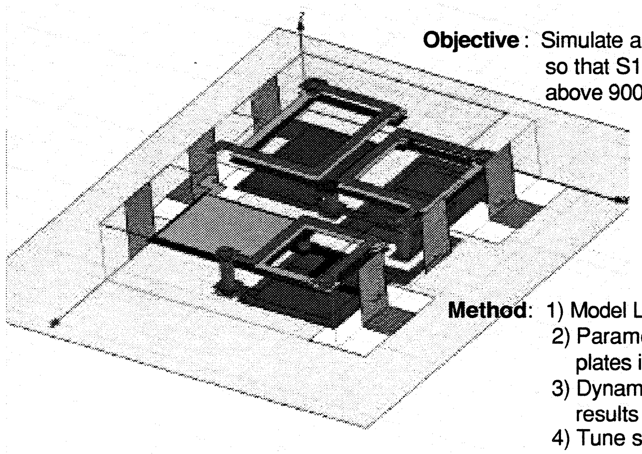
- Parameterize the design; send model to system group for higher level tradeoffs
- Start with known Conical Feed horn
- Design OMT junction
- Sweep design parameters
- Insert into Designer

System Level Tradeoffs

- Insert N-port blocks for Antenna & OMT
- Wire together the two modes from the Antenna to the circular port on the OMT
- Add Microwave ports to the RWG ports of the OMT
- Analyze & Tune interpolated solutions
- Fix optimum values for parameters and "Simulate Missing Solutions"



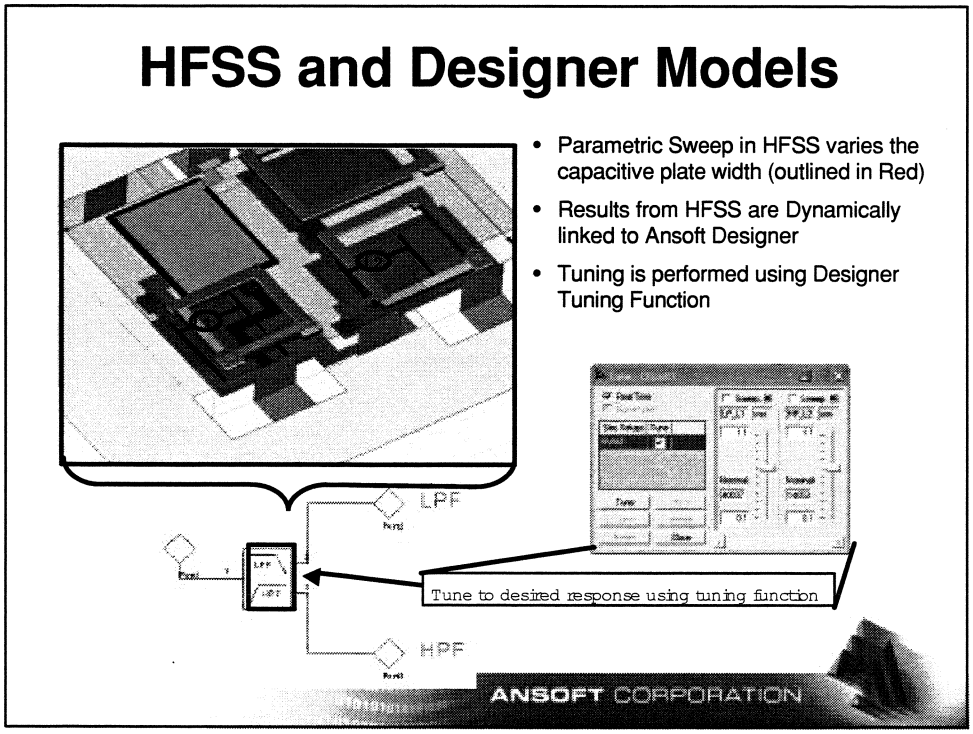
LTCC Diplexer



Objective : Simulate and tune LTCC Diplexer so that S11 <= -10 dB above 900 MHz

Method: 1) Model LTCC in HFSS
2) Parametrically sweep capacitive plates in model
3) Dynamically link HFSS results into Designer
4) Tune structure in Designer

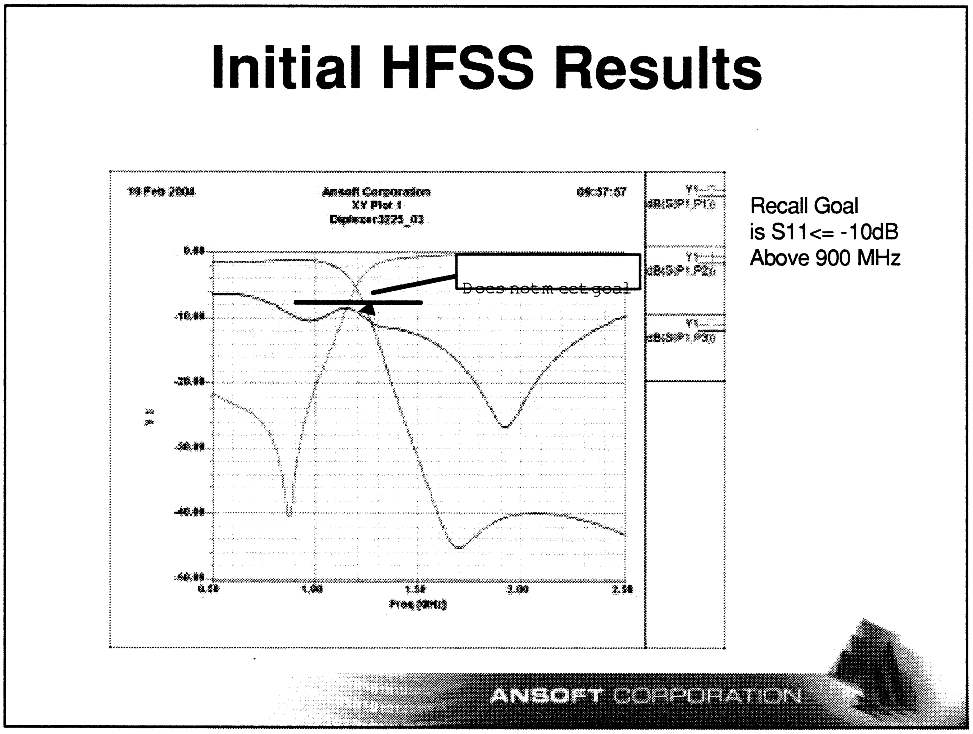
Benefits : Diplexer was tuned in real time.
Engineer has visual indication of filter performance while optimizing it.



HFSS and Designer Models

- Parametric Sweep in HFSS varies the capacitive plate width (outlined in Red)
- Results from HFSS are Dynamically linked to Ansoft Designer
- Tuning is performed using Designer Tuning Function

Tune to desired response using tuning function

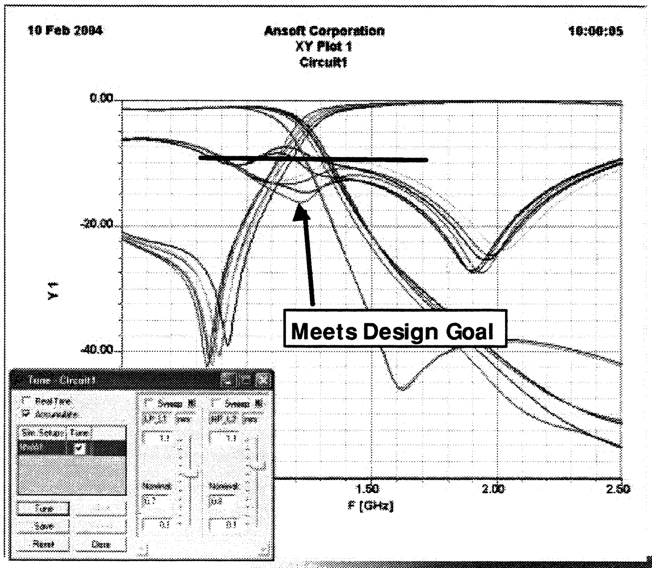


Initial HFSS Results

Does not meet goal

Recall Goal
is S11<= -10dB
Above 900 MHz

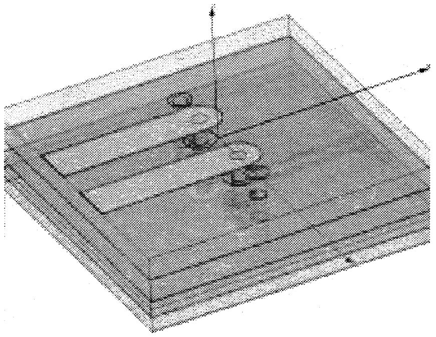
Tuned HFSS Results



Multiple tuned responses are shown.

Indicated trace well exceeds performance specification

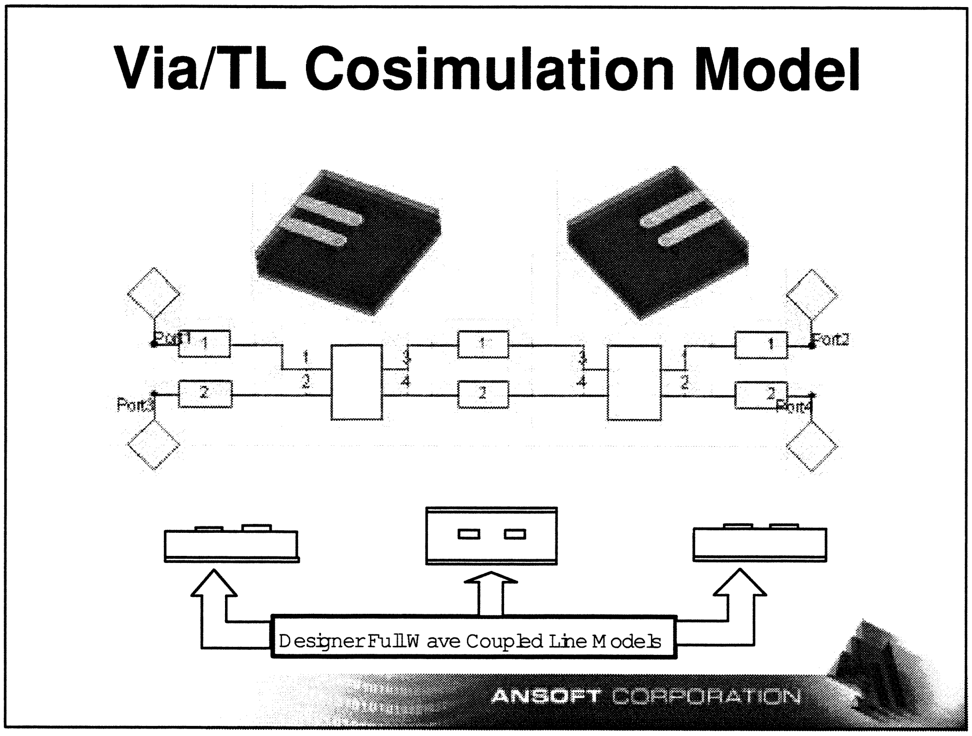
Via/Transmission Line/System



Objective: Simulate a transmission path of a differential pair and produce an Eye Diagram for the path.

Method:
1) Model Differential via in HFSS
2) Dynamically link HFSS results into Designer Circuit tool
3) Model long microstrip and stripline sections in Designer Circuit tool
4) Analyze the HFSS/Circuit at the system level
5) Produce desired Eye Diagram

Benefit: Modeling discontinuity in HFSS only reduces solution time. Combining circuit and HFSS models in Designer allows rapid design changes. System level performance is quickly determined by using system tool



Via/TL Cosimulation Model

Designer FullWave Coupled Line Models

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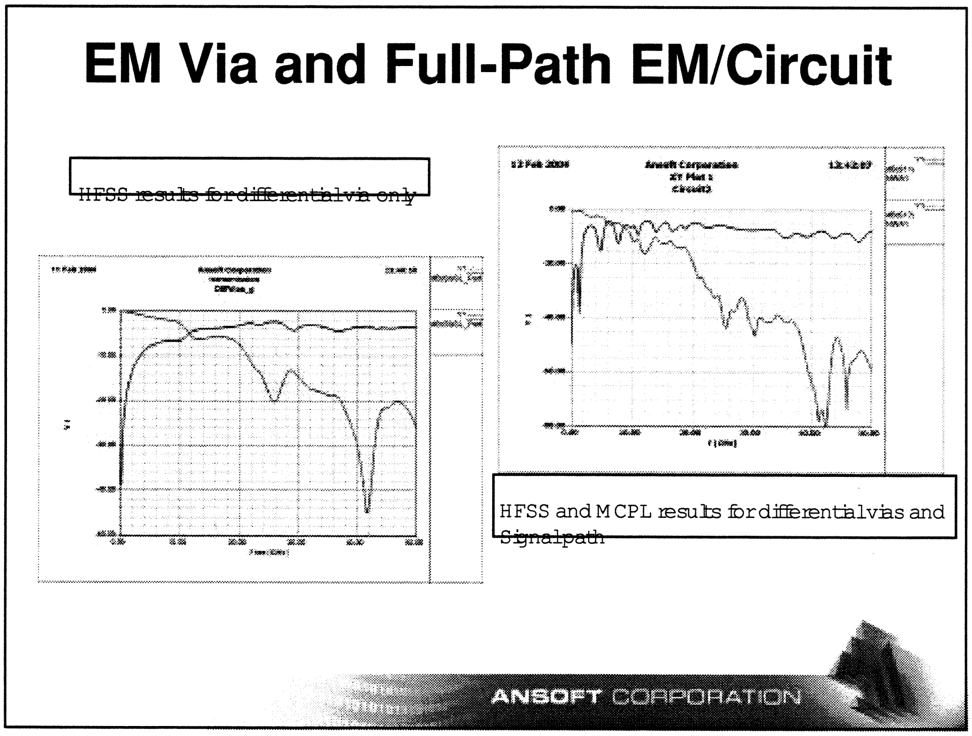

EM Via and Full-Path EM/Circuit

HFSS results for differential via only

HFSS and MCPL results for differential vias and Signal path

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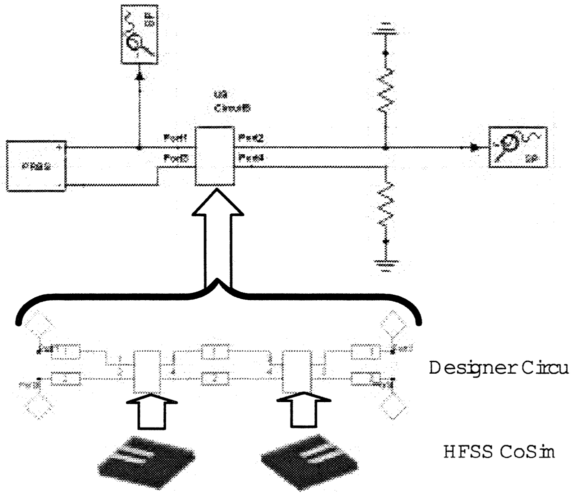
System Simulation



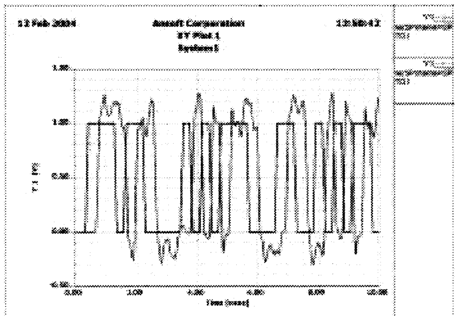
Bit Rate: 3.2 Gbps
Rise Time: 100 ps

Designer Circuit CoSim

HFSS CoSim

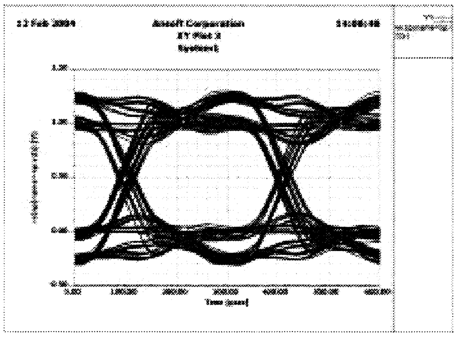
Full System Results

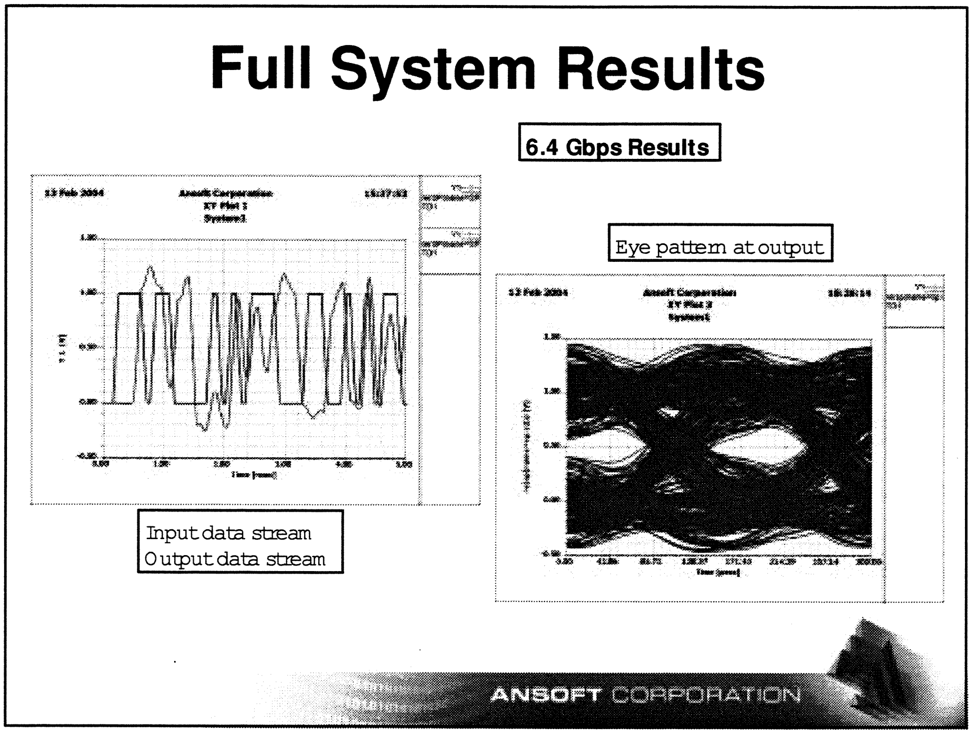
3.2 Gbps Results



Input data stream
Output data stream

Eye pattern at output



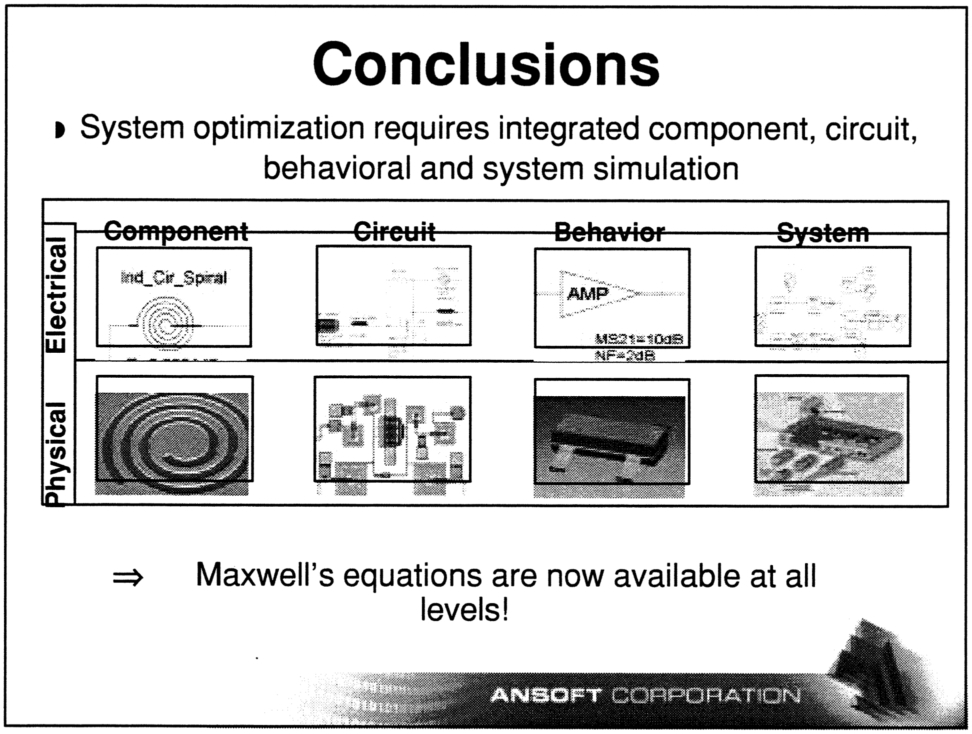


Full System Results

6.4 Gbps Results

Input data stream
Output data stream

Eye pattern at output



Conclusions

- System optimization requires integrated component, circuit, behavioral and system simulation

⇒ Maxwell's equations are now available at all levels!

IEEE MTT Symposium 2004 Workshop

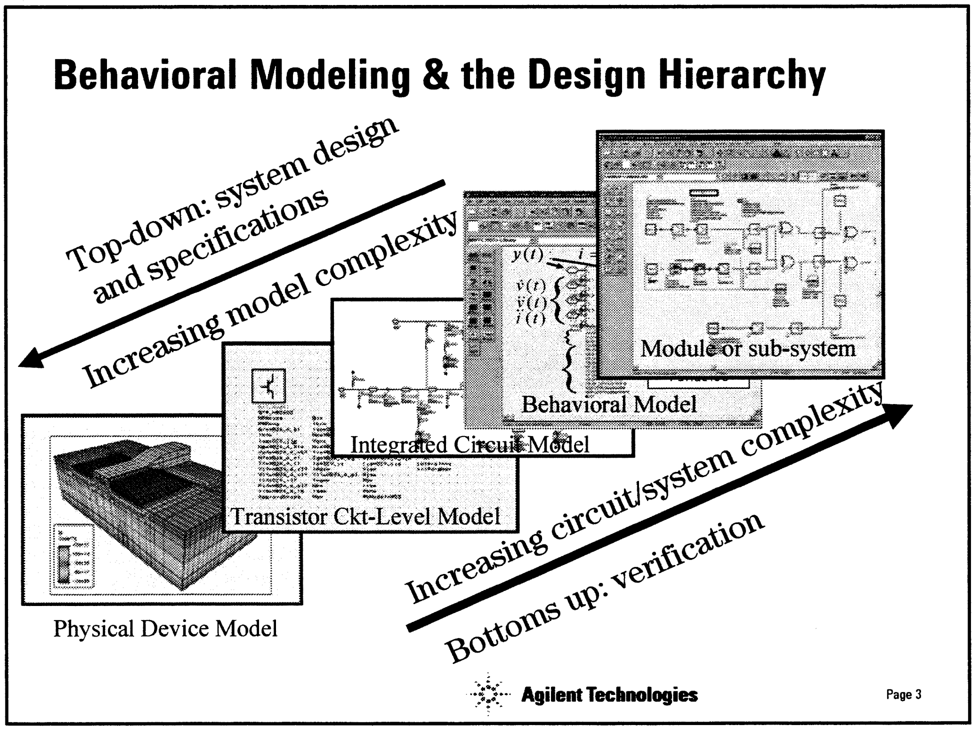
Advances in Compact Device Models and Data-Based Behavioral Models

David E. Root, John Wood, and Steve Chen

Agilent Technologies

Outline

- The Model and Simulation Hierarchy
- Trends and Advances in Nonlinear Device Modeling
- Advances in Nonlinear Black-Box Data-based Modeling
- Conclusions



Behavioral Modeling & the Design Hierarchy

Top-down: system design and specifications

Increasing model complexity

Module or sub-system

Behavioral Model

Integrated Circuit Model

Transistor Ckt-Level Model

Physical Device Model

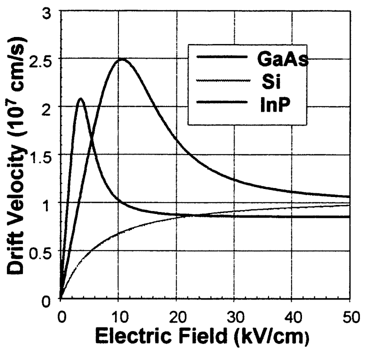
Increasing circuit/system complexity

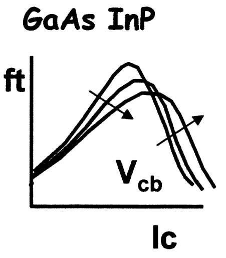
Bottoms up: verification

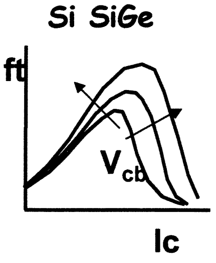
Trends in Nonlinear Device Modeling (1)

- Apply standard models to new device technologies
  - Extract Curtice Cubic model for GaN FET devices
  - Extract Gummel-Poon based Si BJT models for GaAs based HBTs
- Develop new models for "new" device technologies
  - III-V HBTs
  - LDMOS

III-V HBTs vs. Si BJTs







Microscopic transport properties -> device-level characteristics
Problems fitting III-V devices to Si BJT models

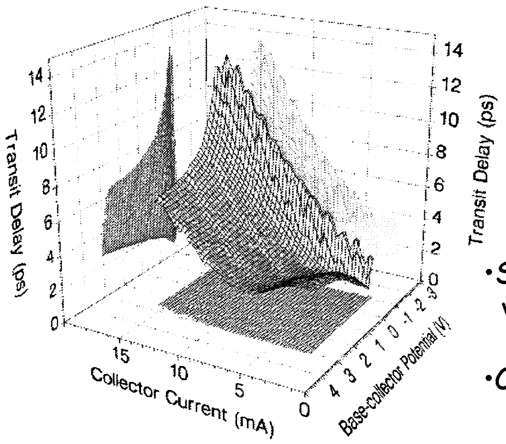
Models with III-V physics work better for GaAs InP HBTs [14]

Nonlinear transit-time models necessary to fit f_t characteristics

---

Charge/Delay Model: Base-Collector Delay [14]

Measured base-collector transit time



$$\tilde{\tau}_C = \frac{\partial Q_{CC}}{\partial I_C} = \frac{\text{Im}(Y_{21}+Y_{22})}{\omega(g_m+g_{CE})}$$

$$\tilde{C}_C = \frac{\partial Q_{CC}}{\partial V_{BC}} = \frac{\text{Im}(Y_{21})}{\omega} - \tilde{\tau}_C \cdot g_m$$

$$Q_{CC} = \int \tilde{C}_C dV_{BC} + \tilde{\tau}_C dI_C$$

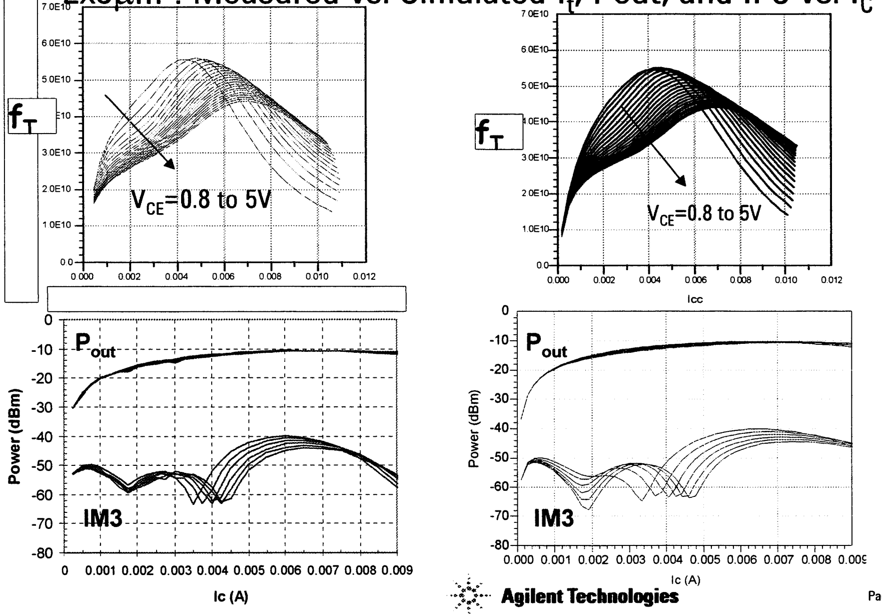
$$(Q_{dif} \neq \tau(V,I)\cdot I)$$

- Shape of τ_C related to velocity-field curves (physics)
- Can compute terminal charges from line-integration of data or adjoint neural networks

Physics, device data, and model equations consistent

Device-Level Validation GaAs Power SHBT > 0.5 W

2x8μm^2: Measured vs. Simulated f_t, Pout, and IP3 vs. I_C

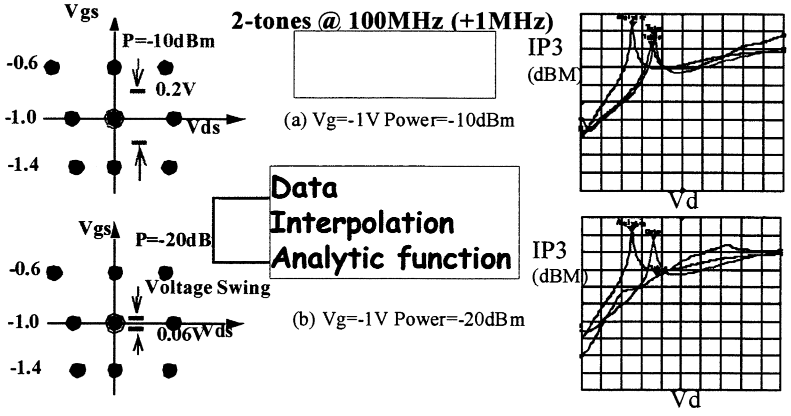


Trends in Nonlinear Device Modeling (2)

- Combine empirical and table-based models
  - Replace model parameters with tables of values [17]
  - Use analytical (empirical) models for I-V; Table-based models for Q-V [18]
- Improved interpolation / approximation techniques for table-based models
  - B-splines [19],[20]
  - Neural Networks [16],[5]

Multi-variate functions (1) Interpolation

$$y(t) = f(V_1(t), V_2(t))$$

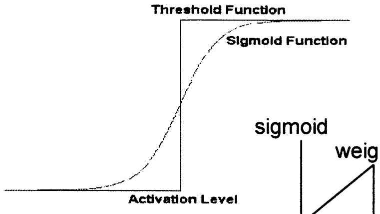


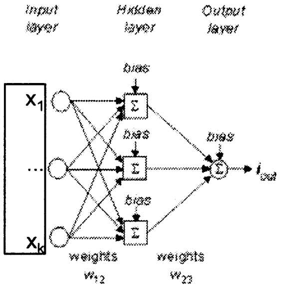
Data
Interpolation
Analytic function

Interpolation algorithms can limit distortion simulation at low powers in otherwise good models

Multi-variate functions (2) Artificial Neural Networks

A NN is a parallel processor made up of simple, interconnected processing units, called *neurons*, with weighted connections.





sigmoid
weights
biases

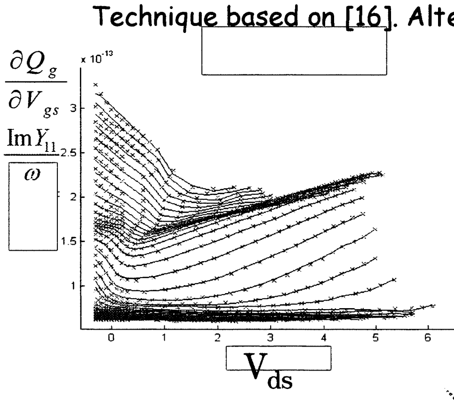
$$F(x_1,\ldots,x_K) = \sum_{i=1}^{I} v_i s\left(\sum_{k=1}^{K} w_{ki}x_k + a_i\right) + b$$

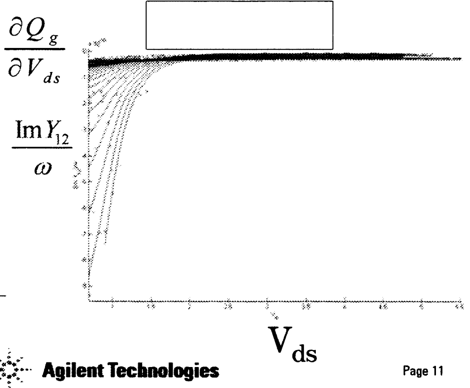
- **Universal Approx. Thm: Can fit any nonlinear function of any # of variables**
- **Infinitely differentiable: good for distortion.**
- **Easy to train (fit) using standard third-party tools (MATLAB)**

Constructing FET Q_g Charge from data (Y_{11i}, Y_{12i}) using Adjoint Neural Networks

For nonlinear data-based transistor models
Terminal Charges for FET & HBT models

Technique based on [16]. Alternative to line-integration [15].





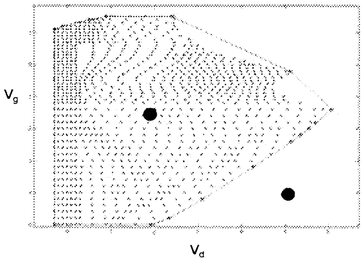
Boundary and Guided Extrapolation

Example: 2-d data

Evaluate $f(V_1, V_2)$ by:

- If (V_1, V_2) within Validity Region
  - Interpolation or
  - Fit function
- Otherwise
  - *Output something smart* that will cause the simulator to return to the validity region
  - Give user a warning that the model is extrapolating

Convex Hull



For an arbitrary region R The *convex hull* of R is the smallest volume convex region containing R.

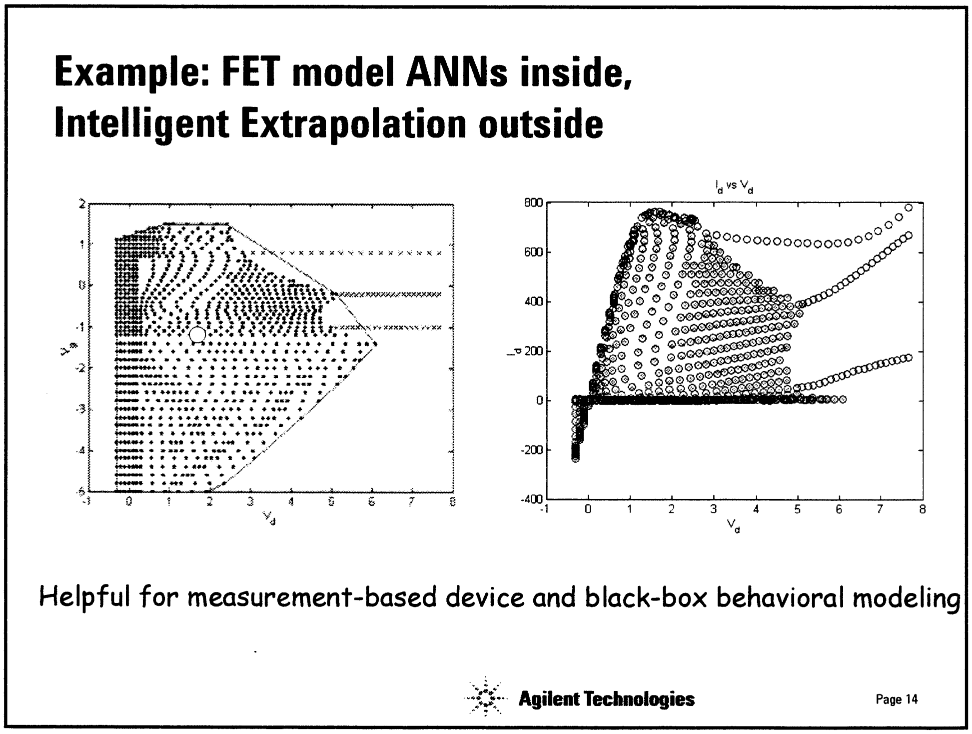
Guided Extrapolation

"Output something smart" means

- Non-singular evaluation
- Desirable asymptotic behavior
  - 'Push' the simulator back toward the Valid region
- Ensure continuity of first derivatives
- No local maxima or minima

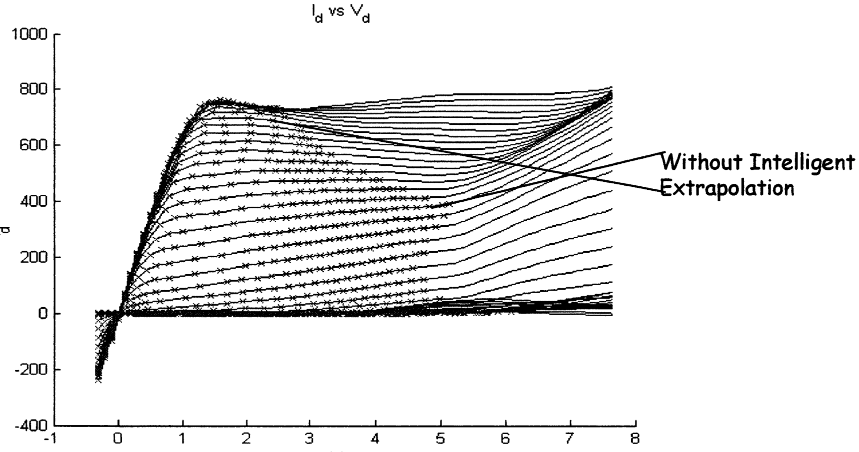
- Speed is important
- Need not be physically correct

Example: FET model ANNs inside, Intelligent Extrapolation outside



Helpful for measurement-based device and black-box behavioral modeling

Example: FET model: ANNs inside, Intelligent Extrapolation outside



"Perfect fit" and Infinitely Differentiable within Boundary
Asymptotic Behavior Keeps Extrapolated Curves from Crossing

Look to new modeling languages: Verilog-A

- Verilog-A : an Analog Hardware Description Language
- Language for modeling behavior of analog components at their external terminals/ports
- Open/public-domain language and an IEEE standard
- Easy enhancements to existing models

- Verilog (digital) - Verification and logging
  - Verilog-A (analog)
    - Verilog-AMS (mixed signal)

Value of Verilog-A

- Dramatically increased modeling productivity
  - Modelers focus on model eqs. not simulator interface
  - No need for tedious partial derivatives or C-code
  - Independence from simulator vendors for compiled code
  - Possible general way to interface with simulator engine for new capabilities: Behavioral Modeling

Simple Verilog-A Example

Define a module in Verilog-A, this example is called `myR` →

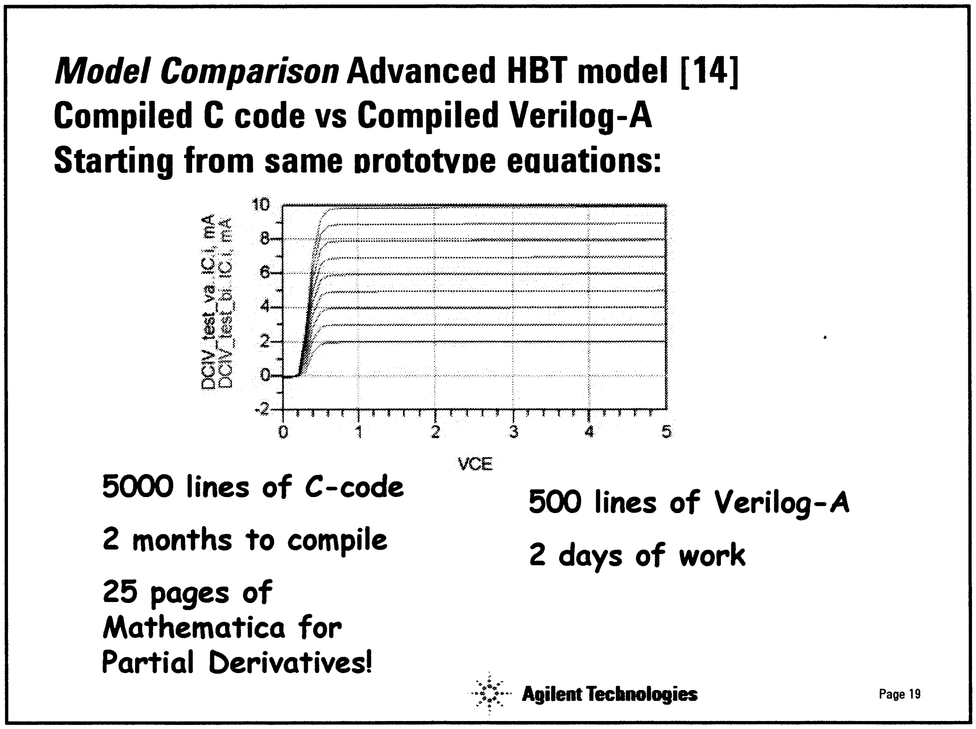
```
module myR(p,n);
  electrical p,n;
  parameter real R=1;
  analog V(p,n) <+ I(p,n)*R;
endmodule
```

```
; Test myR
V1:vsrc 1 0 vdc=1
myR:r1 1 2 R=1
myR:r2 2 0 R=1
dc:dc1
```

← **The device `myR` is now available for use. Here we create two instances r1 and r2.**

Results stored in the dataset in the standard way. →

```
Node    v(1)    v(2)
        1.0     0.5
```



Model Comparison Advanced HBT model [14] Compiled C code vs Compiled Verilog-A Starting from same prototype equations:

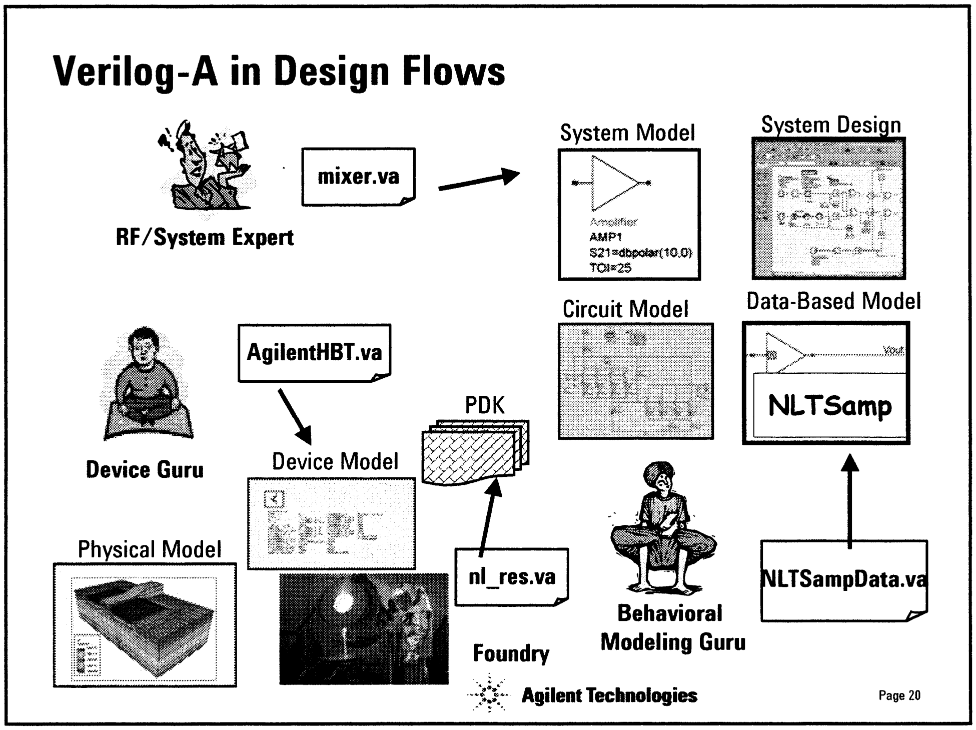
5000 lines of *C*-code

2 months to compile

25 pages of Mathematica for Partial Derivatives!

500 lines of Verilog-*A*

2 days of work



Verilog-A in Design Flows

RF/System Expert

mixer.va

System Model

Amplifier
AMP1
S21=dbpolar(10,0)
TOI=25

System Design

Circuit Model

Data-Based Model

NLTSamp

Device Guru

AgilentHBT.va

Device Model

PDK

Physical Model

nl_res.va

Foundry

Behavioral Modeling Guru

NLTSampData.va

Verilog-A Applications and Limitations

- Circuit Design
  - Transistor models
  - Limited interpolation capability a problem for table-based models
  - Custom component models in PDKs
- A/RF System Design
  - Top-down models for mixers, amps, PLLs, oscillators...
  - Bottom-up black-box behavioral models in V-A not fully ready

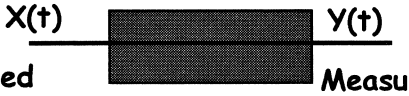
Black-Box Behavioral Models

- Simplified (fast) model of essential nonlinear behavior

- Derivable from simulations and measurements
  - Derive models at higher abstraction levels using the simulator
  - Data-based models can be more accurate.
    No errors propagated from lower level models

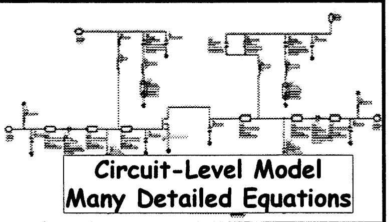
- For Black-Box Models: add constraint:
  At ports connected to the external world
  - No internal information about component necessary

Behavioral Modeling Flow

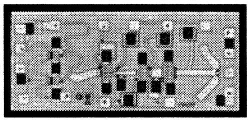


Simulation-Based

Measurement-Based



Circuit-Level Model
Many Detailed Equations



Physical Component
Probe at inputs/outputs

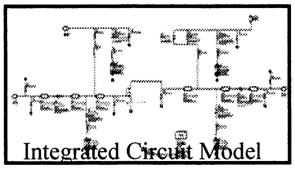
Equation
Order
Reduction

Excite and Measure
in Simulator
TA, HB, Envelope
(black-box)

Excite and Measure
Response with NVNA,
Scope, VSA, etc.
(black-box)

$F[x(t), y(t)] = 0$ **Functional Relationship between input/output**
Differential equations; describing functions; mixed domain

Circuit-level simulation and nonlinear device models



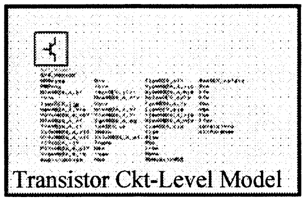
Integrated Circuit Model

Simulators:

SPICE (Time domain)

Harmonic Balance (Freq. Domain)

Transient Envelope (Mixed Domain)



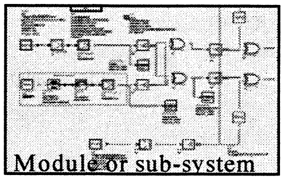
Transistor Ckt-Level Model

"Compact" (transistor, diode) Models

Time-domain "SPICE Models"

$$I(t) = I_s(e^{v(t)/kT} - 1) + \frac{C_{j0}}{\sqrt{1 - v(t)/\Phi}} \frac{dv(t)}{dt}$$

System level simulation & nonlinear behavioral models



Simulators:

- **Transient Envelope (Mixed Domain)**
- **Harmonic Balance (Freq. Domain)**
- **SPICE (Time-domain)**

Black Box Behavioral Model of IC

Behavioral Models

- **Still mostly time-domain models**
- **sometimes used w. freq. domain filters**

Behavioral models do not take full advantage of the benefits of available simulation algorithms

---

Native Behavioral Models

Objectives:

Formulate black-box behavioral models in a language native to the simulator solution technique.

- **Most efficient for simulation**
- **Most efficient for model generation**
- **But not all models will simulate in all domains**

Requirements:

- **Access to the underlying simulator algorithm!**
- **Open simulator architecture**
- **Ability to schedule simulations**

Time Domain Behavioral Models and Simulation

Modeling problem: map input waveforms to output waveforms. (Functionals) y(t)=F[x(t)] or G[x(t),y(t)]=0

- **Models usually defined by nonlinear ordinary differential equations (NL ODEs)**
  - **Developed by Nonlinear Time Series or Dynamic Neural Networks**
  - **ODE models will simulate, in principle, in all in all other domains (HB, Envelope)**

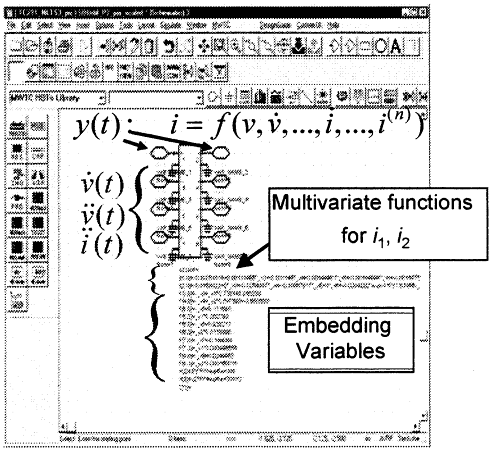
Examples: $y(t) = f(x(t)) + g(x(t))\frac{dx(t)}{dt}$ **static**

$y^{(n)} = f(y^{(n-1)}, \ldots y, x, \dot{x}, \ldots x^{(m)})$ **dynamic**

NLTS Models Implemented in Simulator

$$i(t) = f\left(v(t), \dot{v}(t), \ddot{v}(t), \ldots, v^{(n)}(t), \dot{i}(t), \ldots, i^{(m)}(t)\right)$$

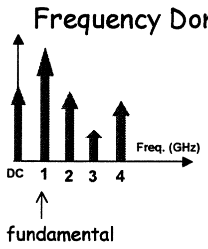
- Implemented as an "SDD" *Symbolically Defined Device* Interpreted user-defined nonlinear ODE interface
- SDD *automatically generated* (proper time derivatives, neural network functions, etc.)



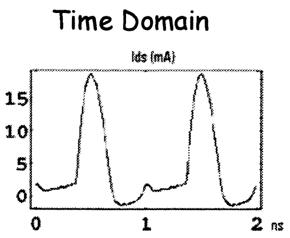
Frequency Domain Models: Harmonic Balance

Modeling problem: map input spectrum to output spectrum

- Applies to systems under large-signal periodic drives
- Discrete harmonic spectra for all inputs, outputs, & state variables


Frequency Domain

$$x(t) = \mathrm{Re}\left(\sum_{h=0}^{H} X_h \, e^{j\,2\pi h f_0 t}\right)$$


Time Domain

$\{X_h\}$ complex (amplitude & phase) numbers at each harmonic index h

Example: "Scattering Function" amplifier model [4]

- Mismatch under large-signal drive
  - Considering only the fundamental signal components

$$b_2 = S_{21}(|a_1|)\cdot a_1 + S_{22}(|a_1|)\cdot a_2 + S'_{22}(|a_1|)\cdot e^{j\cdot 2\cdot\phi(a_1)}\cdot a_2^*$$

  - More accurate "power-dependent" S-parameter behavioral model
  - Basis for rigorous definition of "Hot S22"

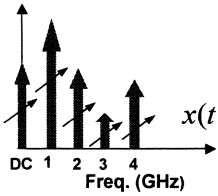
Envelope Domain Models and Simulation

Modeling problem: map input envelopes to output envelopes

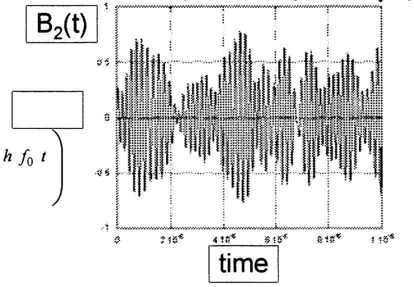
- Applies to systems under large-signal modulated drives
- Time-varying spectra for all inputs, outputs, & state variables

Time Domain (envelope)

Time-varying spectrum



$$x(t) = \mathrm{Re}\left(\sum_{h=0}^{H} X_h(t)\; e^{j\,2\pi\,h\,f_0\,t}\right)$$



{$X_h(t)$} complex (amplitude & phase) waveforms at each harmonic index h

Envelope Models: Recent Advances

- Replace static output envelope function with a *functional*

$$\hat{Y}(t) \neq f(\hat{X}(t)) \to F[\hat{X}(t)]$$

$$F[\hat{X}(t)] = f(\hat{X}(t), \hat{X}(t-\tau), \ldots, \hat{X}(t-n\tau))$$

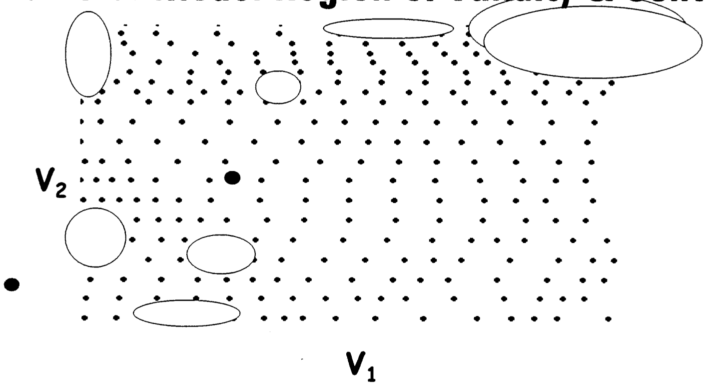
- Various approximations for f lead to different models [10,11]
- Alternatively, embed the envelopes using time derivatives

$$F[\hat{X}(t)] = f(\hat{X}(t), \dot{\hat{X}}(t), \ldots, \hat{X}^{(m)}(t))$$

Or, more generally, write (coupled) ordinary differential equation models for envelopes

$$\hat{Y}^{(n)}(t) = f(\hat{Y}^{(n-1)}, \ldots, \hat{Y}(t), \hat{X}(t), \dot{\hat{X}}(t), \ldots, \hat{X}^{(m)}(t))$$

Behavioral Model Region of Validity & Convergence



Example: static model $y(t) = f(V_1(t), V_2(t))$

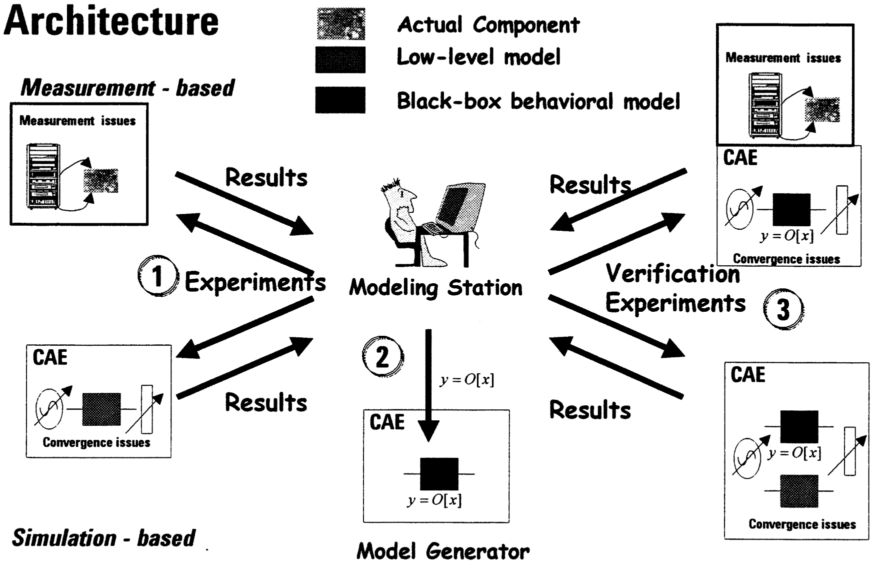
Assume model valid for black point (interpolation in state-space)

Extrapolation required for red point.

Apply Neural Networks w. Boundary Recognition / Extrapolation

Data-Based Black Box Modeling: Open Architecture

▨ Actual Component
■ Low-level model
■ Black-box behavioral model



Conclusions

- Advances and trends in nonlinear device modeling were reviewed
  - Physics-based, measurement-based, CAD-oriented approaches, and new languages were highlighted
- Advances in Black-Box Behavioral models can be achieved by formulating them in language native to the simulator
  - Development of open simulator architectures and enhancements to standard modeling languages are suggested to enable progress

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- Doug Rytting
- Agilent Management

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WFD – Advances and New Directions in Device Modeling and Design Optimization for Microwave CAD

Introduction and Application of Cortical Theory to RF Amplifier Modeling and Compensation

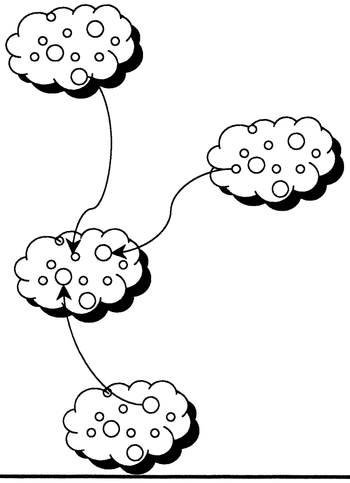
Paul Draxler
QUALCOMM Incorporated
University of California, San Diego
June 11th, 2004

Adding Smarts into RF Amplifiers

- Increasing radio environment complexity
  - Versatility (switched matching and bias networks,...)
  - Signal Environments (jammers, modulation, ...)
- The world is a very nonlinear place, with time varying nonlinear transfer functions, and evolving scenarios.
- Direct analytical techniques become overly complex, but even mice can deal with life.

Outline

- An Application Example - Phrase Completion
- Biological Model
- Other Combinations
- RF Power Amplifier Modeling
- Pre-Distortion
- Applied Cortical Theory
- Conclusions



Application Example – Phrase Completion

down the garden → ______

it was a → ______

take the train → ______

are covered with → ______

there were many → ______

when lockheed makes → ______

are easy to → ______

score the winning ______

Application Example – Discrete Input / Output States

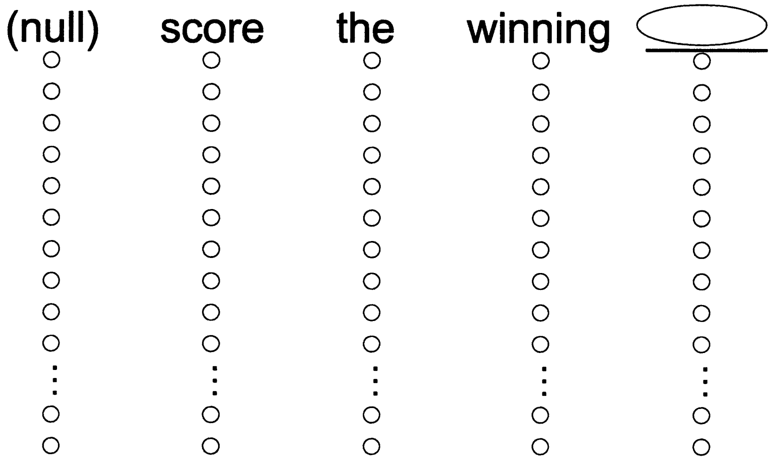
For the case of this example,

- we will have the same discrete input and output states.
- 10,000 English word lexicon.

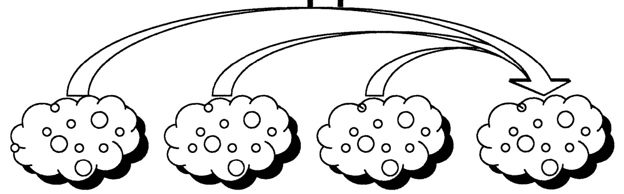
Active state is denoted by the solid circle.

1	○	the
2	○	of
3	○	to
4	○	##
5	○	a
6	○	and
7	○	in
8	○	that
.	○	.
.	○	.
.	○	.
9,999	○	venice
10,000	○	congressmen

Application Example – Phrase Representation



Constructs: Antecedent Support Networks



Source [$src_{(to-3)}$] $src_{(to-2)}$ $src_{(to-1)}$ Target [$tgt_{(to)}$]

- Build Antecedent Support Networks
  - Parse corpus – set rules, lexicon, then count events
    - Training (raw counts)
    - Education (encouraged)
  - Calculate ASN probabilities
  - Honing – least likely wrong answer…

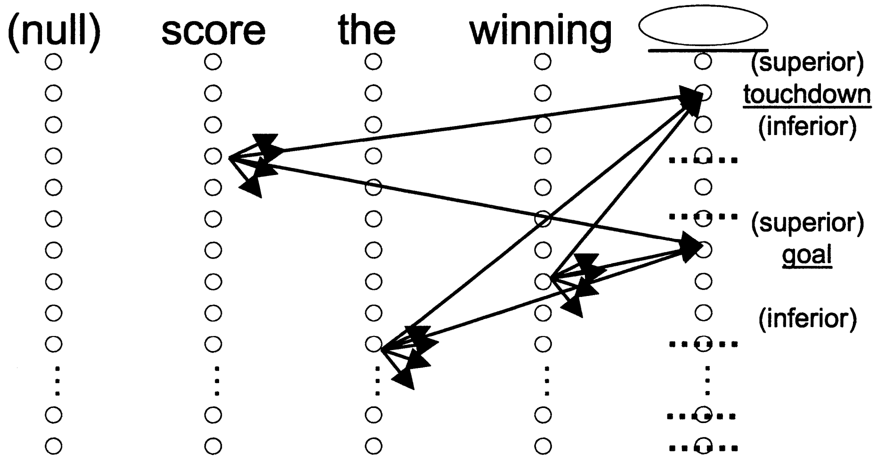
$$C(src,tgt)$$

$$C(tgt)$$

$$P(i_{src} \mid j_{tgt}) = C(i,j) / C(j)$$

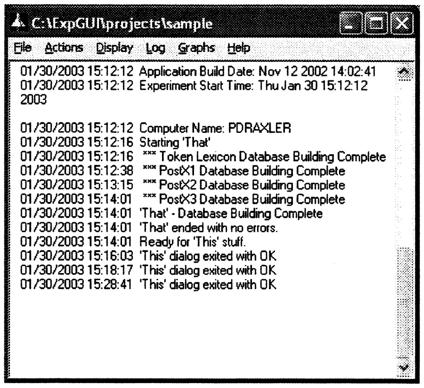
$$\max\{\min[P(i_1,j),P(i_2,j),P(i_3,j),...]\}$$

Application Example – Finding Candidate Answers



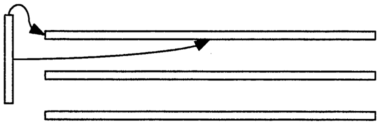
Implementation

- Build a ASN Knowledge Base using:
  - ~3,000,000,000 word training corpus
  - Lexicon: 10000 words
  - Sparse PostX1 Count Data
    - 13.5M / 100M
  - Sparse PostX2 Count Data
    - 22.6M / 100M
  - Sparse PostX3 Count Data
    - 25.8M / 100M

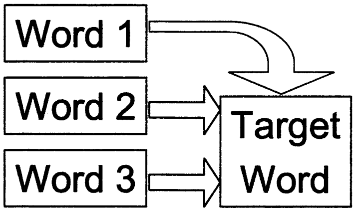


```
01/30/2003 15:12:12 Application Build Date: Nov 12 2002 14:02:41
01/30/2003 15:12:12 Experiment Start Time: Thu Jan 30 15:12:12
2003

01/30/2003 15:12:12 Computer Name: PDRAXLER
01/30/2003 15:12:16 Starting 'That'
01/30/2003 15:12:16  *** Token Lexicon Database Building Complete
01/30/2003 15:12:38  *** PostX1 Database Building Complete
01/30/2003 15:13:15  *** PostX2 Database Building Complete
01/30/2003 15:14:01  *** PostX3 Database Building Complete
01/30/2003 15:14:01 'That' - Database Building Complete
01/30/2003 15:14:01 'That' ended with no errors.
01/30/2003 15:14:01 Ready for 'This' stuff.
01/30/2003 15:16:03 'This' dialog exited with OK
01/30/2003 15:18:17 'This' dialog exited with OK
01/30/2003 15:28:41 'This' dialog exited with OK
```



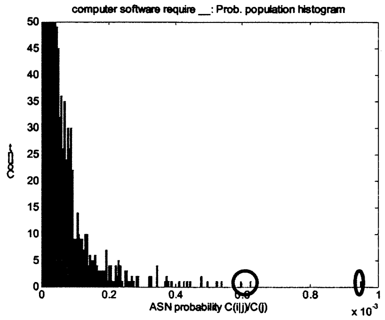
Outputs



- String outputs:
  - Three input words and indices
  - Top contenders with p(i|j)'s
- MATLAB outputs:
  - Indexed result plots
  - Zoomed Histograms of results
  - Log-Log Historgams (w/ bars)
    - Each contributing p(i|j)
    - Post honing support
- Show some interesting words

  Computer software require ____

  xxx, yyy, zzz



Text Output FAST…

```
01/30/2003 10:52:10     ***   NEW RUN   ***
01/30/2003 10:52:10  Source Word1:      target    #1489
01/30/2003 10:52:10  Source Word2:      strange   #4405
01/30/2003 10:52:10  Source Word3:      effects   #2182
01/30/2003 10:52:12  ------- Paul's Top 3 ------- final AS --- X3 --- X2 --- X1 ---
01/30/2003 10:52:12   Target1: #9024  harmful  0.000085 = min(0.000169, 0.000085, 0.000254)
01/30/2003 10:52:12   Target2: #5563  emerge  0.000082 = min(0.000082, 0.000082, 0.000123)
01/30/2003 10:52:12   Target3: #8126  bent  0.000073 = min(0.000073, 0.000073, 0.000073)
```

Application Example – Phrase Completion Answers

down	the	garden	(path, ridge, carpet, soil)
it	was	a	(clear, logical, frustrating)
take	the	train	(ride, trip, trips, rides, journey)
are	covered	with	(medicaid, thick, mud, plastic)
there	were	many	(surprises, casualties, factors)
when	lockheed	makes	(airplane, missiles, fighter)
are	easy	to	(install, understand, dismiss)
score	the	winning	(touchdown, goal, basket)

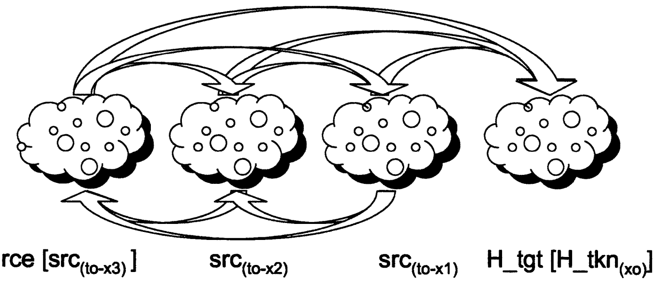
Summary of Attaching the Problem

- Define discrete input / output target states.
- Build pairwise statistical dependencies (from a training corpus) between assumed facts or context and the target or solution region.
- Track the context and identify best solution states.
- Separate non-answers from candidate answers.
- Separate best answers from inferior answers.

Language:
A Severely NL Transfer Function

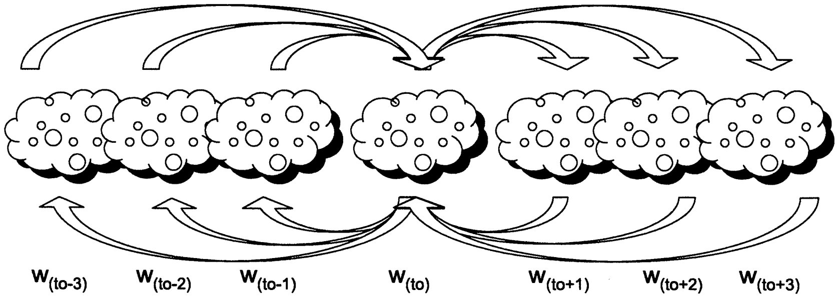
- Language has been formulated by humans, for humans and is quite natural for modeling the dynamics between the thalamus and the cerebral cortex.
- Confabulation (as observed in the thalamocortical dynamic) allows one to capture and predict the performance of very nonlinear transfer functions, by using context (non-uniform historical weighting).

Constructs:
Hierarchical ASN's



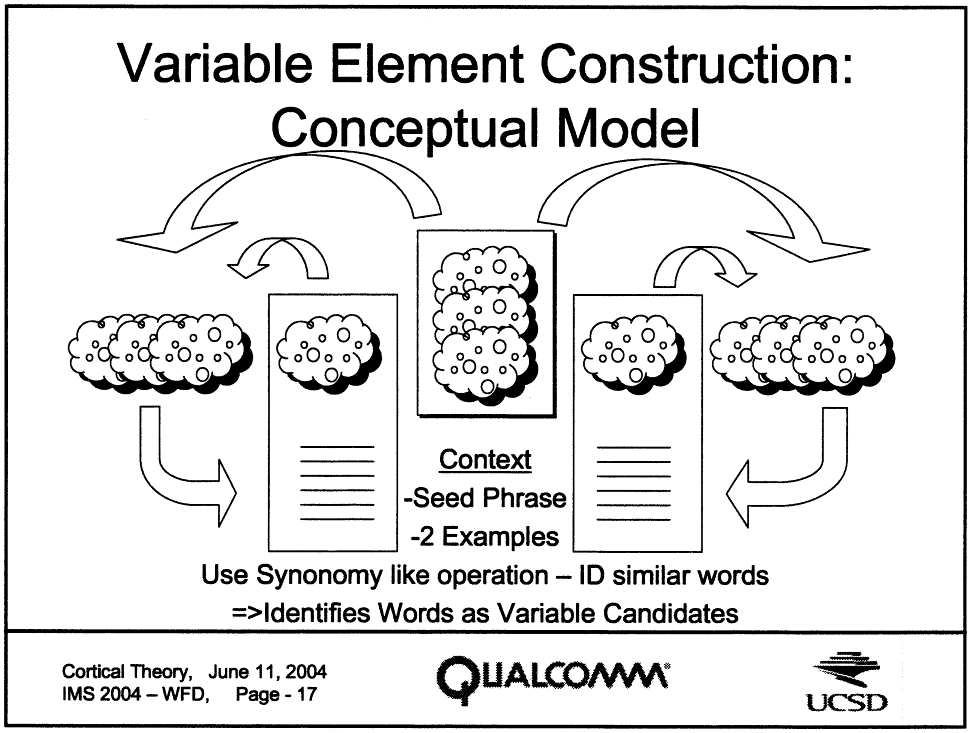
- Identify phrases: Hierarchical ASN's
  - Strong support from flat ASN's
  - Differentiated context from individual words
  - Simplify and clarifies context for future events

Constructs: Synonymy



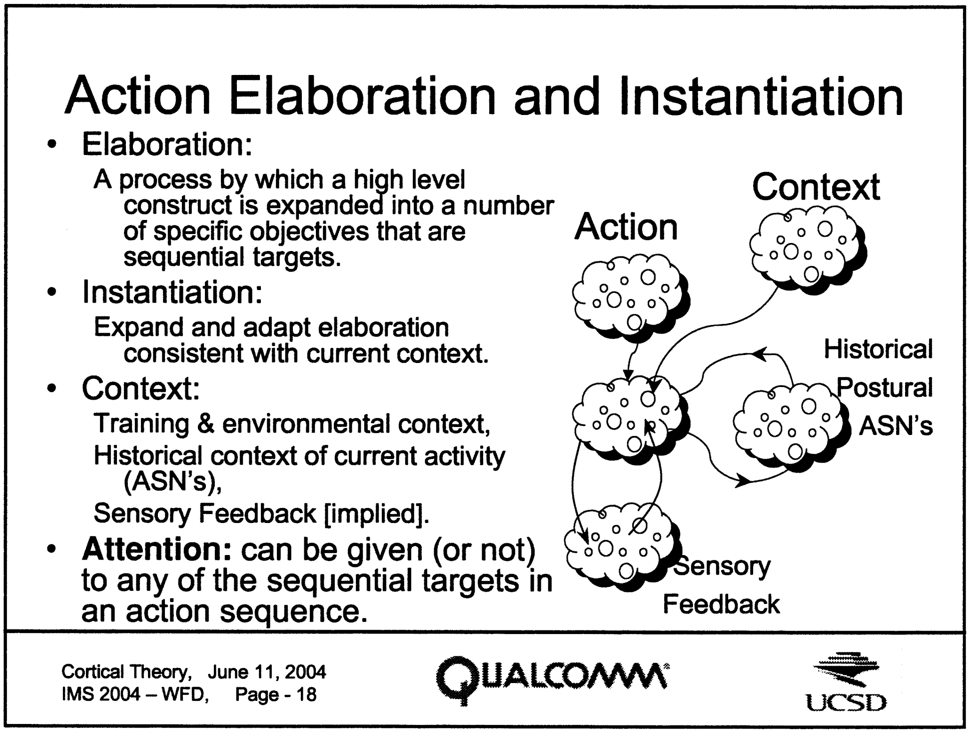
- From an initial word
- Generate lists of common context through outward ASN's
- Use lists and inward ASN's to obtain synonyms.

- Phrase Differentiation
  - Word 1 – context/synonyms
  - Word 2 – context/synonyms
  - Words 1&2
    - strong mutual ASN's
    - very different context/synonyms



Variable Element Construction: Conceptual Model

Use Synonomy like operation – ID similar words
=>Identifies Words as Variable Candidates

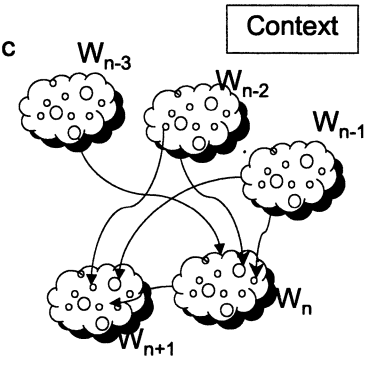


Action Elaboration and Instantiation

- Elaboration:
  A process by which a high level construct is expanded into a number of specific objectives that are sequential targets.
- Instantiation:
  Expand and adapt elaboration consistent with current context.
- Context:
  Training & environmental context,
  Historical context of current activity (ASN's),
  Sensory Feedback [implied].
- **Attention:** can be given (or not) to any of the sequential targets in an action sequence.

Cocktail Party: Segmenting, Interference, and Expectation

- Segmenting:
  An object isolation process where a specific item is either associated with a context or rejected (ignored).
- Interference:
  Multiple, simultaneous item options, of which only one is part of the target.
- Expectation:
  A process of setting pre-established weights on tokens based on context.



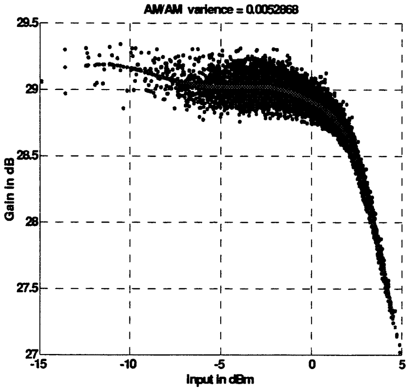
Specific Example (96% accuracy)

lebanon has arrested nine people plotting an attack on the u.s. embassy and kidnapping to try to force the release of islamic militant prisoners

```
05/16/2003 01:35:27 ***************************
05/16/2003 01:35:27 ***** New Simulation *****
05/16/2003 01:35:27 ***************************
05/16/2003 01:35:27 Sentence 0: 24 words
05/16/2003 01:35:27 Sentence 1: 22 words
05/16/2003 01:35:27 Sentence 2: 20 words
05/16/2003 01:35:30 ***************************
05/16/2003 01:35:30 Starting Word: lebanon
05/16/2003 01:35:30 Starting Word: has
05/16/2003 01:35:30 Starting Word: arrested
05/16/2003 01:35:31 Word matched: nine
05/16/2003 01:35:31 Word matched: people
05/16/2003 01:35:31 Word matched: plotting
05/16/2003 01:35:31 Word matched: an
05/16/2003 01:35:32 Word matched: attack
05/16/2003 01:35:32 Word matched: on
05/16/2003 01:35:32 Word matched: the
05/16/2003 01:35:33 Word matched: u.s.
05/16/2003 01:35:33 Word matched: embassy
05/16/2003 01:35:33 Matching FAILED 1: housing
05/16/2003 01:35:33 Word matched: kidnapping
05/16/2003 01:35:33 Word matched: to
05/16/2003 01:35:34 Word matched: try
05/16/2003 01:35:34 Word matched: to
05/16/2003 01:35:34 Word matched: force
05/16/2003 01:35:35 Word matched: the
05/16/2003 01:35:35 Word matched: release
05/16/2003 01:35:35 Word matched: of
05/16/2003 01:35:35 Word matched: islamic
05/16/2003 01:35:36 Word matched: militant
05/16/2003 01:35:36 Word matched: prisoners
05/16/2003 01:35:37 *** DONE ***
```

RF Power Amplifier Problem

- **RF amplifiers requirements today include**
  - **High efficiency for extended battery life,**
  - **Minimized distortion to limit interference**
  - **Minimized spectral regrowth over frequency**
  - **Multimode operations – extreme modulation variations**
- **Modeling only instantaneous distortions (red line) is insufficient to achieve these goals, so accurate modeling of the memory effect (blue dots) is required.**
- **Once one has a sufficient model for these impairments, the next task is to devise ways compensate for them.**



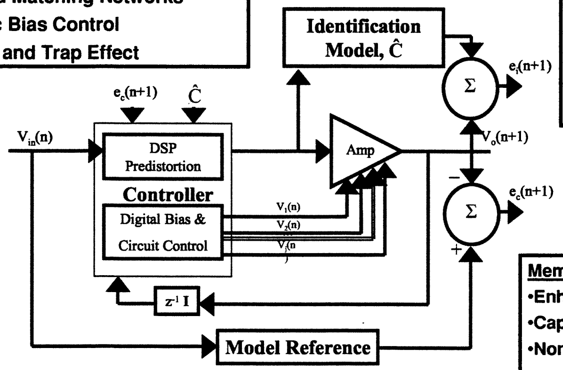
Impact and Reaction to Memory in Wireless Systems

Causes

- Matching and Bias Network Impact
- Switched Matching Networks
- Dynamic Bias Control
- Thermal and Trap Effect

Performance Impact

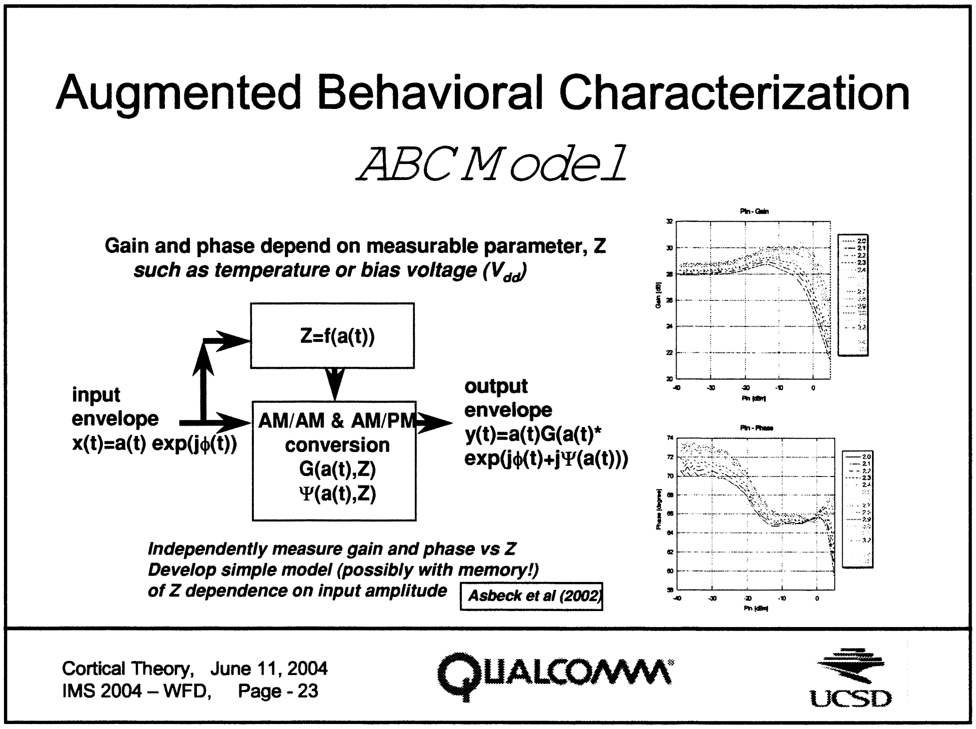
- Increased ACPR
- Degraded EVM
- Limits to Memoryless Predistortion
- Decreased network capacity



Enabling

<u>Memory Compensation</u>
- Enhanced Smart Front Ends
- Capability to Adaptive Controls
- Nonlinear MIMO Control

Similar to Model Reference Adaptive Control, Landau (1979) , Haykin (1999)



Augmented Behavioral Characterization

ABC Model

Gain and phase depend on measurable parameter, Z
such as temperature or bias voltage (V_{dd})

Z=f(a(t))

input
envelope
x(t)=a(t) exp(jϕ(t))

AM/AM & AM/PM
conversion
G(a(t),Z)
Ψ(a(t),Z)

output
envelope
y(t)=a(t)G(a(t)*
exp(jϕ(t)+jΨ(a(t)))

Independently measure gain and phase vs Z
Develop simple model (possibly with memory!)
of Z dependence on input amplitude ***Asbeck et al (2002)***

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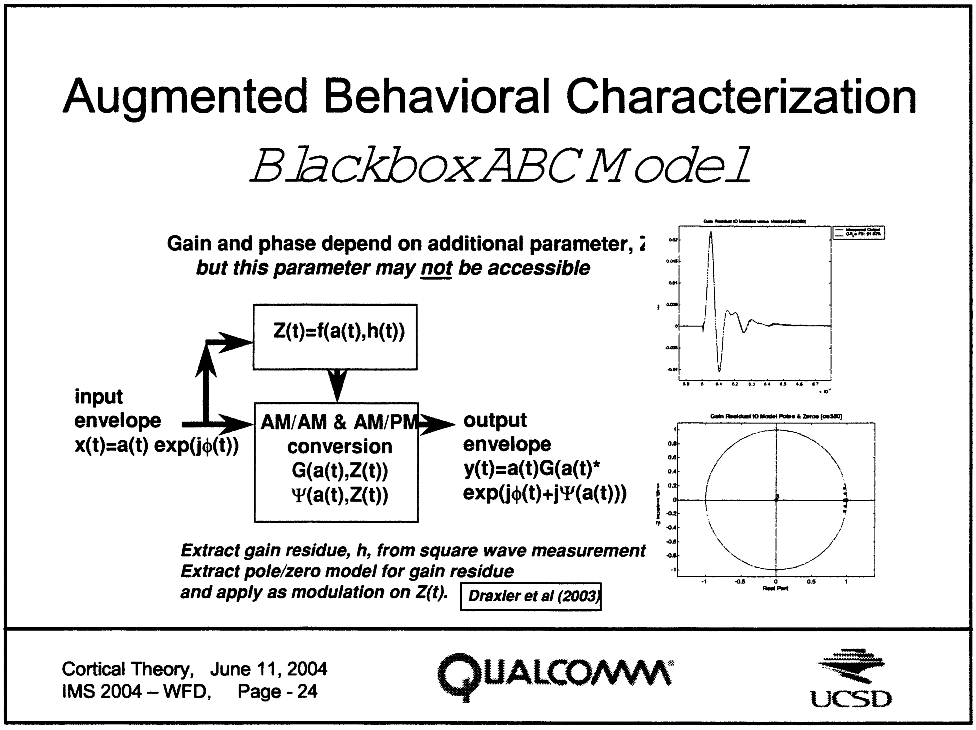

Augmented Behavioral Characterization

Blackbox ABC Model

Gain and phase depend on additional parameter, Z
but this parameter may <u>not</u> be accessible

Z(t)=f(a(t),h(t))

input
envelope
x(t)=a(t) exp(jϕ(t))

AM/AM & AM/PM
conversion
G(a(t),Z(t))
Ψ(a(t),Z(t))

output
envelope
y(t)=a(t)G(a(t)*
exp(jϕ(t)+jΨ(a(t)))

Extract gain residue, h, from square wave measurement
Extract pole/zero model for gain residue
and apply as modulation on Z(t). ***Draxler et al (2003)***

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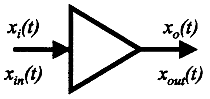
Modeling Theory

Behavioral Model



$$x_i(t) = x_{in}(t)\exp(j(2\pi f_o t + \varphi_{in}(t)))$$
$$x_o(t) = x_{out}(t)\exp(j(2\pi f_o t + \varphi_{out}(t)))$$

Gain dependence on hidden parameter Z

$$\boxed{x_{out}(t) = G(x_{in}(t), Z(t))x_{in}(t)}$$

Approximate with Taylor expansion

$$G(x_{in}, Z(t)) \approx G(x_{in}, Z_\infty) + \left.\frac{\partial G}{\partial Z}\right|_{|x_{in}|, Z_\infty}(Z(t) - Z_\infty(x_{in}))$$

$$G(x_{in}, Z(t)) \approx G_o(x_{in}) \cdot (1 + k_g(x_{in}) \cdot (Z(t) - Z_\infty(x_{in})))$$

Initially, let us assume a linear response of Z to changes of x_{in}

$$\boxed{G(x_{in}, Z(t)) \approx G_o(x_{in}) \cdot (1 + k_g h_z(t) \otimes x_{in}(t))}$$

or

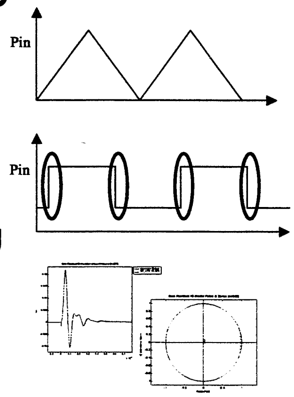
$$\boxed{G(x_{in}, Z(t)) \approx G_o(x_{in}) \cdot (1 + \gamma(x_{in}))}$$

where, $\gamma(x_{in})$ is the gain residue

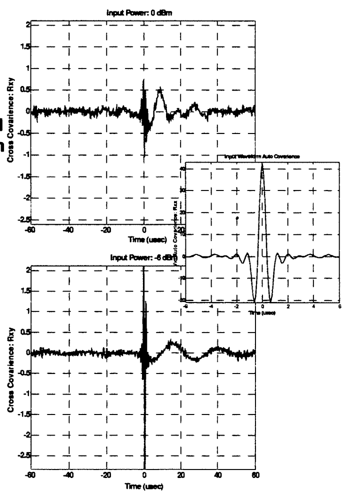
and $G_o(x_{in})$ is the memoryless compression model

Black-box ABC Model: Extraction Methodology for PA with Memory

- Slow - Triangle Waveform
  - AM/AM and AM/PM model
- Square Waveform at various power levels
  - Extract gain residue
- Model gain residue modeled using system identification techniques
- Validate with other waveforms
  - Triangle, Square, Two tone, CDMA, 802.11 and others



CDMA Measurements



The dependence of the gain residue on the signal history shows up as a correlation with the input signal.

Gain residue $$\frac{G(t)-G_o}{G_o} = k_g h(t)\otimes x_{in}(t)$$ **Depends on input signal roughly as linear system**

Input-output cross-correlation theorem: $$\phi_{io}(t) = \int h(\tau)\phi_{ii}(\tau - t)d\tau$$

Crosscorrelation of input and output of linear system is equal to convolution of impulse response of system and input autocorrelation function

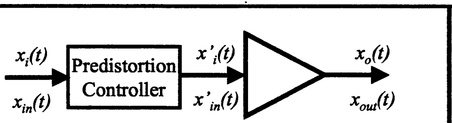
Input Autocorrelation has a very narrow shape (related to FIR filter and baseband LPF).

$\phi_{II}(t)$ is approximately a delta-function.

=> expect $\phi_{Io}(t)$ to approximately equal the impulse response of the gain residue

---

Memoryless and Memory Effect Predistortion – Initial Test



Behavioral Model

We want:

$$x_{out}(t) = G_r\cdot x_{in}(t) = P(x'_{in}(t))*(1+\gamma(x'_{in}(t))$$

Where G_r is a constant, minimal gain of the chain, *P(.)* is the memoryless gain compression model, and $\gamma(.)$ is the gain residue, and,

$$x'_{in}(t) = x_{in}(t)+\delta$$

And if δ is small, and the gain residue is small, we can approximate,

$$P(x'_{in}(t)) = \frac{G_r\cdot x_{in}(t)}{(1+\gamma(x_{in}(t))}$$

We can then define a memoryless predistortion based on an inverted polynomial of the compression characteristic, $P^{-1}(.)$

$$\tilde{x}'_{in}(t) = P^{-1}(G_r\cdot x_{in}(t))$$

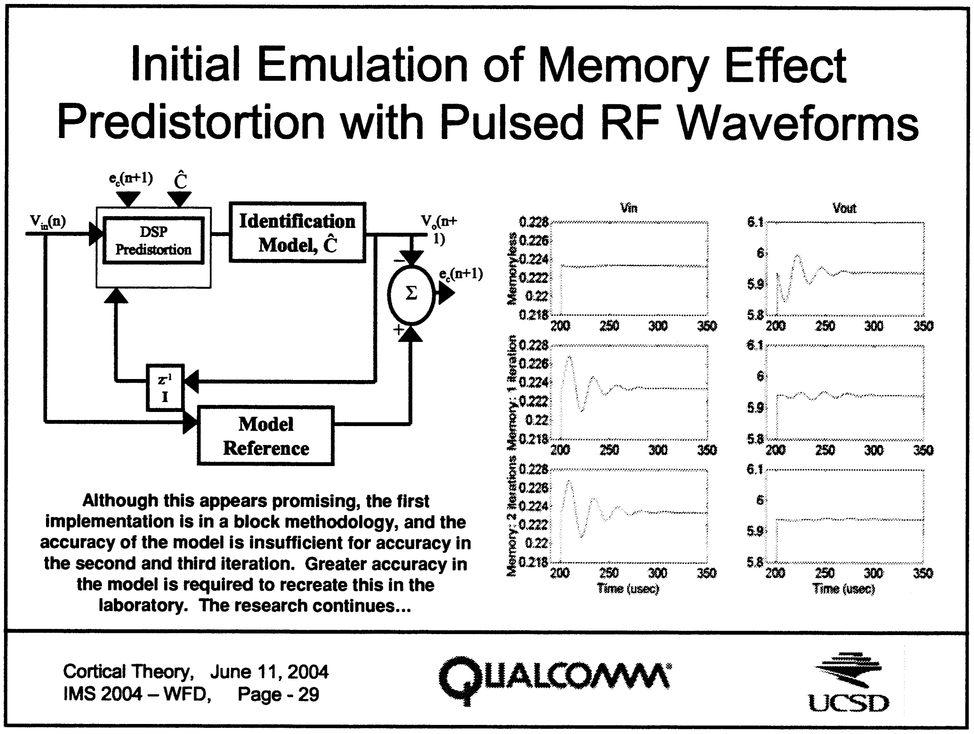
We can then define a first order correction error e_1,

$$e_1 = P(\tilde{x}'_{in}(t))\cdot(1+\gamma(\tilde{x}'_{in}(t))) - G_r\cdot x_{in}(t))$$

$$e_1 = G_r\cdot x_{in}(t)\cdot\left(\frac{\gamma(\Delta)}{1+\gamma(x_{in}(t))}\right),$$

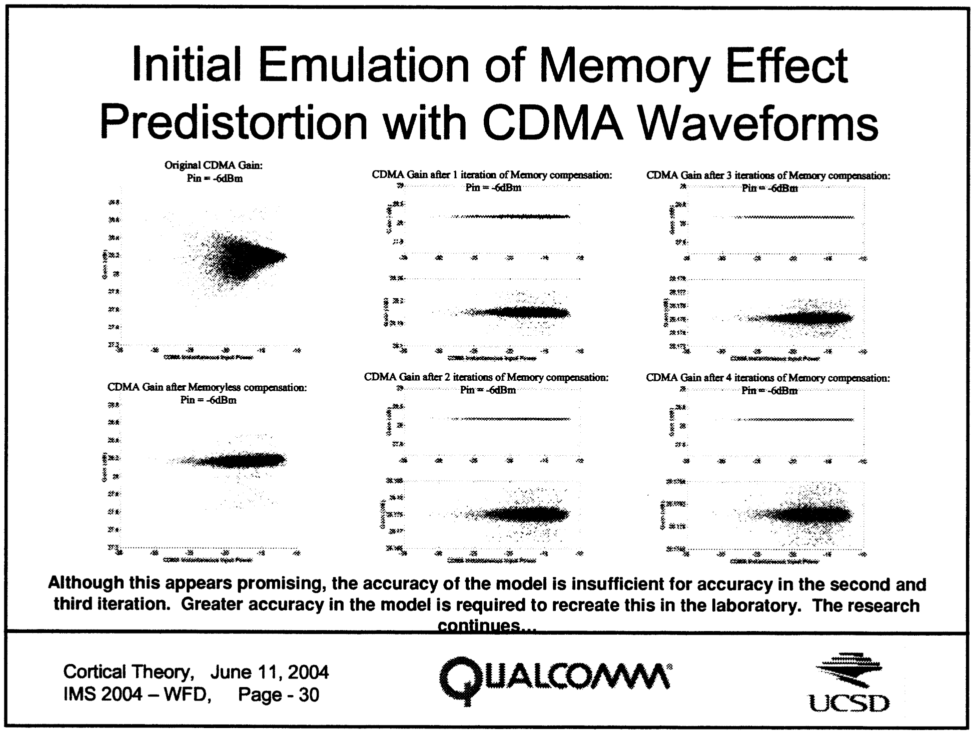
$$\Delta = |x+\delta| - |x|$$

We have applied this to the Blackbox ABC model with favorable results (next few slides) but the actual system is further complicated by a nonlinear gain residue, $\gamma(.)$, and compression characteristic, *P(.)*.



Initial Emulation of Memory Effect Predistortion with Pulsed RF Waveforms

Although this appears promising, the first implementation is in a block methodology, and the accuracy of the model is insufficient for accuracy in the second and third iteration. Greater accuracy in the model is required to recreate this in the laboratory. The research continues...



Initial Emulation of Memory Effect Predistortion with CDMA Waveforms

Although this appears promising, the accuracy of the model is insufficient for accuracy in the second and third iteration. Greater accuracy in the model is required to recreate this in the laboratory. The research continues...

DSP Motivation and Context

- Corrective structures can be classified as block based or adaptive.
- Adaptive systems adapt faster, more agile, and efficient than block based algorithms.
- Standard adaptive filter structures are generally very good at correcting for stationary, or slowly varying signal contamination, but not fast enough for NL transfer functions.
- Controls to correct for nonlinear signal contamination require a nonlinear adaptive system – adaptive time constants and wide eigen value spreads are common issues.

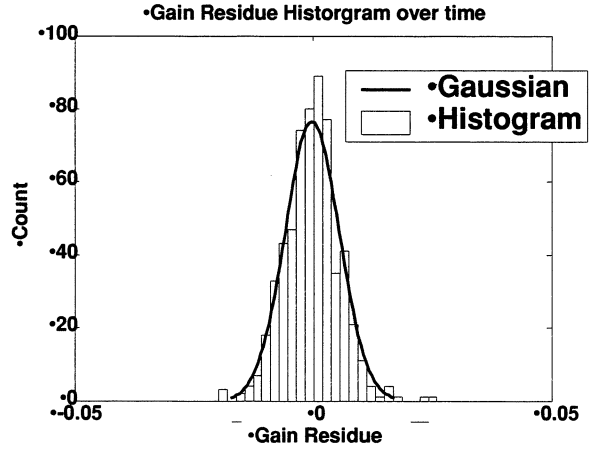
Cortical Theory Additions

Define discrete input and output values.
Input: |Vin| and ΔVin
Output: gain residue

Build pairwise knowledge bases (try Gaussian curvefit to histograms).

Define predicted context based on Blackbox ABC model, previous inputs to RF Power Amplifier and generate the most likely mean and standard deviation for Gain Residue.

One RF PA Knowledge Base



Conclusions

Confabulation is based on observations and insight of mammalian Thalamocortical dynamics.

Language transfer functions are severely nonlinear, leading one to conclude that confabulation will work with simpler control systems.

Complex digital modulation, combined with solid state electronics have created new multi-disciplinary problems for RF systems.

By exploring how other communities have solved similar problems and by applying new approaches, we can further improve channel, power, and component efficiency of RF transceivers.

Acknowledgements

QUALCOMM Inc.

Prof. Peter M. Asbeck
Prof. Robert Hecht-Nielsen
Ian Langmore
Tsai-Pi Hung

References

Cortical Theory

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RF Power Amplifier Behavioral Modeling

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Simulation Tools

[8] Matlab, R12, Student Version, 2002.

Measurement-Based ANN Modeling in the Frequency-Domain

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2004 IEEE MTT-S International Microwave Symposium
Advances and New Directions in Device Modeling and Design
Optimization for Microwave CAD

Fort Worth, TX – June 11, 2004

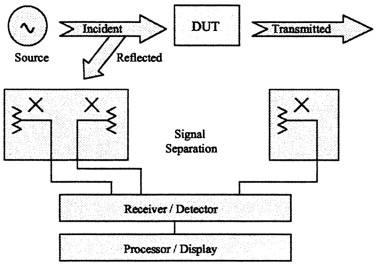
Presentation Outline

★ ANN Modeling for Improved Vector Network Analyzer Calibrations

- Developing Frequency-Domain Models for Nonlinear Circuits Based on Large-Signal Measurements

Vector Network Analyzers

- VNAs are one of the most versatile instruments in the RF and microwave industry.
- Used to measure complex scattering parameters of device and circuits.
- Engineers use them to verify designs, confirm proper performance, and diagnose problems.



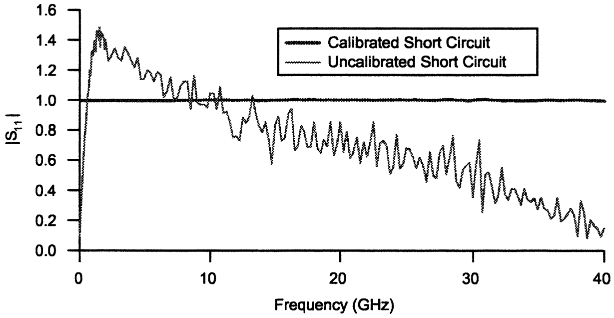
Importance of Calibration

Calibration accounts for:

- Directivity and crosstalk errors due to signal leakage.
- Impedance mismatches.
- Frequency response errors in the source and receiver.
- Losses in the cables and connectors.

Calibration doesn't account for:

- System drift.
- Repeatability in the switches and connectors.
- Instrument noise.
- Errors in calibration standards.



Vector Network Analyzer Calibrations

Procedure:

- Measure known standards.
- Process data to determine error coefficients.
- Correct measured data for device under test.

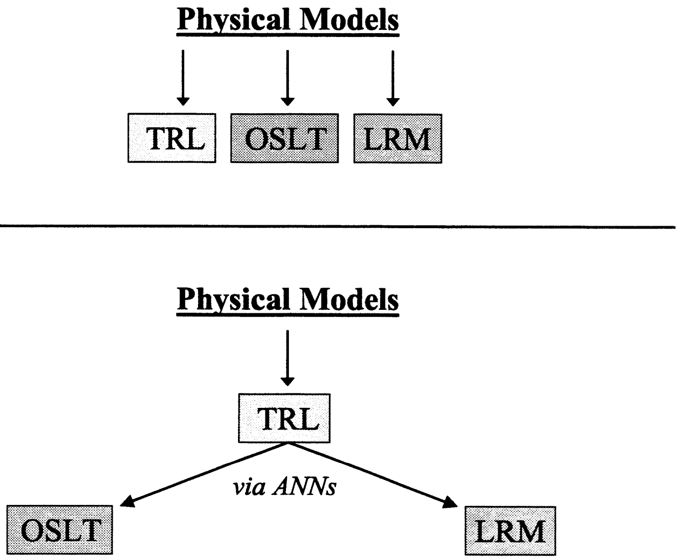
Methods:

Lumped-element standards (OSLT, LRM).
Transmission line standards (TRL, Multiline TRL).
Electronic transfer standards.

Motivations

- Multiline TRL highly accurate means of VNA calibration but it does have disadvantages, namely standards require a lot of real estate and changing separation of probes is necessary.
- Compact calibration kits, such as LRM and OSLT, are preferred on-wafer, but these also have disadvantages, namely lumped-element artifacts tend to be less accurate.
- If standards can be characterized using a benchmark calibration, it is possible to perform an accurate on-wafer LRM or OSLT calibration.
- Once the standards are characterized, one must decide whether to develop a model for them or directly use measurement data obtained from benchmark calibration.

Calibration Standard Philosophies



Physical Models

TRL OSLT LRM

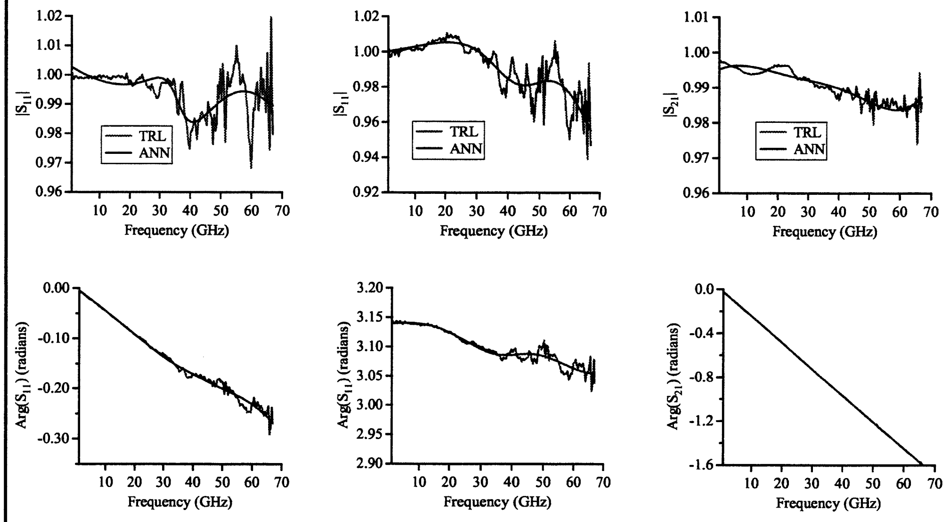
Physical Models

TRL

via ANNs

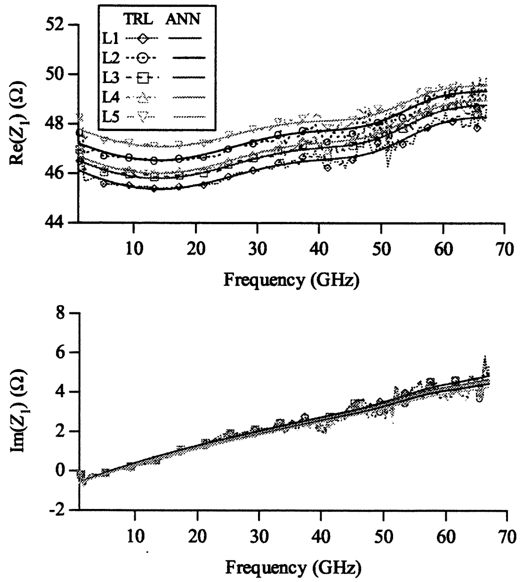
OSLT LRM

On-Wafer OSLT – Open, Short, and Thru Standards



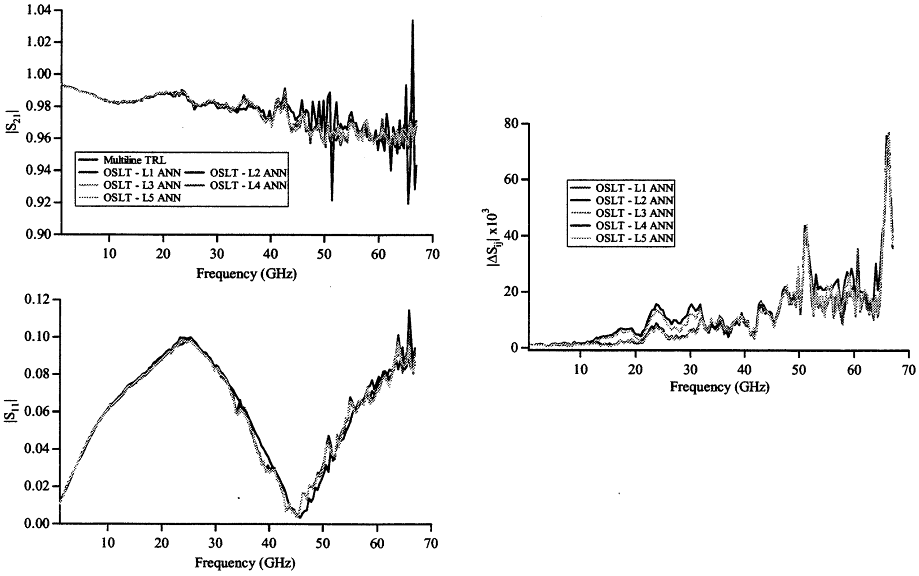
Open **Short** **Thru**

On-Wafer OSLT – Varying Loads

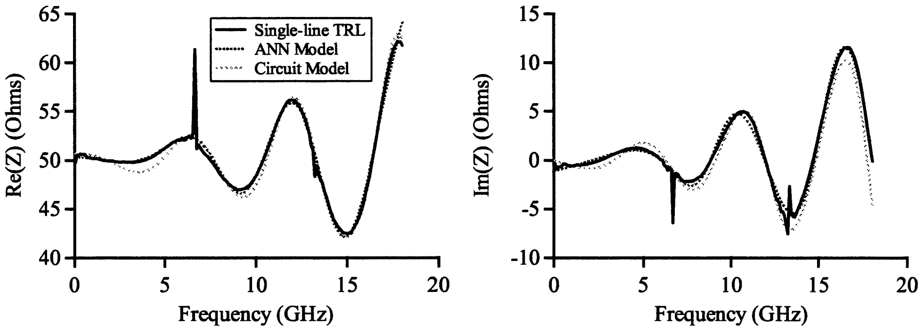


Load	DC Resistance (Ω) Port 1
1	44.73
2	45.85
3	45.20
4	45.38
5	46.45

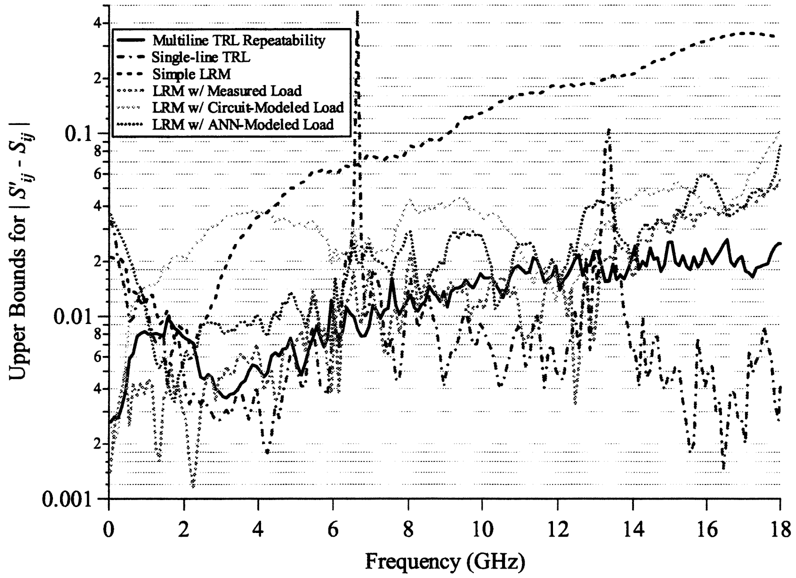
On-Wafer OSLT – Calibration Comparisons



Coaxial LRM – Match



Coaxial LRM – Calibration Comparisons



Summary

- We have successfully applied ANNs to model on-wafer and coaxial lumped-element calibration standards.
- We have shown that calibrations that make use of ANN-modeled standards compare favorably to benchmark TRL calibrations.
- ANN-modeled calibration standards can be implemented using existing or custom software packages.

  - MultiCal: TRL and LRM calibrations
  - NeuroModeler: ANN model development
  - Custom: OSLT calibrations

Advantages of ANN Models

- They do not require detailed physical models.
- ANN models are much more compact than large measurement files (62 real-valued parameters for S_{11} vs. 2475 for measurement database of just 5 loads).
- Possible to generate accurate models using only a few measurement points (9 of the 165 available points). Thus, calibration times can be reduced.
- ANN models, trained on only a few measurement points, can be much more accurate than direct calibrations when limited data are available.
- They are less susceptible to the noise inherent in measured data.
- ANN models are able to accurately model loads with measured DC resistances slightly outside their training range.

Presentation Outline

- ANN Modeling for Improved Vector Network Analyzer Calibrations
- ★ Developing Frequency-Domain Models for Nonlinear Circuits Based on Large-Signal Measurements

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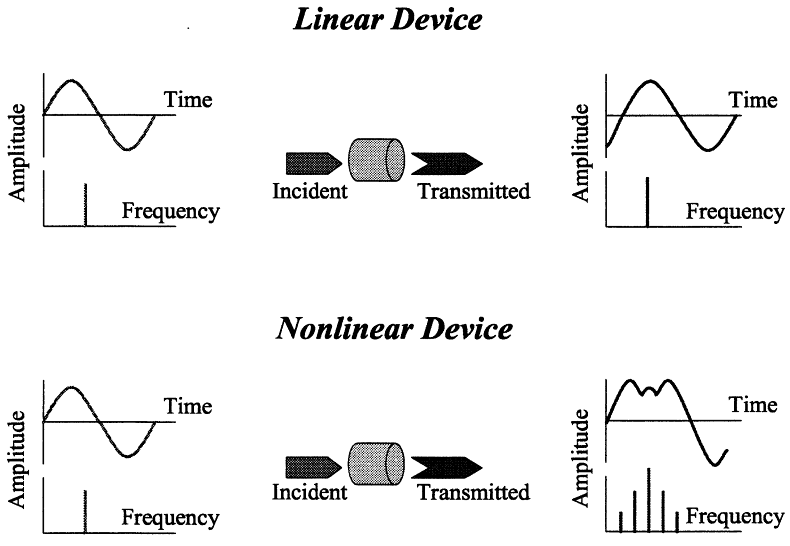
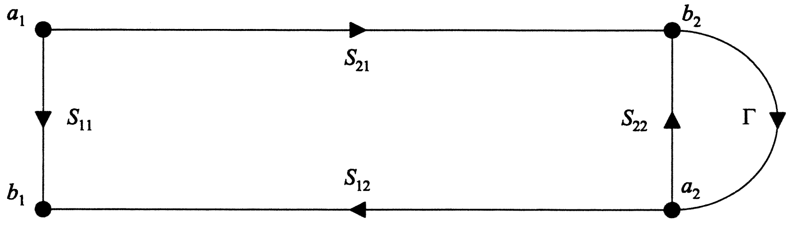
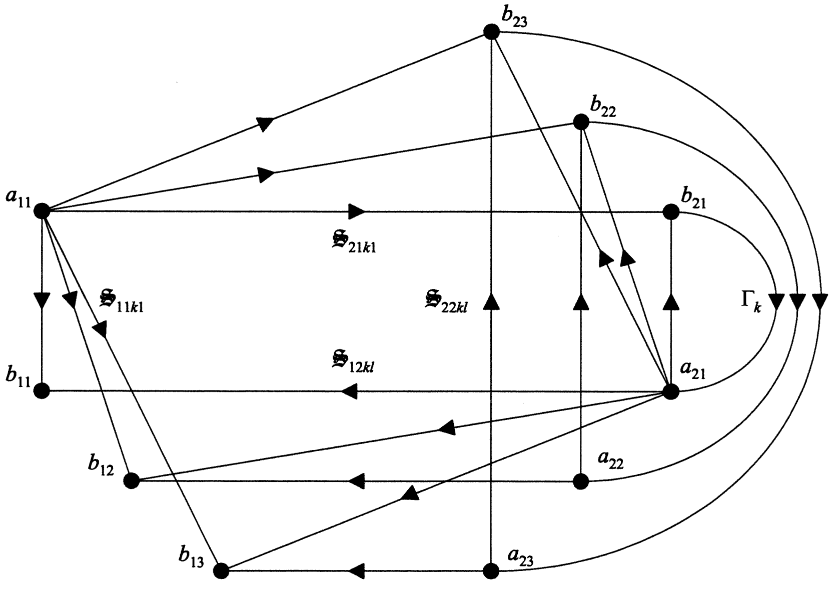
Linear Versus Nonlinear Device



Flow Diagram for a Linear Two-Port Device

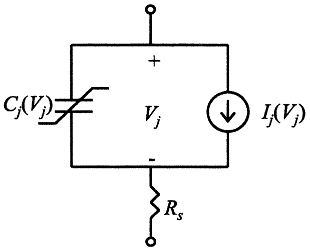


Flow Diagram for a Nonlinear Two-Port Device



Common Approaches to Nonlinear Device Modeling

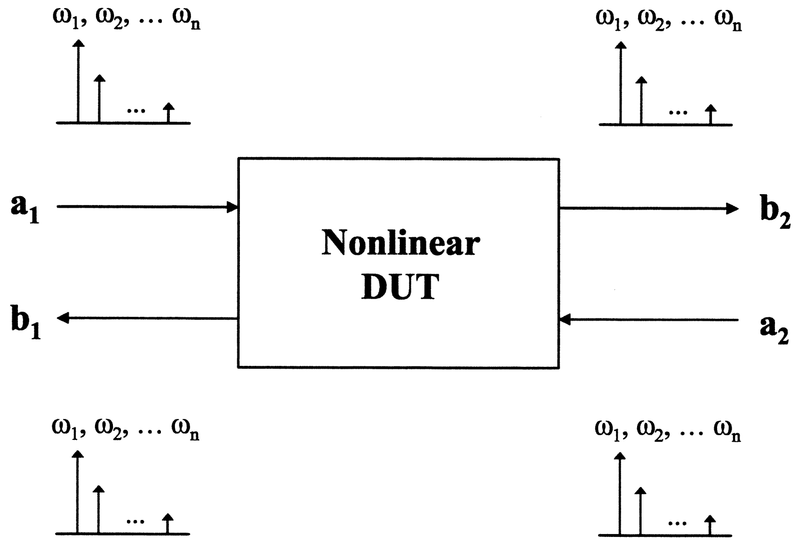
- **Physical Models**
  - Electrical behavior from basic parameters (dimensions and material characteristics).
  - PDEs and carrier transport physics in semiconductors.
  - Not used much for circuit design.
- **Equivalent Circuit Models**
  - Linear and nonlinear circuit elements in a configuration that emulates the device.
  - Circuit elements often associated with the physical structure of the device.
  - Most widely used approach.



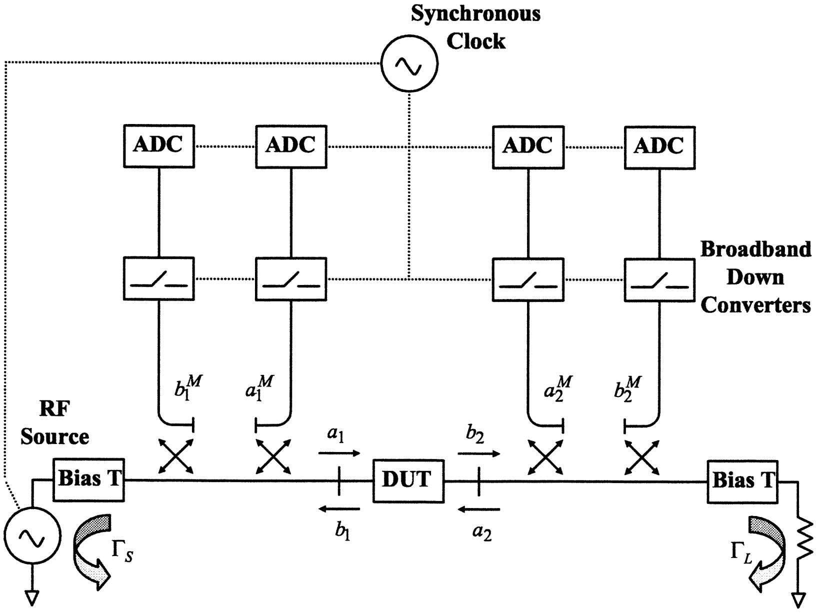
Newer Approach to Nonlinear Device Modeling

- Measurement-Based Frequency Domain Models
  - Instrumentation (NVNA's) now available for accurate nonlinear RF characterization.
  - Device physics not needed for model derivation.
  - Neural network modeling appropriate.
  - "Traceable" models.
  - Especially useful when de-embedding is difficult, components have distributed characteristics, and with new less-understood technology.

Power-Normalized Waves at the Ports of a Nonlinear Device



Block Diagram of a Nonlinear Vector Network Analyzer



Highlights of NVNAs

- Ability to measure a nonlinear device in realistic large-signal operating conditions.

- Allow accurate measurements of amplitude and phase of the spectral components of incident and reflected voltage waves at the ports of a nonlinear device.

Nonlinear Large-Signal $\mathfrak{S}$-Parmeters

Two-port network with a single-tone excitation.

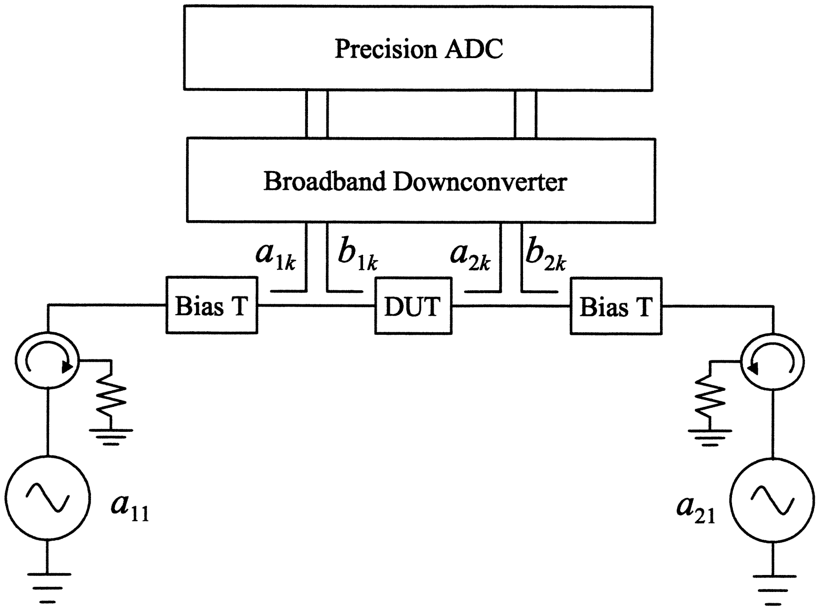
Input reflection coefficient:

$$\mathfrak{S}_{11k1} = \left.\frac{|b_{1k}|\angle\left(\phi_{b_{1k}} - k\phi_{a_{11}}\right)}{|a_{11}|}\right|_{a_{mn} = 0 \text{ for } \forall m \forall n[(m \neq 1) \wedge (n \neq 1)]}$$

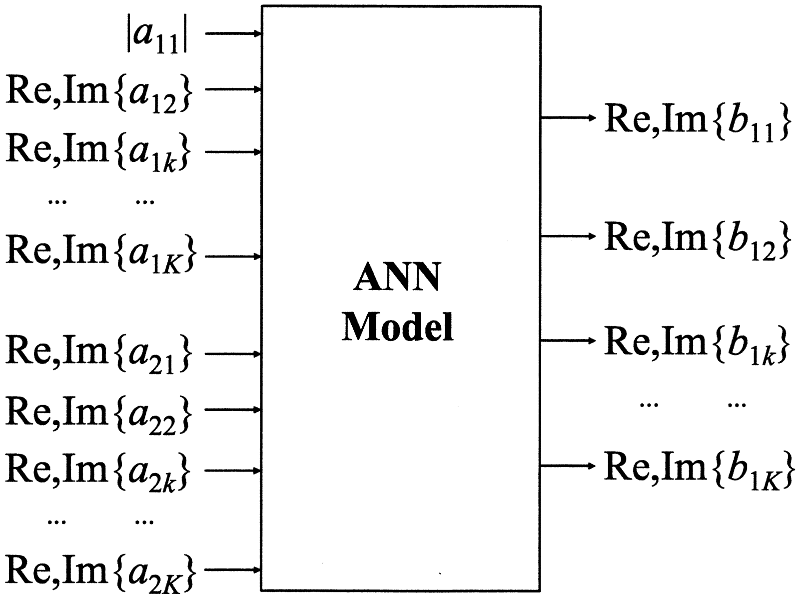
Forward transmission coefficient:

$$\mathfrak{S}_{21k1} = \left.\frac{|b_{2k}|\angle\left(\phi_{b_{2k}} - k\phi_{a_{11}}\right)}{|a_{11}|}\right|_{a_{mn} = 0 \text{ for } \forall m \forall n[(m \neq 1) \wedge (n \neq 1)]}$$

NVNA with Two Sources



Inputs & Outputs for Training ANN Model


$|a_{11}|$

$\mathrm{Re},\mathrm{Im}\{a_{12}\}$

$\mathrm{Re},\mathrm{Im}\{a_{1k}\}$

… …

$\mathrm{Re},\mathrm{Im}\{a_{1K}\}$

$\mathrm{Re},\mathrm{Im}\{a_{21}\}$

$\mathrm{Re},\mathrm{Im}\{a_{22}\}$

$\mathrm{Re},\mathrm{Im}\{a_{2k}\}$

… …

$\mathrm{Re},\mathrm{Im}\{a_{2K}\}$

ANN Model

$\mathrm{Re},\mathrm{Im}\{b_{11}\}$

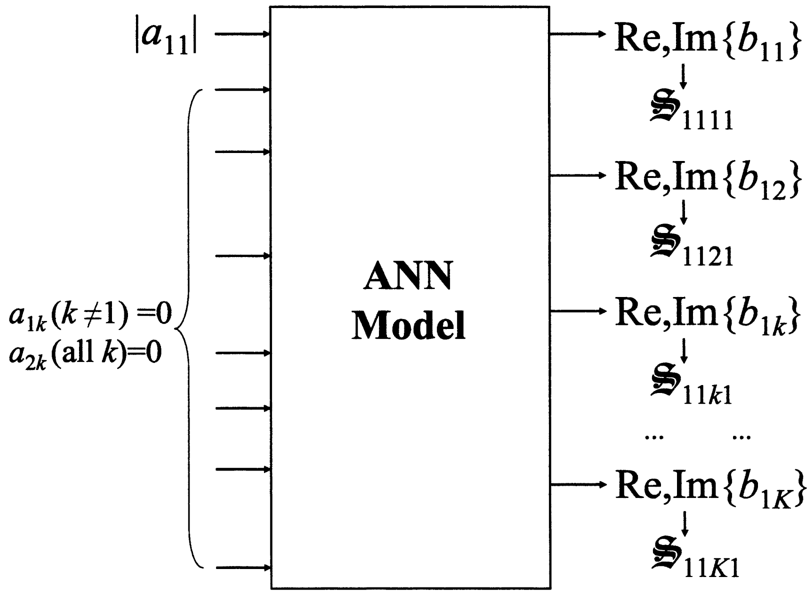
$\mathrm{Re},\mathrm{Im}\{b_{12}\}$

$\mathrm{Re},\mathrm{Im}\{b_{1k}\}$

… …

$\mathrm{Re},\mathrm{Im}\{b_{1K}\}$

Inputs & Outputs for Obtaining $\mathfrak{S}_{11k1}$



$|a_{11}|$

$a_{1k}(k \neq 1) = 0$
$a_{2k}(\text{all } k) = 0$

ANN Model

$\text{Re,Im}\{b_{11}\} \rightarrow \mathfrak{S}_{1111}$

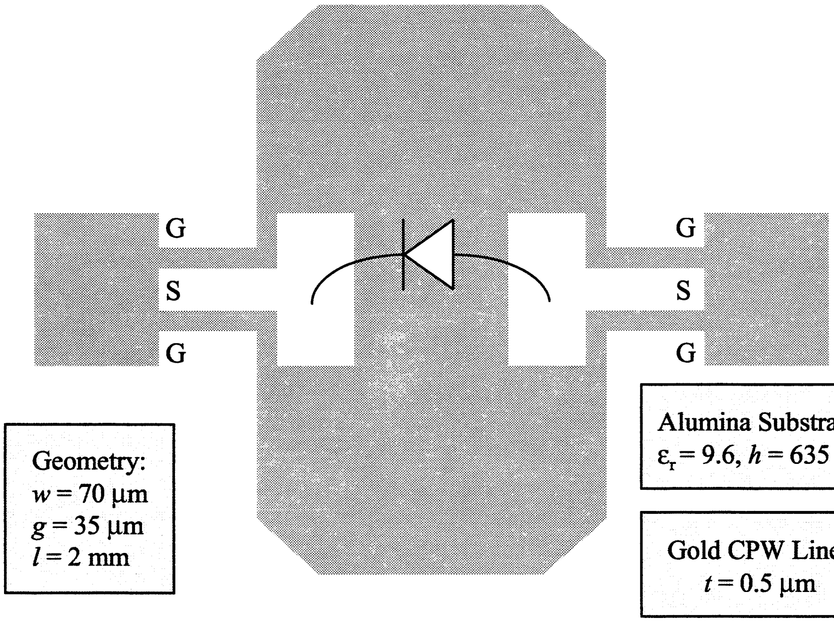
$\text{Re,Im}\{b_{12}\} \rightarrow \mathfrak{S}_{1121}$

$\text{Re,Im}\{b_{1k}\} \rightarrow \mathfrak{S}_{11k1}$

… …

$\text{Re,Im}\{b_{1K}\} \rightarrow \mathfrak{S}_{11K1}$

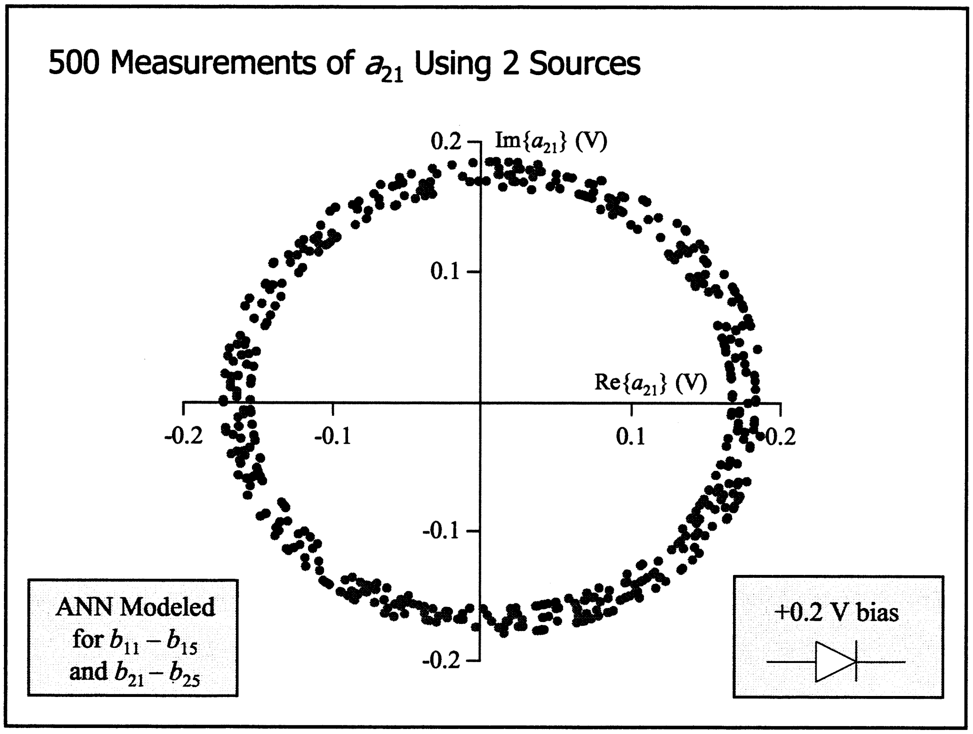
Schottky Diode in Series Configuration in CPW



Geometry:
$w = 70\ \mu\text{m}$
$g = 35\ \mu\text{m}$
$l = 2$ mm

Alumina Substrate:
$\varepsilon_r = 9.6$, $h = 635\ \mu\text{m}$

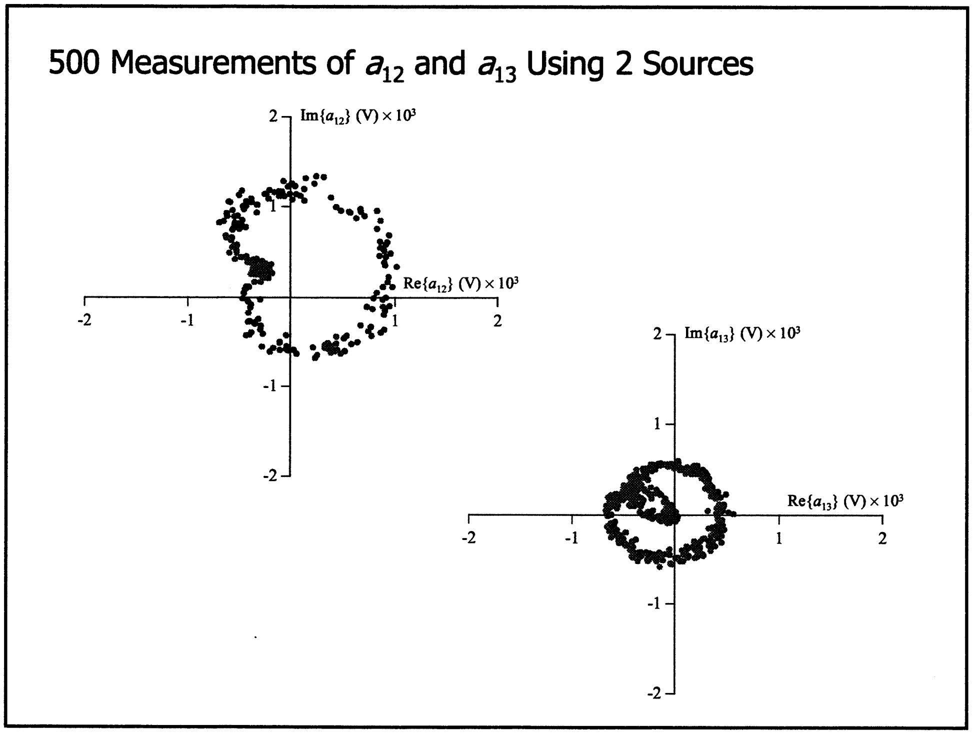
Gold CPW Lines:
$t = 0.5\ \mu\text{m}$



500 Measurements of a_{21} Using 2 Sources

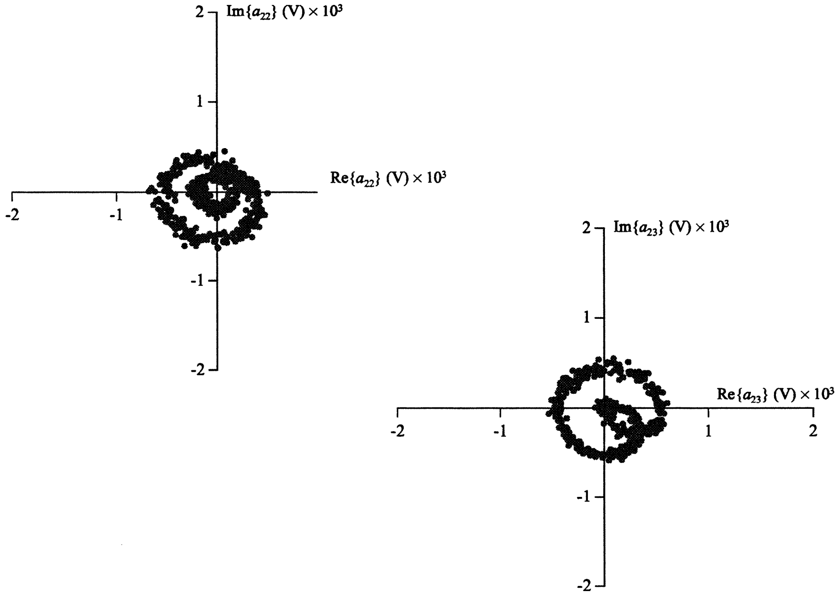
ANN Modeled
for $b_{11} - b_{15}$
and $b_{21} - b_{25}$

+0.2 V bias

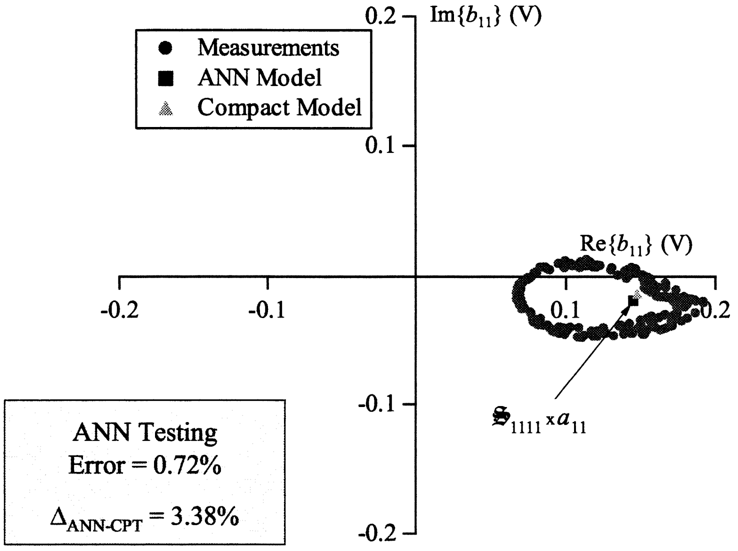


500 Measurements of a_{12} and a_{13} Using 2 Sources

500 Measurements of a_{22} and a_{23} Using 2 Sources



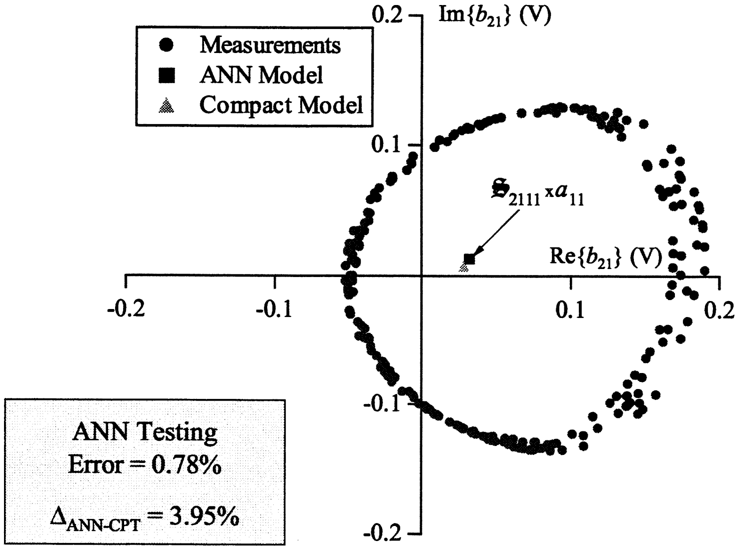
Determining $\mathfrak{S}_{1111}$



ANN Testing
Error = 0.72%

$\Delta_{\text{ANN-CPT}} = 3.38\%$

Determining $\mathfrak{S}_{2111}$



ANN Testing
Error = 0.78%

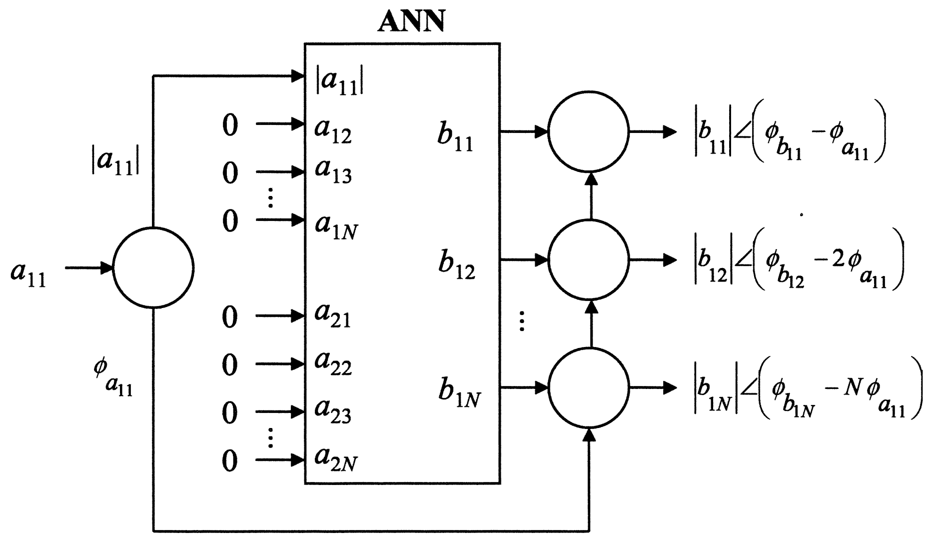
$\Delta_{\text{ANN-CPT}} = 3.95\%$

Determining $\mathfrak{S}_{11k1}$ and $\mathfrak{S}_{21k1}$

Qty.	Diff. (%)	Diff. (dBV)
$\mathfrak{S}_{1111}$	3.38	-44.5
$\mathfrak{S}_{1121}$	1.23	-53.3
$\mathfrak{S}_{1131}$	3.29	-44.8
$\mathfrak{S}_{1141}$	0.40	-63.1
$\mathfrak{S}_{1151}$	1.67	-50.6
$\mathfrak{S}_{2111}$	3.95	-43.2
$\mathfrak{S}_{2121}$	7.15	-38.0
$\mathfrak{S}_{2131}$	5.93	-39.6
$\mathfrak{S}_{2141}$	0.72	-57.9
$\mathfrak{S}_{2151}$	0.85	-56.5

Compact model in commercial CAD software is close to the measurement-based results.

Using ANN Models When the Phase of $a_{11} \neq 0$



Applications of Nonlinear Large-Signal $\mathfrak{S}$-Parameters

- Values of $\mathfrak{S}_{11kl}$ and $\mathfrak{S}_{21kl}$ may be obtained for chosen dc bias and drive levels ($|a_{11}|$).
- Source impedance required for conjugate match (maximum power transfer) at the excitation frequency may be directly obtained from $\mathfrak{S}_{1111}$ at the large-signal operating point.
- Neural networks may be trained for yielding $\mathfrak{S}_{11kl}$ and $\mathfrak{S}_{21kl}$ as a function of bias level and excitation amplitude over a pre-selected range of parameters.
- Similarly the quasi-linear $\mathfrak{S}_{2211}$ is useful for selecting the optimum load impedance for output at the fundamental frequency in a power amplifier.
- Values of $\mathfrak{S}_{ijkl}$ may also be obtained not only for a specified input power level (a_{11}) but also for specified terminations at various harmonics (a_{2k}/b_{2k}).

New directions in addressing complexity of high-speed circuits and interconnects

Michel Nakhla

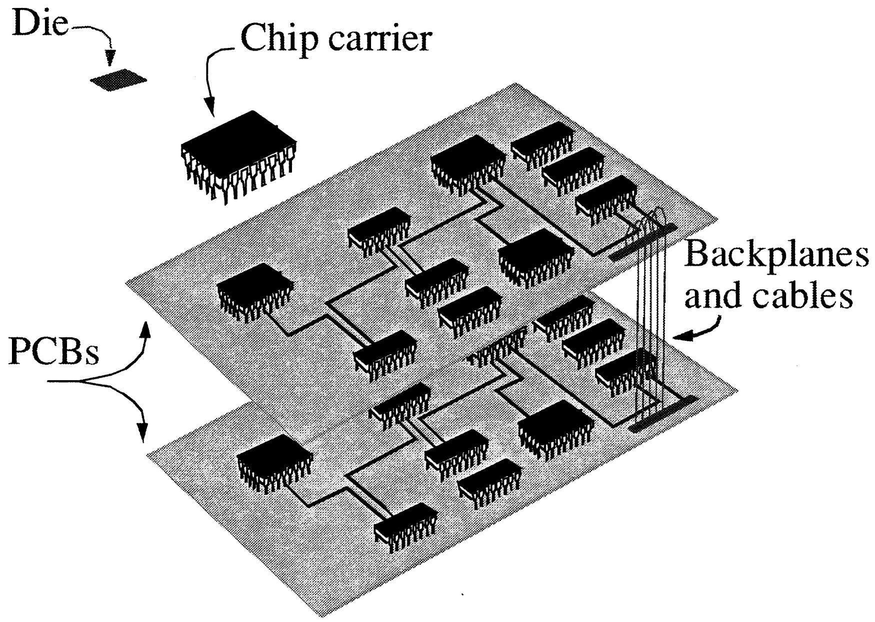
Carleton University

Canada

IMS2004-workshop on :

Advances and New Directions in Device Modeling and Design Optimization for Microwave CAD

Interconnect Hierarchy



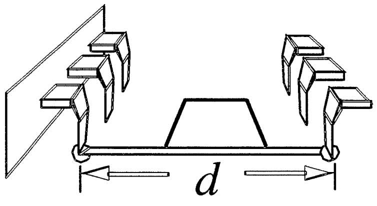
Die

Chip carrier

Backplanes and cables

PCBs

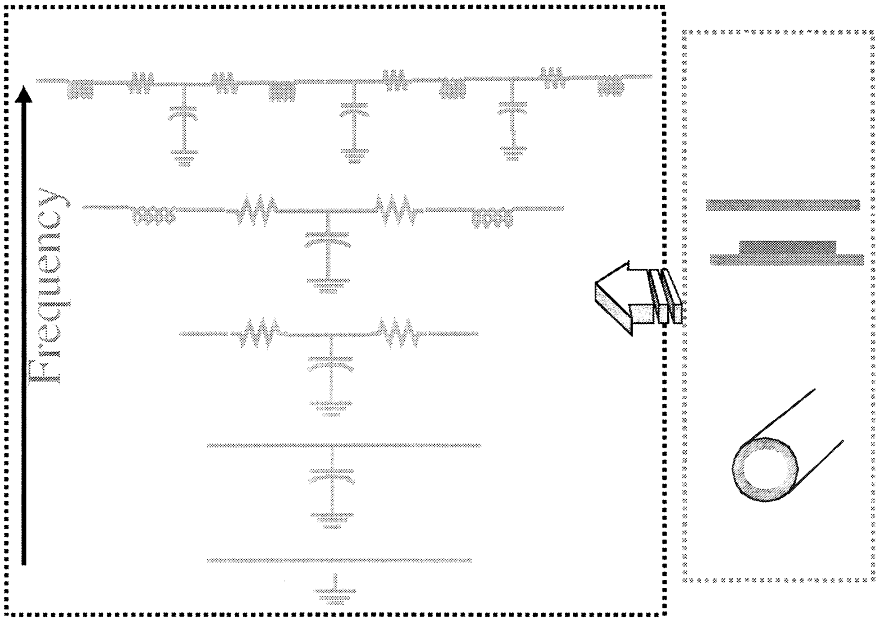
High-Speed Interconnect Effects



Interconnect length becomes comparable to the Wavelength ➡ $\Downarrow \lambda = \frac{v}{f \Uparrow} \approx d$

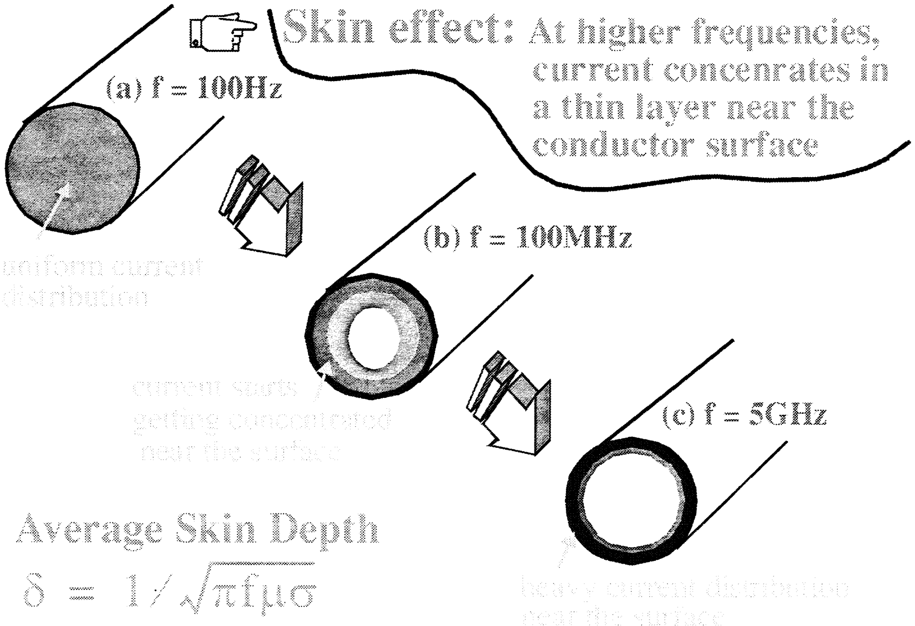
Sharper pulses contain higher frequency harmonics ➡ $\Uparrow f_{max} = \frac{0.35}{t_r \Downarrow}$

Interconnect Models



Interconnect Models

☞ **Skin effect:** At higher frequencies, current concenrates in a thin layer near the conductor surface



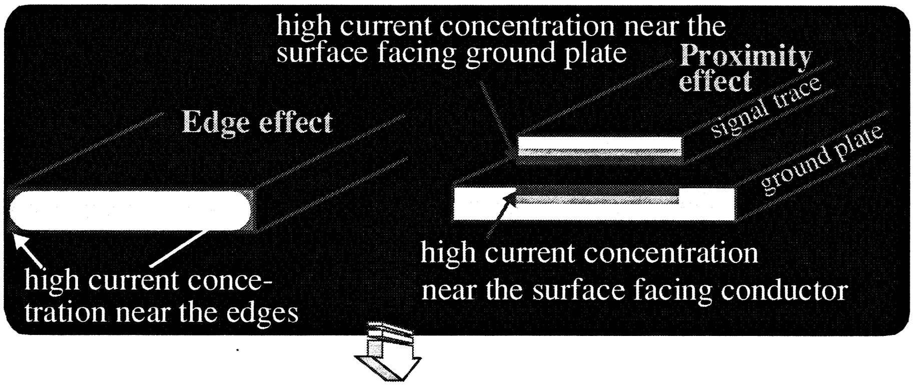
Average Skin Depth

$$\delta = 1/\sqrt{\pi f \mu \sigma}$$

Interconnect Models

☞ **Edge Effect:** current concenrates near the sharp edges of the conductor

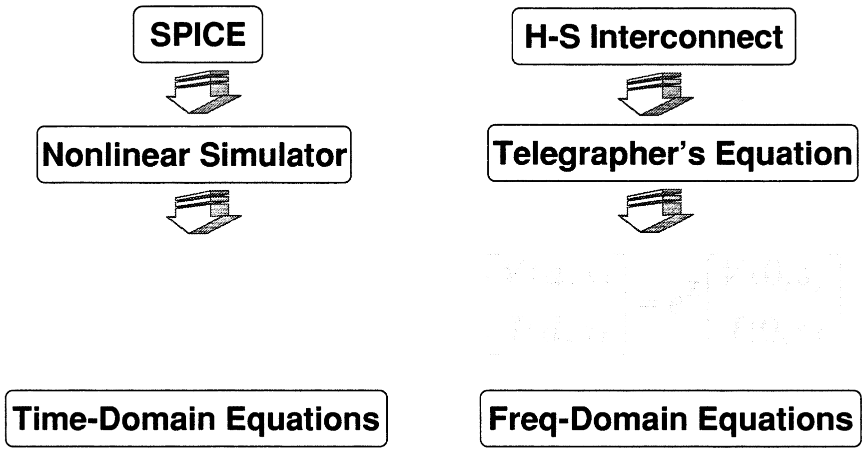
☞ **Proximity Effect:** current in the ground plane concentrates under the signal conductor



Frquency dependent RLCG parameters

Analysis Difficulties

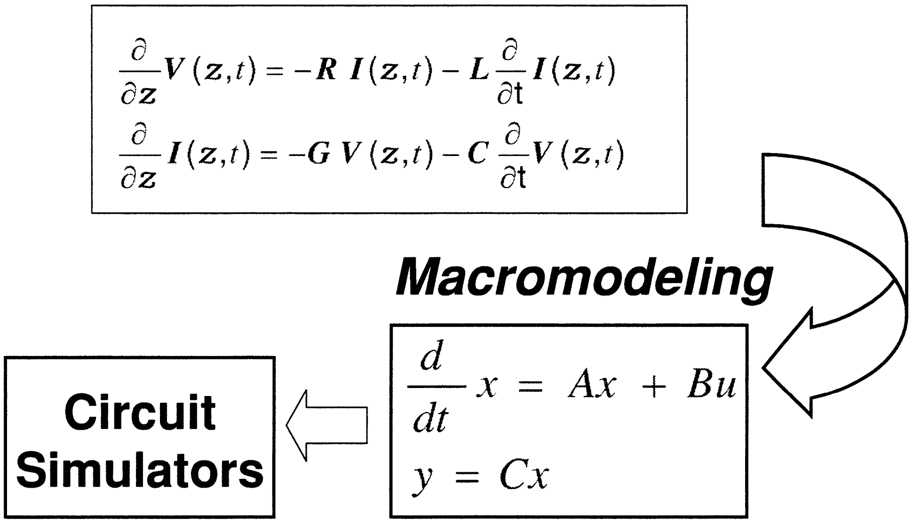
- **Mixed Frequency/Time Problem**



SPICE → Nonlinear Simulator → Time-Domain Equations

H-S Interconnect → Telegrapher's Equation → Freq-Domain Equations

MACROMODELLING



$$\frac{\partial}{\partial z}V(z,t) = -R\,I(z,t) - L\frac{\partial}{\partial t}I(z,t)$$

$$\frac{\partial}{\partial z}I(z,t) = -G\,V(z,t) - C\frac{\partial}{\partial t}V(z,t)$$

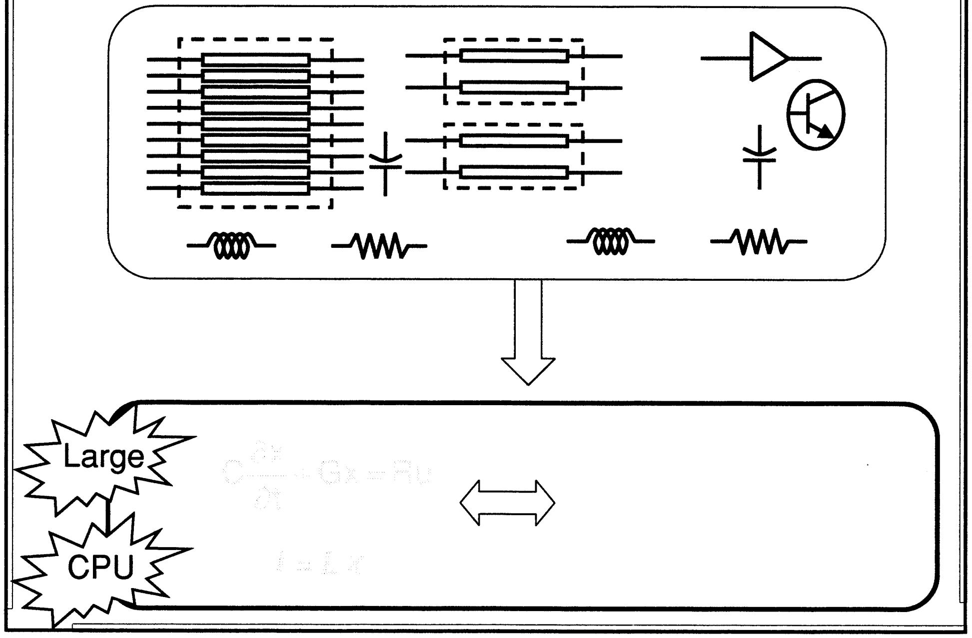
Macromodeling

$$\frac{d}{dt}x = Ax + Bu$$

$$y = Cx$$

Circuit Simulators

Discretization



$$C\frac{\partial x}{\partial t} + Gx = Bu$$

$$i = Lx$$

Solution is based on two concepts:

1) Macromodelling

2) Circuit Reduction

MACROMODELLING ISSUES

1) Accuracy/Efficiency

2) Stability

3) Passivity

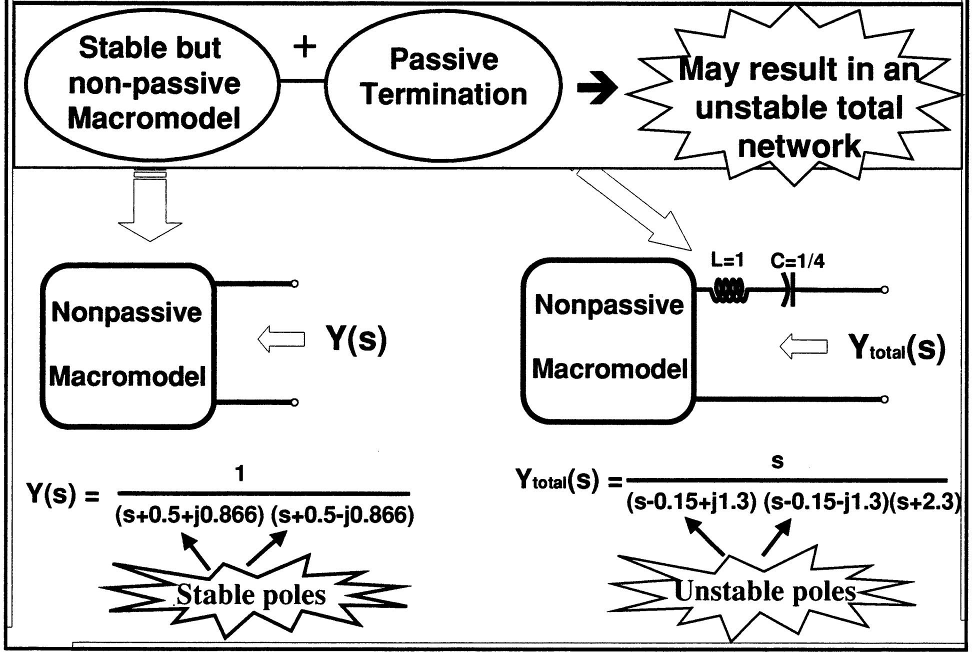
Importance of Passivity

➢ Passivity vs Stability

Stable Macromodel + Passive Termination ➔ May result in an unstable total network

Passive Macromodel + Passive Termination ➔ Asymptotically stable

Significance of Passivity



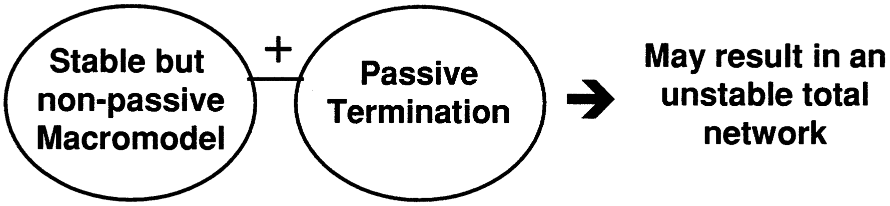
$$Y(s) = \frac{1}{(s+0.5+j0.866)\,(s+0.5-j0.866)}$$

Stable poles

$$Y_{total}(s) = \frac{s}{(s-0.15+j1.3)\,(s-0.15-j1.3)(s+2.3)}$$

Unstable poles

Importance of Passivity



- •Error: Failure to Converge
- •Error: Time Step Too Small – Abort
- •Error: Unstable Network

Passivity Conditions

Passive Macromodel + Passive Termination ➔ Asymptotically stable

Y(s) is passive iff: 1)

2) Re(s)>0

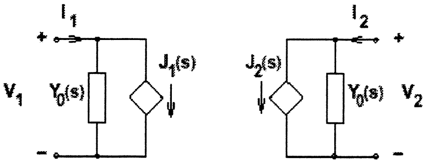
Y(s) is a positive real matrix

Possible Approaches

1) MoC based Algorithms

2) Rational Approximations

Equivalent Circuit : Lossless Line



Basic equivalent circuit for the method of characteristics (MoC) model

$$\Gamma = s\sqrt{LC} \qquad Y_0 = \sqrt{\frac{C}{L}}$$

$$J_1 = -e^{-d\Gamma}[Y_0V_2 + I_2] \qquad \Rightarrow \qquad J_1(t) = -Y_0v_2(t-\tau) - i_2(t-\tau)$$

$$J_2 = -e^{-d\Gamma}[Y_0V_1 + I_1] \qquad\qquad J_2(t) = -Y_0v_1(t-\tau) - i_1(t-\tau)$$

Lossless

$\Gamma \longrightarrow$ Delayed sources

Delay Extraction (Lossy Lines)

Lossless

$$\Gamma = s\sqrt{LC} \qquad Y_0 = \sqrt{\frac{C}{L}}$$

Lossy

$$Y_0 = \sqrt{\frac{G+sC}{R+sL}}$$

$$\Gamma(s) = \sqrt{(G+sC)(R+sL)}$$

$$J_1 = -e^{-d\Gamma}[Y_0V_2 + I_2]$$

$$J_2 = -e^{-d\Gamma}[Y_0V_1 + I_1]$$

$$e^{-d\Gamma(s)} = e^{-ds\sqrt{LC}}P(s)$$

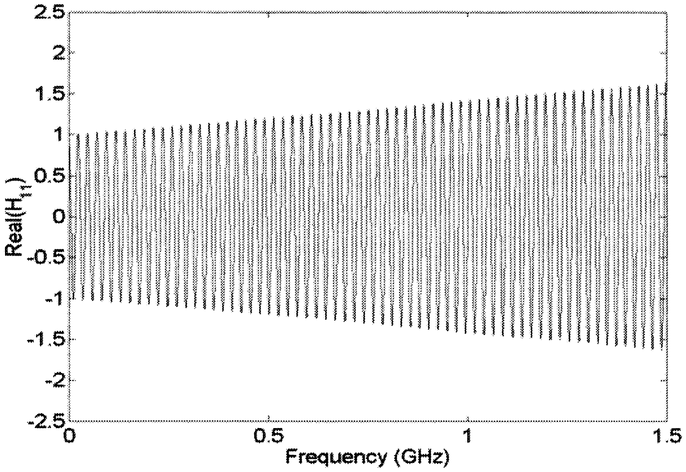
$$P(s) = e^{sT}\left[e^{-d\Gamma(s)}\right]$$

Approximation Problems

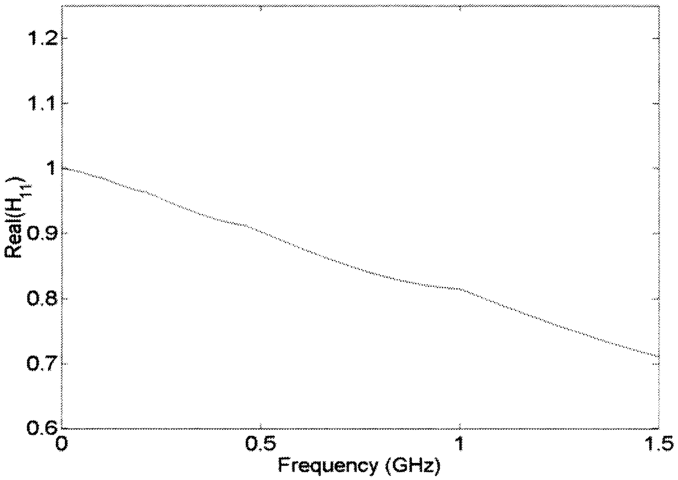
$$Y_0, P(s) \approx \frac{\Phi(s)}{\Psi(s)}$$

Why?

Without Delay Extraction



With Delay Extraction



MoC based Algorithms

→ Delay Extraction + Rational Approximation

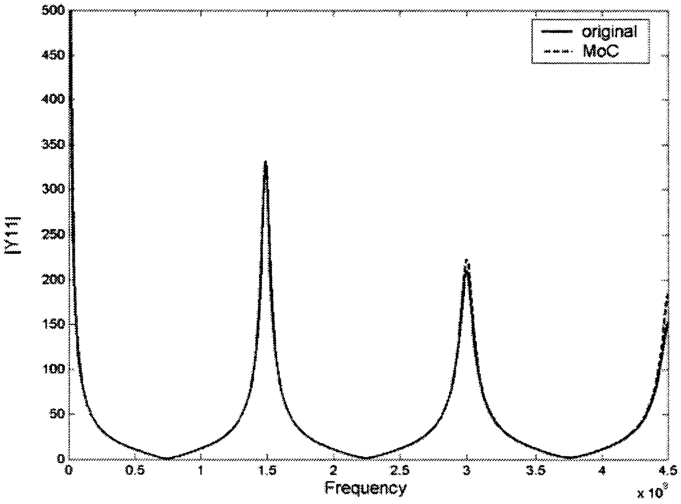
→ Efficient for Long Low Loss Lines

Difficulties

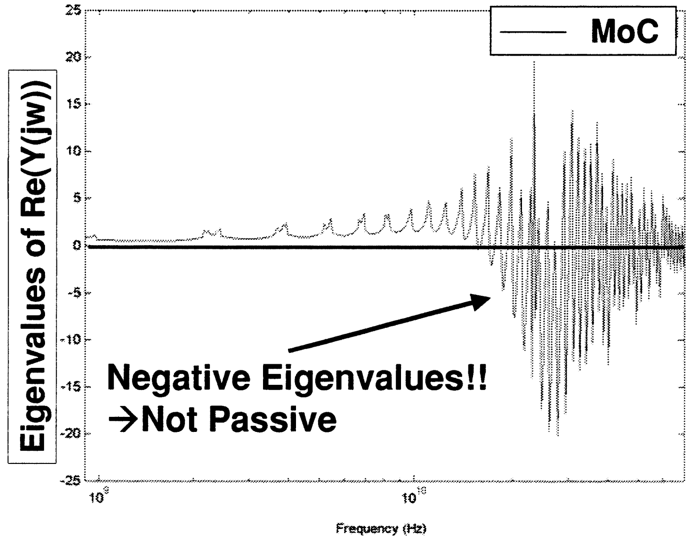
1) Coupled Lines: Curve Fiting
n Lines → $(2n^2 + n)$ Tr. functions
Eg. 10 Lines → 210 Tr. functions

2) Does not Guarantee Passivity

MoC - Accurate fit



MoC – Passivity Check



Negative Eigenvalues!!
→Not Passive

Possible Approaches

1)MoC based Algorithms

➩ ***2) Rational Approximations***

⇘

MRA (Matrix Rational Approximation)

MRA: Matrix Rational Approximation

<u>Concept:</u>

MRA: Matrix Rational Approximation

Telegrapher's Equations

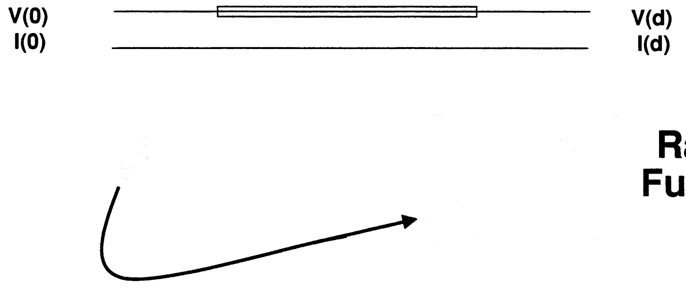
⇓

e^{Z}

⇓

$P_N(Z)^{-1} Q_N(Z)$

MRA Algorithm



Rational Functions

1) Closed-Form

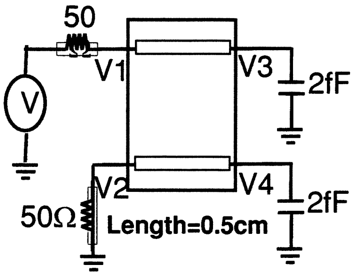
→ Large Number of Coupled Lines

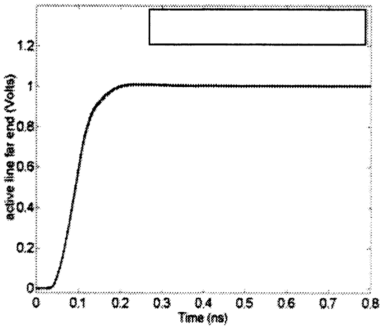
2) Passivity is guaranteed

✓

1. A. Dounavis, R. Achar and M. Nakhla, T-MTT, Oct. 2001

On-Chip Interconnect



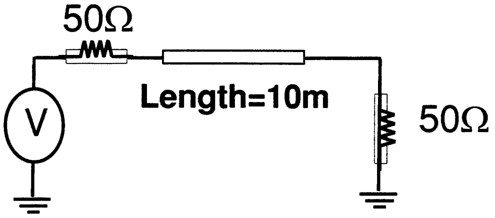


CPU Time: MRA vs MoC

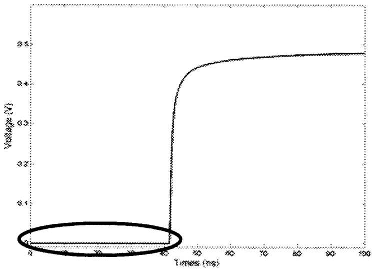
**Speed-up Ratio
10.9x**

How About Long Low-Loss Lines???

*(Low Loss TL)**



MoC	MRA
Time (s)	Time (s)
4	463



*Ruehli, Cangellaris and Huang, *Proc. EPEP-2002*

Delay Extraction: Major Challenge

e^Z

$$e^{Z} = e^{A+sB} \neq e^{A}\, e^{sB}$$

Lossy

Lossless (delay line)

[illegible]

Proposed Delay Extraction

Special Case: Lie Product[1]

Proposed Delay Extraction

<u>Modified Lie Product</u>

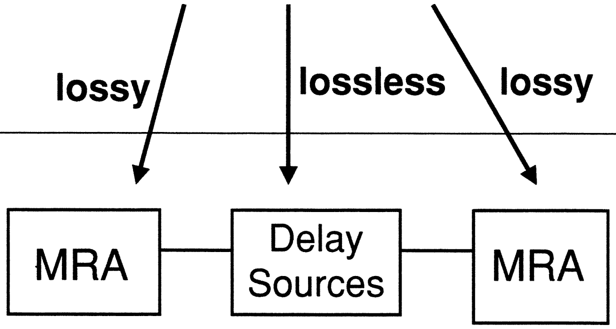
$$\text{error} \cong O\left(\frac{1}{m^2}\right)$$

Proposed Delay Extraction

Modified Lie Product - 1

$$A = \begin{bmatrix} 0 & -R \\ G & 0 \end{bmatrix}$$

$$B = \begin{bmatrix} 0 & L \\ -C & 0 \end{bmatrix}$$

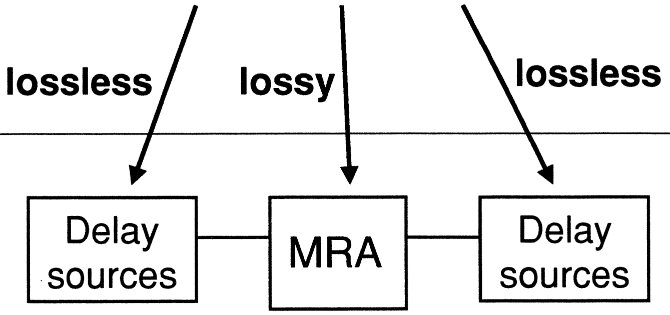


Proposed Delay Extraction

Modified Lie Product - 2

$$A = \begin{bmatrix} \text{[illegible]} & -R \\ -G & 0 \end{bmatrix}$$

$$B = \begin{bmatrix} H & -L \\ -C & 0 \end{bmatrix}$$



lossless

lossy

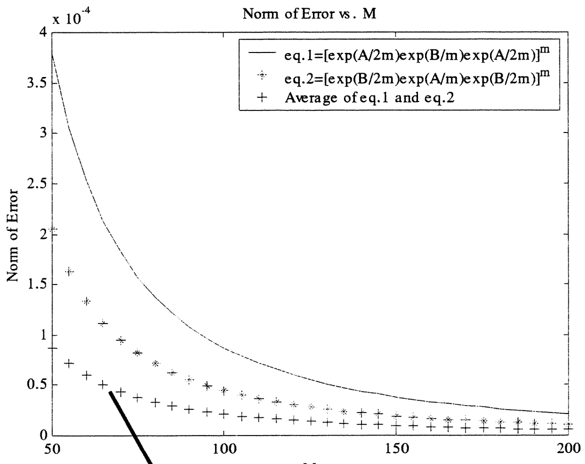
lossless

Delay sources

MRA

Delay sources

Error Estimates

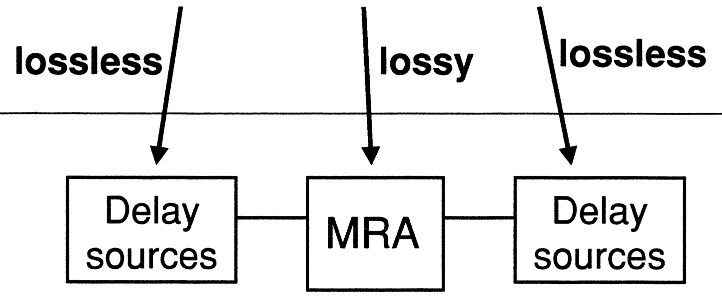


Average of Modified Lie Formula 1 and 2

Frequency Dependent Parameters

Modified Lie Product - 2

$$B_{ext} = \begin{bmatrix} 0 & -L \\ -C & 0 \end{bmatrix}$$



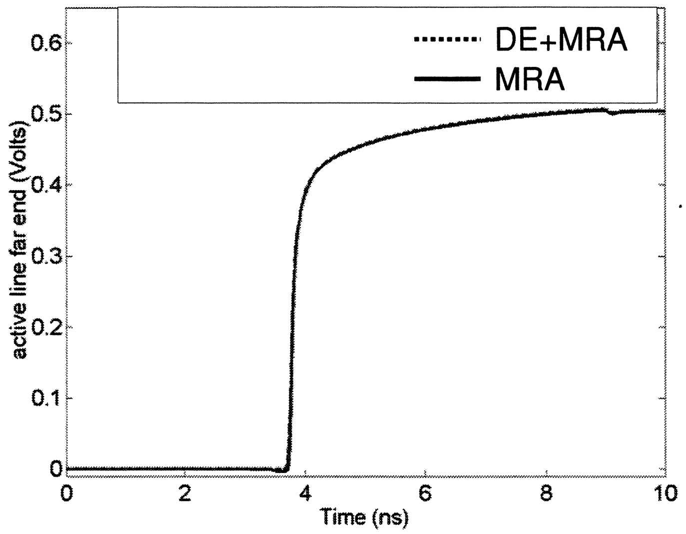
Example: Long Cable

Line	DE+MRA	MRA	Lumped
	time (s)	time (s)	time (s)
10m	4	463	1272

Computer: SUN Blade-1000 workstation with 900MHz UltraSPARC-III CPU.

Dounavis, N. Nakhla, Achar, M. Nakhla, Proc. EPEP-2003

Line 6 (40cm)



Simulation	DE+MRA	MRA	Lumped
	time (s)	time (s)	time (s)
5cm	0.89	4.16	32.4
20cm	2.47	25	292
40cm	4.32	74	4641

Computer: SUN Blade-1000 workstation with 900MHz Ultra SPARC-III CPU.

Dounavis, N. Nakhla, Achar, M. Nakhla, Proc. EPEP-2003

SUMMARY

- ***HS Interconnect Issues***
  - **Mixed F/T**
  - **Complexity**
- ***Possible Solutions***
  - **Macromodelling**
  - **Circuit Reduction**
- ***Macromodelling***
  - **Accuracy/Order**
  - **Stability**
  - **Passivity**

Workshop WFD
Advances and New Directions in Device Modeling and Design Optimization

Accurate and Efficient Analysis of Large Spiral Inductors with Thick Metal and Narrow Gaps Using Space Mapping

SONNET®

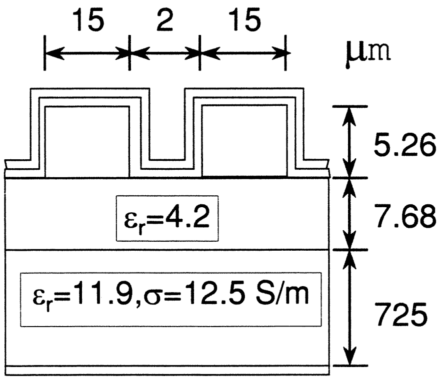
James C. Rautio
Sonnet Software

Introduction

- Increasing spiral inductor Q is critical.
  - Use circular spiral geometries.
  - Use very thick metal.
  - Make conductors as wide as possible and gaps as narrow as possible.
- This is a worst case nightmare for EM analysis!

The Problem

- Gap = 2 microns.
- Thickness = 5.26 microns.
- Gap coupling significant at high frequency.
- Thin conformal passivation is non-planar.

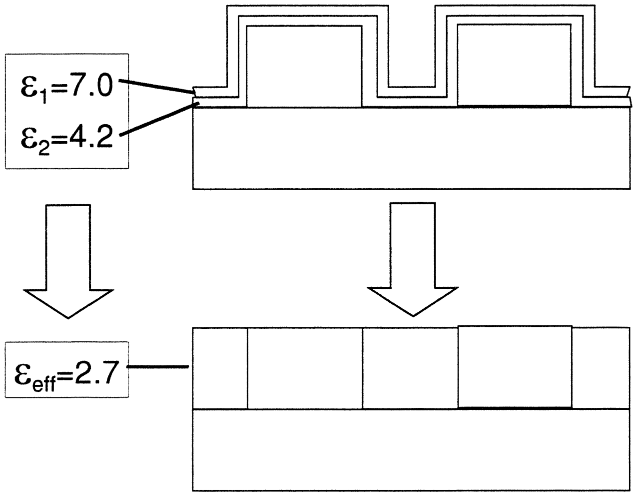


Data courtesy Jazz Semiconductor.

Drawing not to scale.

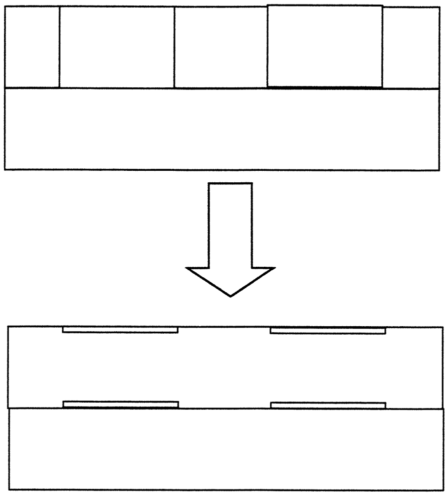
Passivation Model

- ε_{eff} chosen so same cap. between lines.



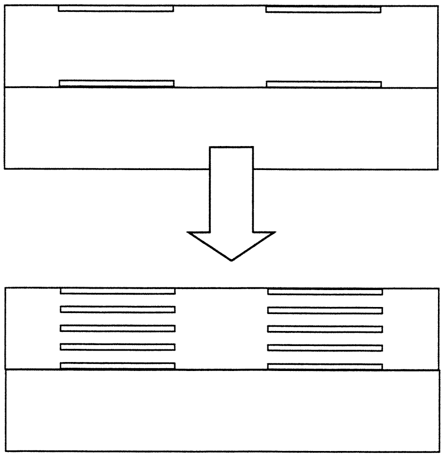
Two-Sheet Model

- Two thin sheets to model thick metal.
- OK if gap ≈ thickness.
- Need more sheets for this one.

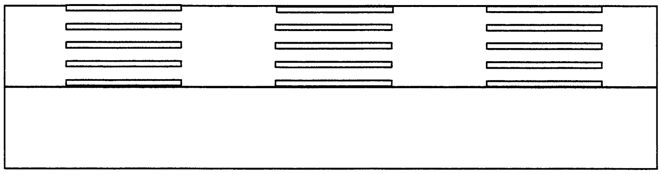


N-Sheet Model

- As N increases, result converges.
- Can only use for small structures.
- Would rather use 2-sheet model, much faster.



N-Sheet Convergence



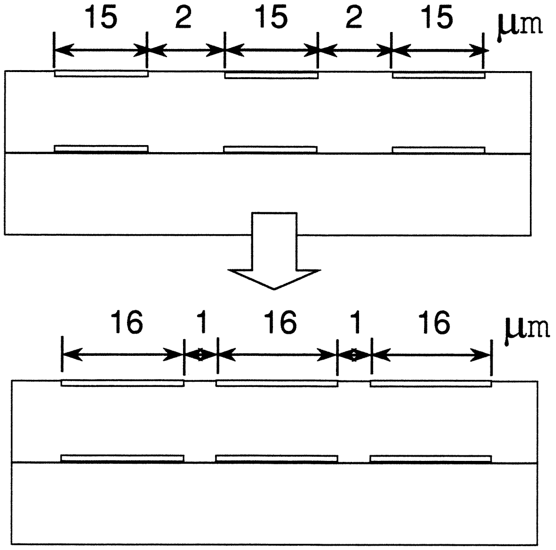
- Check convergence for simple structure: CPW through line.
- Calculate L and C per unit length as N increases.

N-Sheet Convergence

- 2-sheet model has:
  - –48 nH/m too much L.
  - –56 pF/m too little C.
- Let's modify the 2-sheet model!

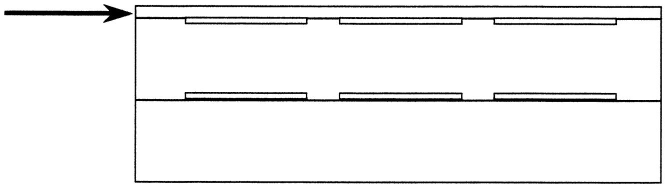
N	L (nH/m)	C (pF/m)
1	296	133
2	188	165
3	168	184
5	154	201
9	146	211
17	143	216
∞	140	221

Modified 2-Sheet Model



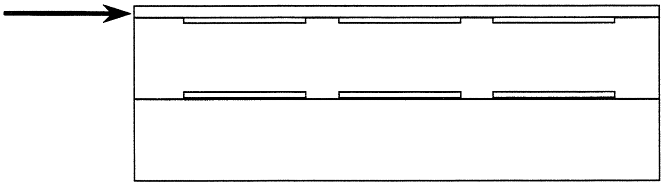
- Wider line = less L/m
- Narrower gap = more C/m.
- But more is needed.

Space Mapping Layer



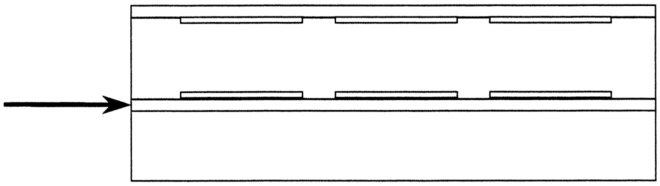
- Add very thin "space mapping layer" on top.
- Adjust ε_r and μ_r until we get right C/m and L/m.***x***

Space Mapping Layer



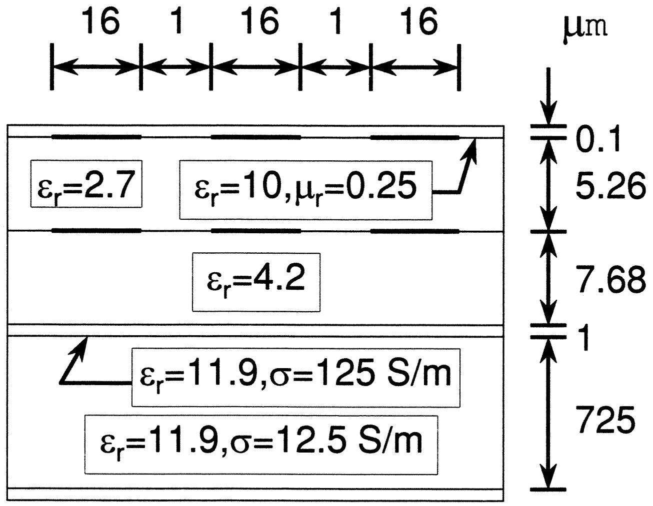
- Arbitrarily set thickness to 0.1 micron.
- Keep increasing ε_r until desired C/m seen ➔ $\varepsilon_r = 10$.
- Keep decreasing μ_r until desired L/m seen ➔ $\mu_r = 0.25$.

Surprise Problem



- In course of work, discovered that there was a 1 micron boron implant in surface of silicon.
- 10X higher conductivity.
- Important at higher frequencies.

Complete Model


16 1 16 1 16

μm

0.1

5.26

7.68

1

725

ε_r=2.7

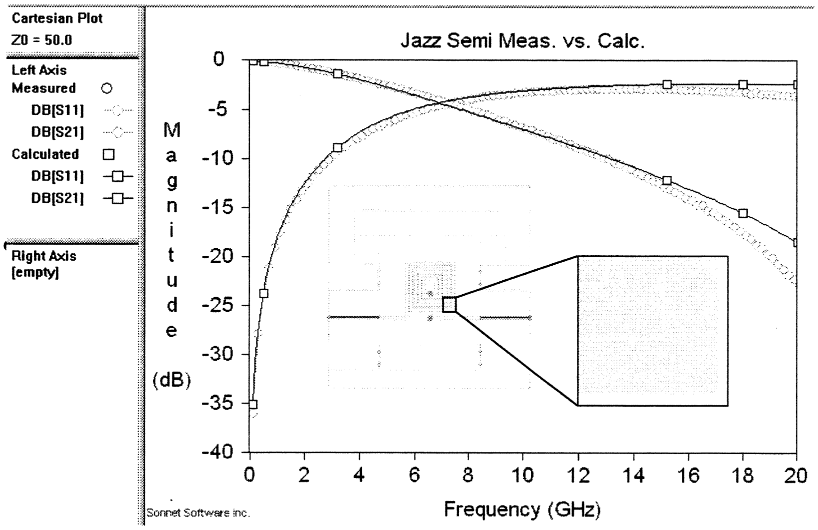
ε_r=10,μ_r=0.25

ε_r=4.2

ε_r=11.9,σ=125 S/m

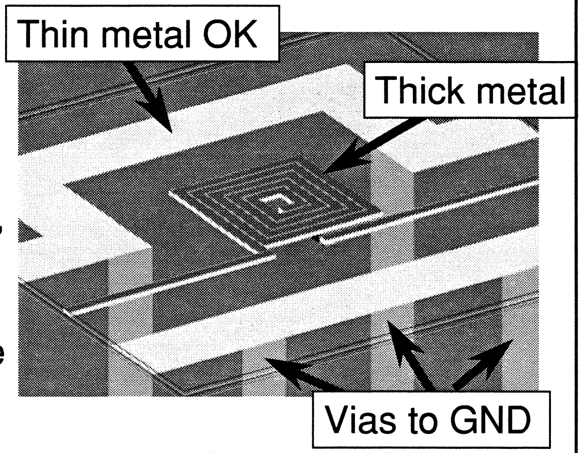
ε_r=11.9,σ=12.5 S/m

Meas. Vs. Calculated



Sonnet Software inc.

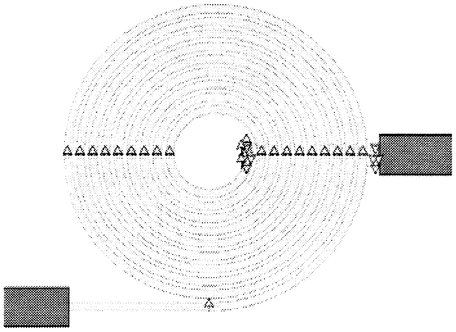
Additional Considerations



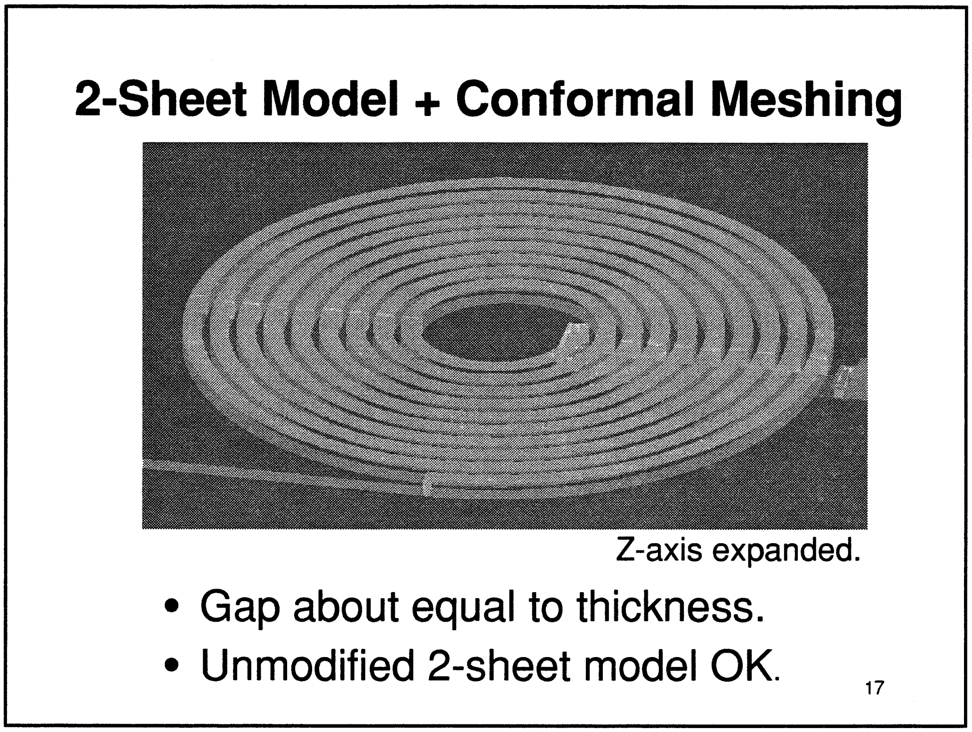
- Path of ground return current is important.
- If "Ground Cage" in measurement, be sure it's in the analysis.
- If ground strips are connected to substrate ground with vias, then include vias in analysis.

8.25-Turn Circ. Spiral

- Important to include natural high edge current.
  - Affects Q and inductance.
- Requires narrow subsections along all edges.
- Very difficult to do for long smooth curves.



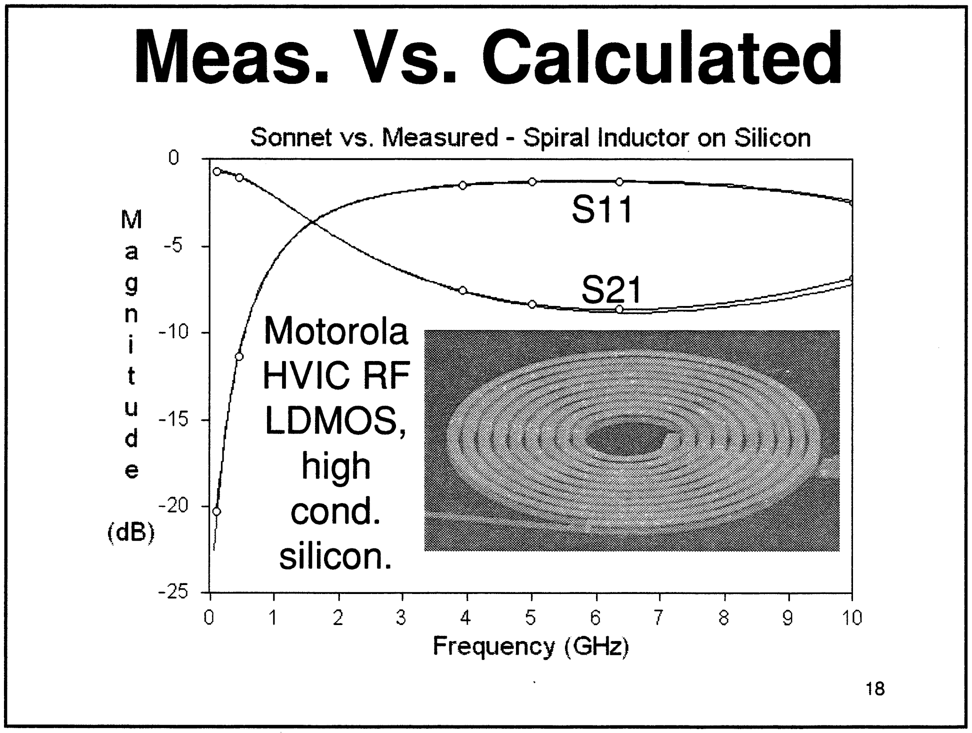
Data courtesy Motorola.



2-Sheet Model + Conformal Meshing

Z-axis expanded.

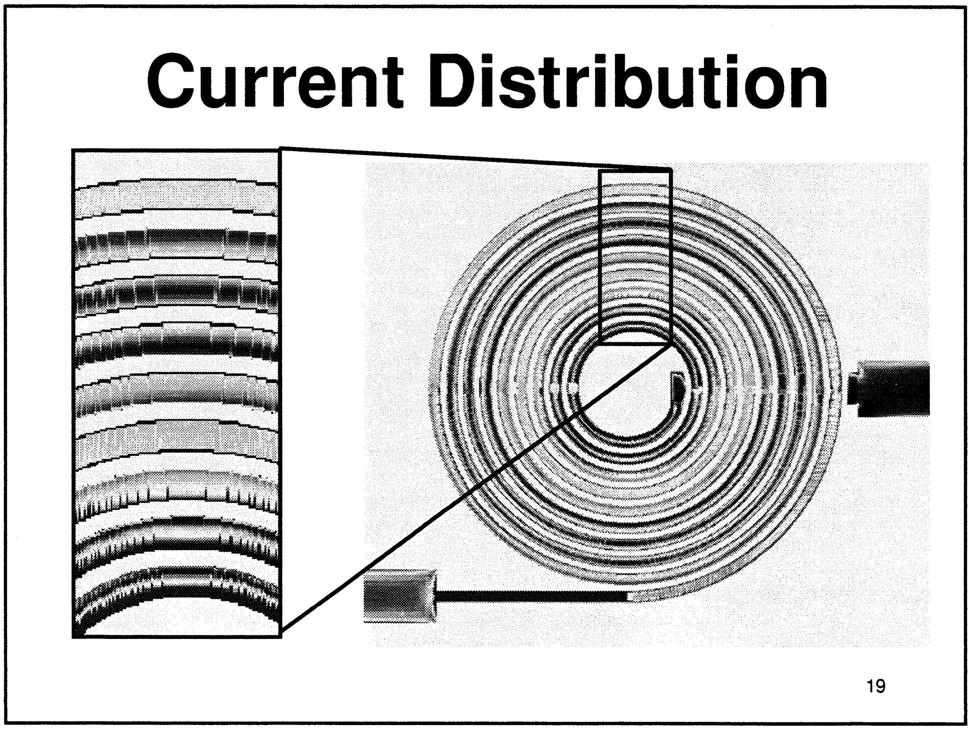
- Gap about equal to thickness.
- Unmodified 2-sheet model OK.



Meas. Vs. Calculated

Sonnet vs. Measured - Spiral Inductor on Silicon

Motorola HVIC RF LDMOS, high cond. silicon.



Current Distribution

Conclusion

- Thick metal + narrow gap difficult for EM analysis.
- If gap and thickness about the same, just use 2-sheet model.
- If gap much less than thickness, use N-sheet, or modified 2-sheet.
- For circular spirals, use conformal meshing.

ITESO

Electromagnetics-based Design through Inverse Space Mapping Techniques

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presented at

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2004 IEEE MTT-S International Microwave Symposium, Fort Worth, TX, June 11, 2004

Electromagnetics-based Design through Inverse Space Mapping Techniques

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Abstract

Inverse space mapping algorithms for designing with accurate but computationally expensive simulators are described and contrasted in this presentation.

Neural Inverse Space Mapping (NISM) optimization was the first space mapping algorithm that explicitly made use of the inverse of the mapping from the fine to the coarse model parameter spaces. NISM follows an aggressive formulation by not requiring a number of up-front fine model evaluations to start building the mapping. An statistical procedure to parameter extraction (PE) is employed in NISM to avoid the need for multipoint matching and frequency mappings. An artificial neural network (ANN) whose generalization performance is controlled through a network growing strategy approximates the inverse mapping at each iteration. The ANN starts from a 2-layer perceptron and automatically migrates to a 3-layer perceptron when the amount of nonlinearity found in the inverse mapping becomes significant. The NISM step consists of evaluating the current neural network at the optimal coarse model solution.

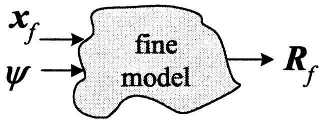
Linear Inverse Space Mapping (LISM) follows a piece-wise linear formulation to implement the inverse of the mapping, avoiding the use of neural networks. LISM approximates the inverse of the mapping function at each iteration by linearly interpolating the last $n + 1$ pairs of coarse and fine model design parameters, where n is the number of optimization variables. The same statistical procedure to PE is used in LISM as in NISM. LISM also follows an aggressive formulation in the sense of not requiring up-front fine model evaluations. LISM has been applied to design linear circuits in the frequency domain and nonlinear circuits in the time domain transient-state.

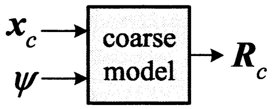
A rigorous comparison between Broyden-based direct space mapping, neural (NISM) and linear (LISM) inverse space mapping is realized using a synthetic example. Two industrially relevant microwave design problems are efficiently solved using inverse space mapping techniques.

Outline

- "Direct" space mapping
- Broyden-based direct space mapping
- Inverse space mapping: linear and neural
- A comparison between Broyden, LISM and NISM
- Examples using LISM and NISM
- Conclusions

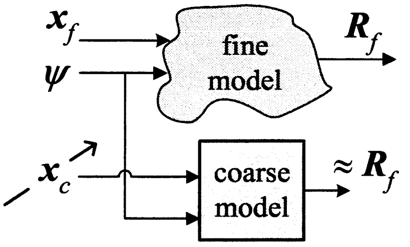
Direct Space Mapping for Design





$$\boldsymbol{x}_c^* = \arg\min_{\boldsymbol{x}_c} U(\boldsymbol{R}_c(\boldsymbol{x}_c,\psi))$$

$$\boldsymbol{f}(\boldsymbol{x}_f) = \boldsymbol{P}(\boldsymbol{x}_f) - \boldsymbol{x}_c^*$$



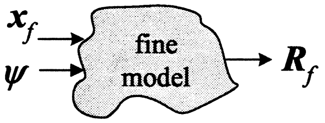
A SM solution $\boldsymbol{x}_f^{SM}$ is found when

$$\boldsymbol{f}(\boldsymbol{x}_f^{SM}) \approx \boldsymbol{0}$$

$$\Longrightarrow \boldsymbol{R}_f(\boldsymbol{x}_f^{SM}) \approx \boldsymbol{R}_c(\boldsymbol{x}_c^*)$$

$$\boldsymbol{P}(\boldsymbol{x}_f) = \arg\min_{\boldsymbol{x}_c} \left\| \boldsymbol{R}_f(\boldsymbol{x}_f,\psi) - \boldsymbol{R}_c(\boldsymbol{x}_c,\psi) \right\|_2^2$$

The Fine Model Solution



$$\boldsymbol{x}_f^* = \arg\min_{\boldsymbol{x}_f} U(\boldsymbol{R}_f(\boldsymbol{x}_f,\psi))$$

When solving

$$\boldsymbol{f}(\boldsymbol{x}_f) = \boldsymbol{P}(\boldsymbol{x}_f) - \boldsymbol{x}_c^* = \boldsymbol{0}$$

we do not attempt to find $\boldsymbol{x}_f^*$

$$\boldsymbol{x}_f^{SM} \neq \boldsymbol{x}_f^*$$

Broyden-based Direct Space Mapping

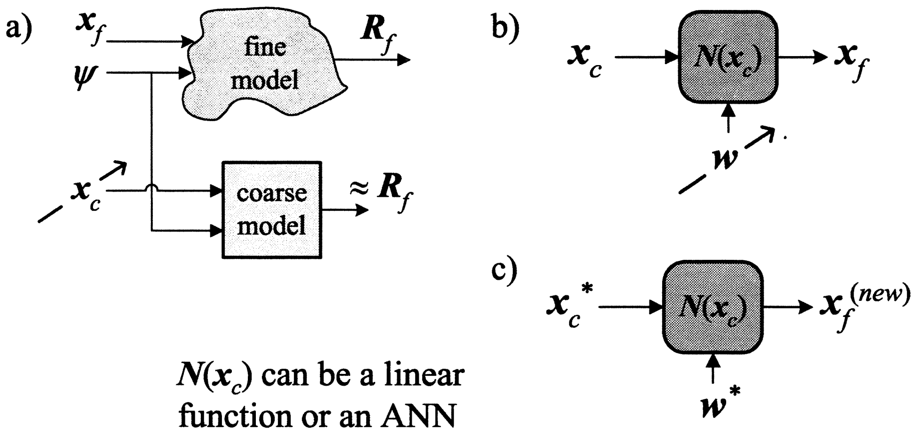
```
begin
   $x_f = x_c^*$, $B = I$, $f = P(x_f) - x_c^*$
   repeat until stopping_criterion
      solve $Bh = -f$ for $h$
      $x_f = x_f + h$
      $f = P(x_f) - x_c^*$
      $B = B + \frac{f h^T}{h^T h}$
end
```

where $\boldsymbol{P}(\boldsymbol{x}_f) = \arg\min_{\boldsymbol{x}_c} \left\| \boldsymbol{R}_f(\boldsymbol{x}_f,\psi) - \boldsymbol{R}_c(\boldsymbol{x}_c,\psi) \right\|_2^2$

(Bandler et al., 1995)

Inverse Space Mapping for Design

Main sub-processes



$N(x_c)$ can be a linear function or an ANN

Inverse Space Mapping

```
begin
  $i = 0, x_f^{(i)} = x_c^*$, initialize $w$
  repeat until stopping_criterion
      $x_c^{(i)} = \arg\min_{x_c} \| R_f(x_f, \psi) - R_c(x_c, \psi) \|_2^2$
      train $N(w)$ to find $w^*$
      $x_f^{(i+1)} = N(w^*, x_c^*)$
end
```

Training a Linear Inverse Mapping

- $\boldsymbol{N}(\boldsymbol{w}, \boldsymbol{x}_c) = \boldsymbol{A}\boldsymbol{x}_c + \boldsymbol{b}$, $\boldsymbol{A} \in \Re^{n \times n}$, $\boldsymbol{b} \in \Re^{n}$
  where $\boldsymbol{w}$ contains $\boldsymbol{b}$ and the columns of $\boldsymbol{A}$
- It is "trained" by solving

$$\min_{\boldsymbol{A},\boldsymbol{b}} \left\| \boldsymbol{e}_{i-n}^T \quad \cdots \quad \boldsymbol{e}_i^T \right\|_2^2$$

where

$$\boldsymbol{e}_k = \boldsymbol{A}\boldsymbol{x}_c^{(k)} + \boldsymbol{b} - \boldsymbol{x}_f^{(k)}$$

(Rayas-Sánchez et al., 2004)

Training a Neural Inverse Mapping

- $\boldsymbol{N}(\boldsymbol{x}_c, \boldsymbol{w}) = \boldsymbol{W}^o \boldsymbol{\Phi}(\boldsymbol{x}_c) + \boldsymbol{b}^o$

  $\boldsymbol{\Phi}(\boldsymbol{x}_c) = [\varphi(s_1) \quad \varphi(s_2) \quad \dots \quad \varphi(s_h)]^T$

  $\boldsymbol{s} = \boldsymbol{W}^h \boldsymbol{x}_c + \boldsymbol{b}^h$

  where $\boldsymbol{w}$ contains $\boldsymbol{b}^o \in \Re^{n}$, $\boldsymbol{b}^h \in \Re^{h}$ and the columns of $\boldsymbol{W}^o \in \Re^{n \times h}$ and $\boldsymbol{W}^h \in \Re^{h \times n}$
- It is trained by solving

$$\min_{\boldsymbol{w}} \left\| \boldsymbol{e}_1^T \quad \cdots \quad \boldsymbol{e}_i^T \right\|$$

where

$$\boldsymbol{e}_k = \boldsymbol{x}_f^{(k)} - \boldsymbol{N}(\boldsymbol{x}_c^{(k)}, \boldsymbol{w})$$

(Bandler et al., 2001)

ASM-Broyden *vs* LISM *vs* NISM

- The same termination criteria are used

$$\left\|\boldsymbol{f}(\boldsymbol{x}_f^{(i)})\right\|_\infty < \varepsilon_1 \quad \text{or} \quad \left\|\boldsymbol{x}_f^{(i+1)} - \boldsymbol{x}_f^{(i)}\right\|_2 \le \varepsilon_2(\varepsilon_2 + \left\|\boldsymbol{x}_f^{(i)}\right\|_2) \quad \text{or} \quad i > 4n$$

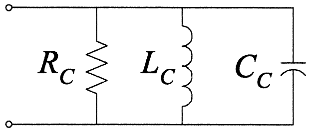
$$\varepsilon_1 = 10^{-4} \qquad \varepsilon_2 = 10^{-5}$$

- The same statistical parameter extraction algorithm is used at the i-th iteration to solve

$$\min_{\boldsymbol{x}_c} \left\|\boldsymbol{R}_f(\boldsymbol{x}_f^{(i)}, \psi) - \boldsymbol{R}_c(\boldsymbol{x}_c, \psi)\right\|_2^2$$

ASM-Broyden *vs* LISM *vs* NISM (cont)

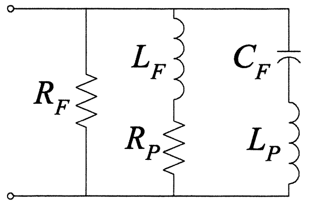
"Coarse" model



$\boldsymbol{x}_c = [R_C\,(\Omega)\;\; L_C\,(\text{nH})\;\; C_C\,(\text{pF})]^T$

$\boldsymbol{x}_c^* = [50\;\; 0.2370\;\; 11.8792]^T$

"Fine" model



$\boldsymbol{x}_f = [R_F\,(\Omega)\;\; L_F\,(\text{nH})\;\; C_F\,(\text{pF})]^T$

$R_P = 0.5\ \Omega$

$L_P = 0.13$ nH

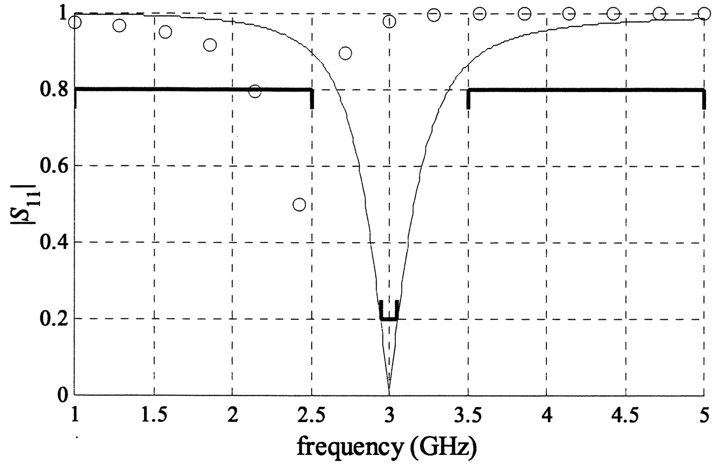
Specifications ($Z_o = 50\ \Omega$)

$|S_{11}| > 0.8$ from 1 GHz to 2.5 GHz and from 3.5 GHz to 5 GHz

$|S_{11}| < 0.2$ from 2.95 GHz to 3.05 GHz

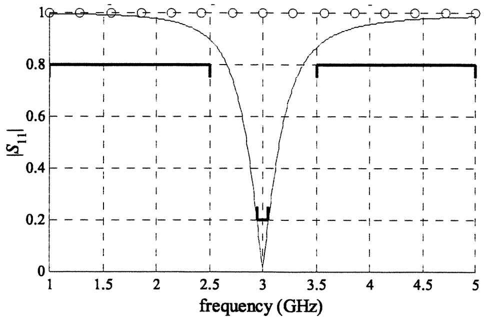
ASM-Broyden *vs* LISM *vs* NISM (cont)

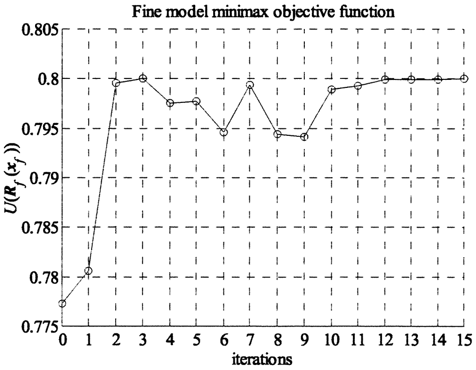
Starting point:
fine (○) and coarse model (–) responses at $\boldsymbol{x}_c^*$



ASM-Broyden *vs* LISM *vs* NISM (cont)

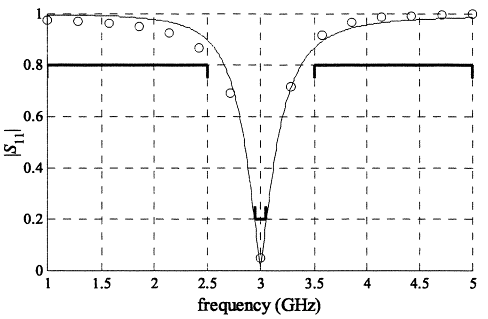
Coarse model response (–) at
$\boldsymbol{x}_c^*$ and fine model response (○)
at $\boldsymbol{x}_f^{\mathrm{BROYDEN}}$

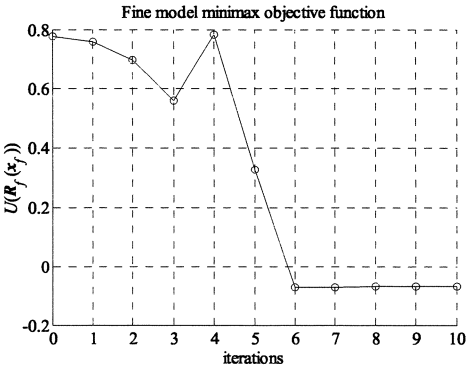




ASM-Broyden *vs* LISM *vs* NISM (cont)

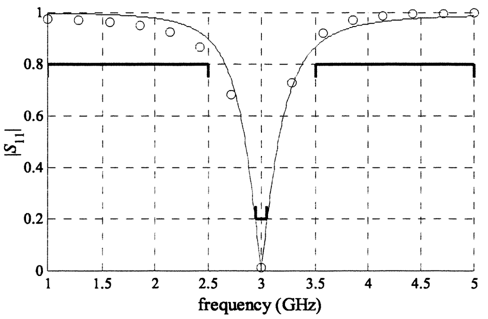
Coarse model response (–) at $\boldsymbol{x}_c^*$ and fine model response (○) at $\boldsymbol{x}_f^{\mathrm{LISM}}$

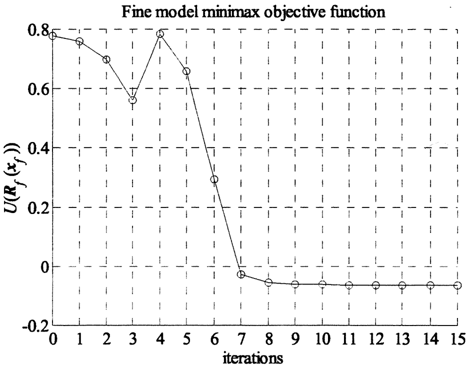




ASM-Broyden *vs* LISM *vs* NISM (cont)

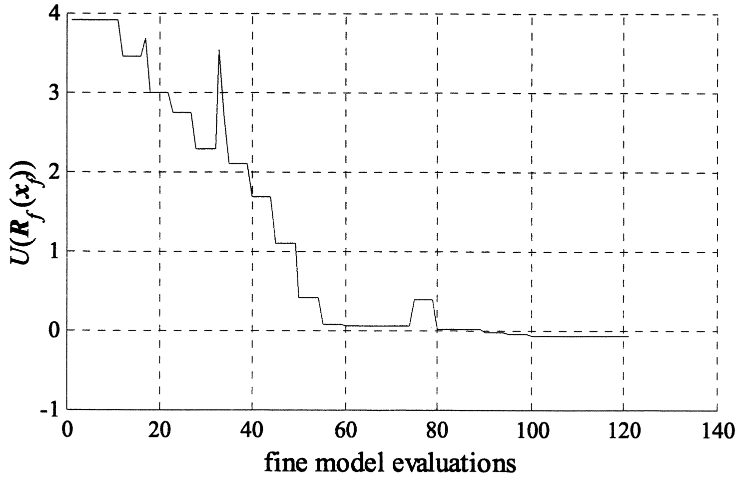
Coarse model response (–) at $\boldsymbol{x}_c^*$ and fine model response (○) at $\boldsymbol{x}_f^{\mathrm{NISM}}$





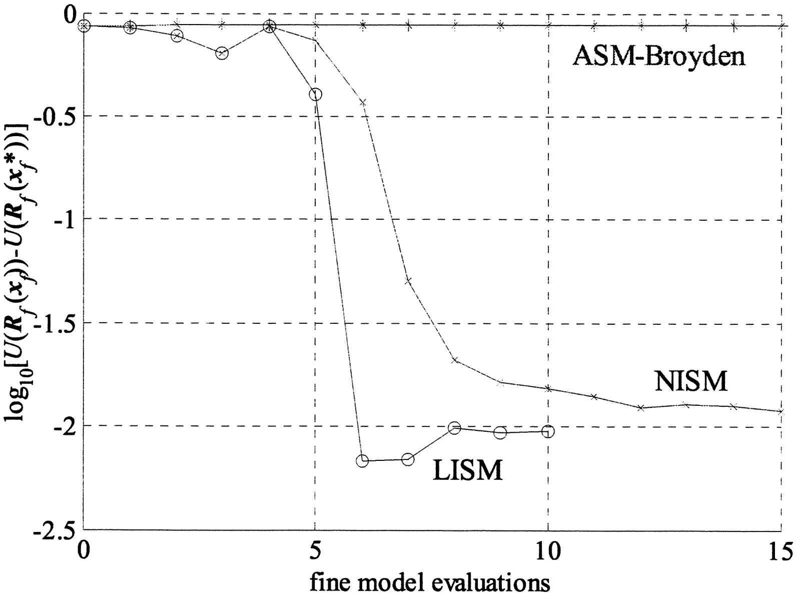
ASM-Broyden *vs* LISM *vs* NISM (cont)

Optimizing the “fine” model directly



$\boldsymbol{x}_f^* = [153.0188 \quad 0.3229 \quad 6.1815]^T$

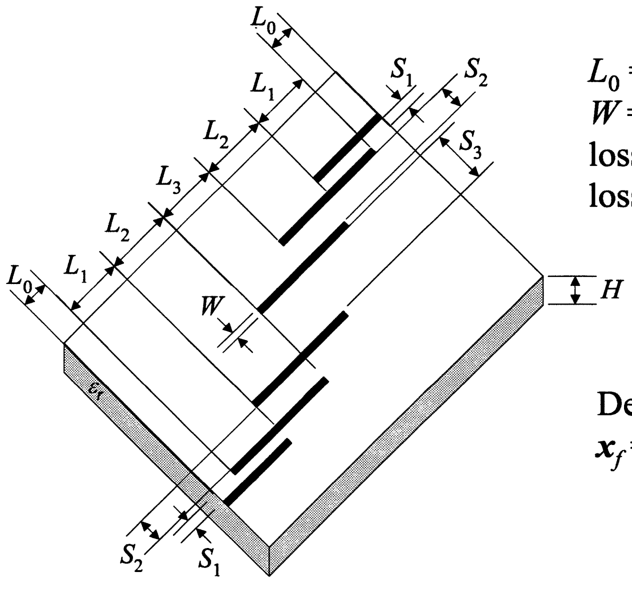
ASM-Broyden *vs* LISM *vs* NISM (cont)



ASM-Broyden *vs* LISM *vs* NISM (cont)

method	solution $x_f^{(i)}$	$U(R_f(x_f^{(i)}))$	i	termination condition
Direct optimization	$[153.0188 \quad 0.3229 \quad 6.1815]^T$	-0.0780	121	-
LISM	$[123.3491 \quad 0.3468 \quad 5.8238]^T$	-0.06854	10	$\left\lVert f(x_f^{(i)}) \right\rVert_\infty < \varepsilon_1$
NISM	$[124.0788 \quad 0.3472 \quad 5.8601]^T$	-0.0661	15	$i > 4n$
ASM-Broyden	fails	0.8000	15	$i > 4n$

HTS Filter *(Westinghouse, 1993)*



$L_0 = 50$ mil, $H = 20$ mil,
$W = 7$ mil, $\varepsilon_r = 23.425$,
loss tangent $= 3\times10^{-5}$;
lossless metalization

Design parameters
$x_f = [L_1 \; L_2 \; L_3 \; S_1 \; S_2 \; S_3]^T$

Optimizing the HTS Microstrip Filter

Specifications

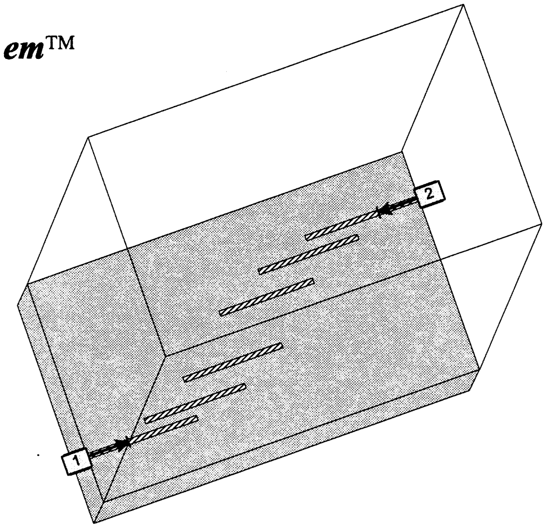
$|S_{21}| \geq 0.95$ for 4.008 GHz $\leq f \leq$ 4.058 GHz

$|S_{21}| \leq 0.05$ for $f \leq$ 3.967 GHz and $f \geq$ 4.099 GHz

Optimizing the HTS Microstrip Filter (cont)

Fine model

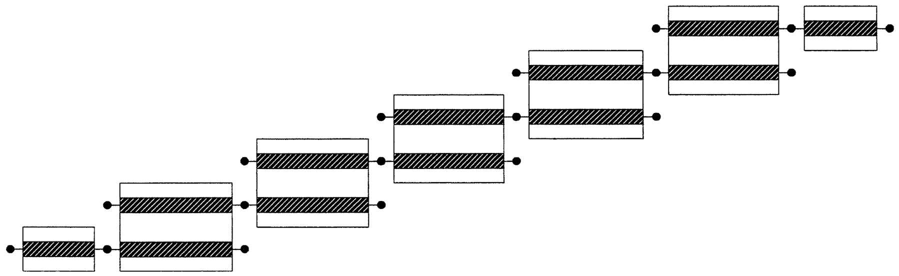
Sonnet's ***em***™



Optimizing the HTS Microstrip Filter (cont)

Coarse model

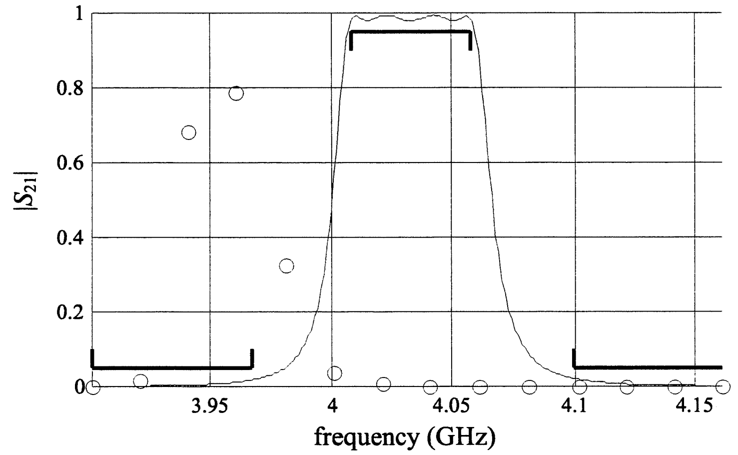
OSA90/hope™ built-in models of open circuits, microstrip lines and coupled microstrip lines



Optimizing the HTS Microstrip Filter (cont)

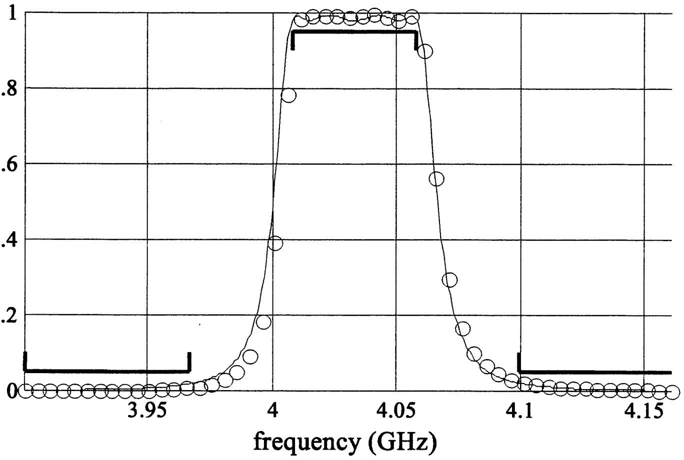
Starting point

OSA90/hope™ (–) and ***em***™ (○) at $\boldsymbol{x}_c^*$



NISM Optimization of the HTS Filter

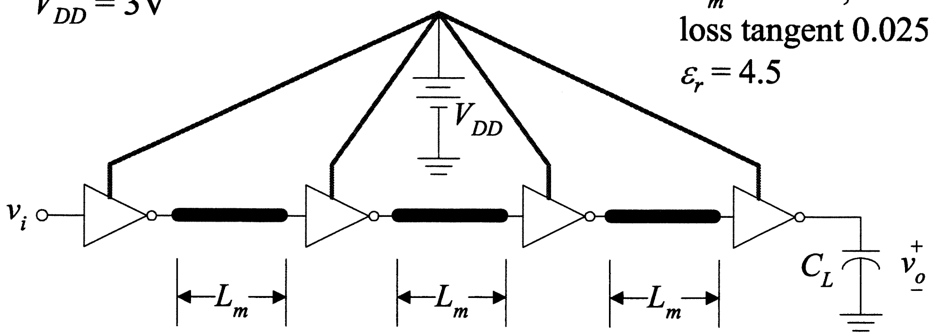
Responses using OSA90/hope™ (–) at $\boldsymbol{x}_c^*$ and
em™ (○) at the NISM solution (after 3 NISM iterations)



CMOS Drivers for a Long Microstrip Line

$L_m = 200\text{mil}$
$C_L = 0.1\text{pF}$
$V_{DD} = 3\text{V}$

FR4 substrate:
$H_m = 10\text{mil}$,
$W_m = 19\text{mil}$,
loss tangent 0.025
$\varepsilon_r = 4.5$



v_i : trapezoidal pulse with a 3V amplitude, 2.5 ns duration, 100 ps rise time and fall time

CMOS Drivers for a Long Line (cont)

Specifications

$v_o(t) < 0.3\text{V}$ for $0 \le t \le 1\text{ns}$

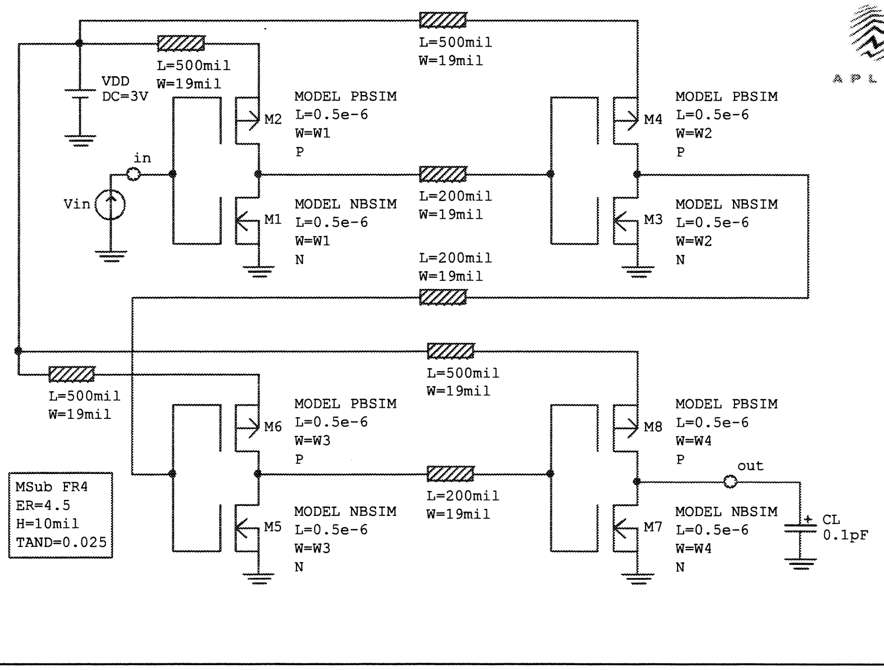
$v_o(t) > 2.7\text{V}$ for $3\text{ns} \le t \le 4.5\text{ns}$

$v_o(t) < 0.3\text{V}$ for $6.5\text{ns} \le t \le 8.5\text{ns}$

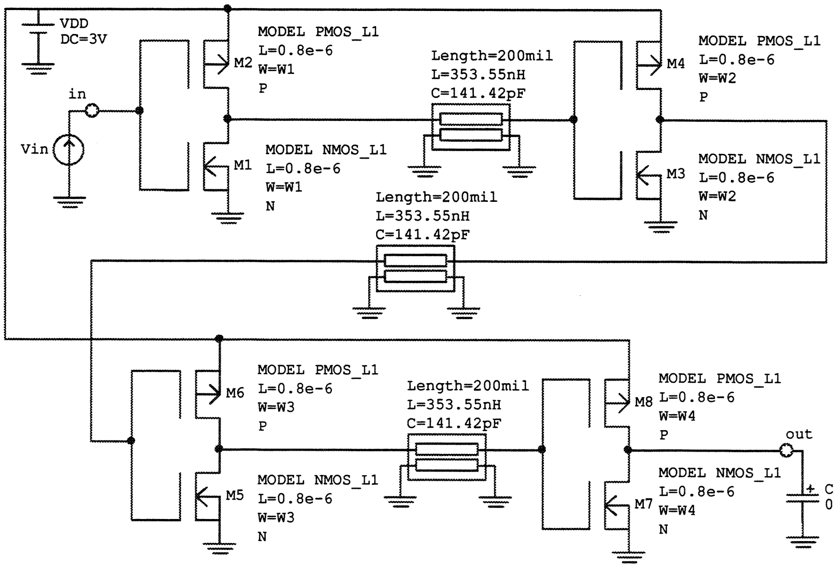
Optimization variables

$\boldsymbol{x}_f = [W_1\ W_2\ W_3\ W_4]^T$
(channel width of each MOSFET,
0.5μm CMOS technology)

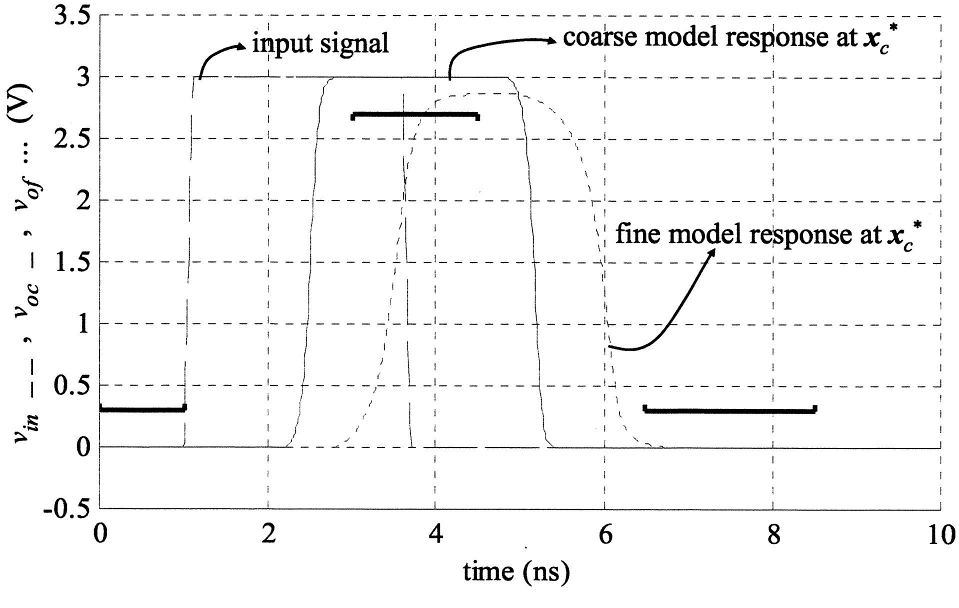
CMOS Drivers for a Long Line – Fine Model



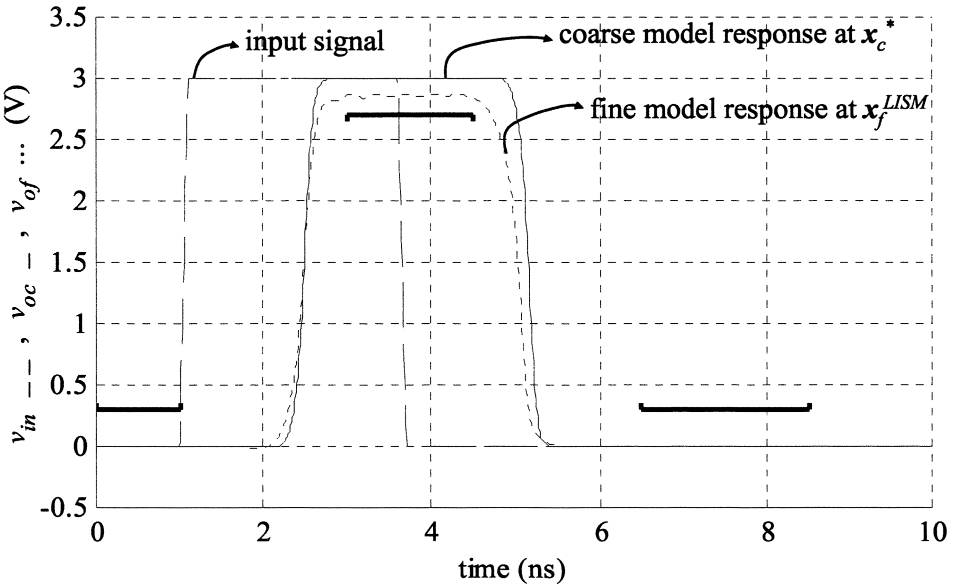
CMOS Drivers for a Long Line – Coarse Model



Optimizing the CMOS Drivers – Starting Point



Optimizing the CMOS Drivers – SM Solution



Conclusions

- Inverse space mapping techniques for designing with accurate but computationally expensive simulators are described
- Broyden-based direct space mapping is reviewed
- NISM optimization implements the inverse mapping with an ANN at each iteration
- LISM follows a piece-wise linear formulation to implement the inverse of the mapping
- A comparison between Broyden, NISM and LISM is realized
- LISM outperforms the other two techniques

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Coarse EM Modeling of LTCC RF Circuits and Its Applications to Optimization Design

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Outline

☐ **Partial Element Equivalent Circuit (PEEC), Coarse EM model for LTCC RF circuits**

☐ **Physically expressive equivalent circuit model including parasitics**

☐ **Dynamic coarse model**

☐ **Application of dynamic coarse model to Aggressive Space Mapping (ASM)**

Challenges in ASM

- ☐ **Requiring an effective coarse model:**

  physically sound, fast though less accurate

- ☐ **Uniqueness in parameter extraction**
- ☐ **The variable match**

Why PEEC model?

- ☐ **PEEC model is suitable for EM simulation of multi-layers LTCC RF circuits**
- ☐ **Static PEEC model is much faster than full wave EM models**
- ☐ **PEEC model works in the same variable space as that in full wave EM models**

Basic concept of PEEC model

For a system of K conductors:

$$E_x^i(\mathbf{r},t)=\frac{J_x(\mathbf{r},t)}{\sigma}+\sum_{k=1}^{K}\sum_{n}\frac{\mu}{4\pi}\left[\int_{v_n}G(\mathbf{r},\mathbf{r}')dv_n'\right]\frac{\partial J_n^x(t')}{\partial t}$$

$$+\sum_{k=1}^{K}\sum_{m}\frac{1}{4\pi\varepsilon}\frac{\partial}{\partial x}\left[q_m(t')\int_{s_m}G(\mathbf{r},\mathbf{r}')ds_m'\right]$$

$\mathbf{E}=0$

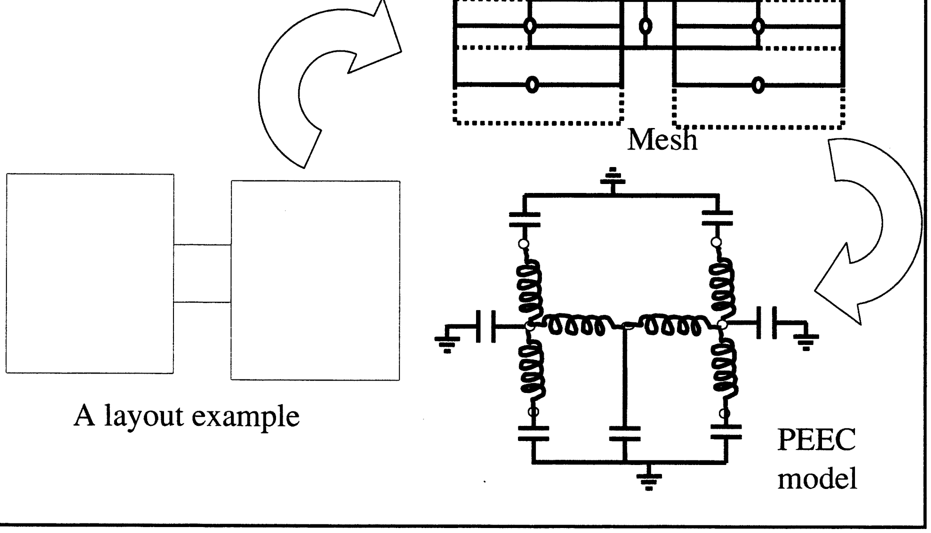
$$0=V_R+V_L+V_C$$

$V_R=0$ for Perfect conductor

$$V_L=\sum_{k=1}^{K}\sum_{n}\left\{\frac{\mu}{4\pi}\frac{1}{a_l a_n}\left[\int_{v_l}\int_{v_n}G(\mathbf{r},\mathbf{r}')dv_n'dv_l\right]\right\}\frac{\partial I_n^x(t')}{\partial t}=\sum_{k=1}^{K}\sum_{n}L_{p_{l,n}}\frac{dI_n^x(t')}{dt}$$

$$V_C=\sum_{k=1}^{K}\sum_{m}\left\{\frac{1}{4\pi\varepsilon}\frac{1}{S_m}\left[\int_{s_m}G(\mathbf{r}_l^+,\mathbf{r}')ds_m'-\int_{s_m}G(\mathbf{r}_l^-,\mathbf{r}')ds_m'\right]\right\}Q_m(t')$$

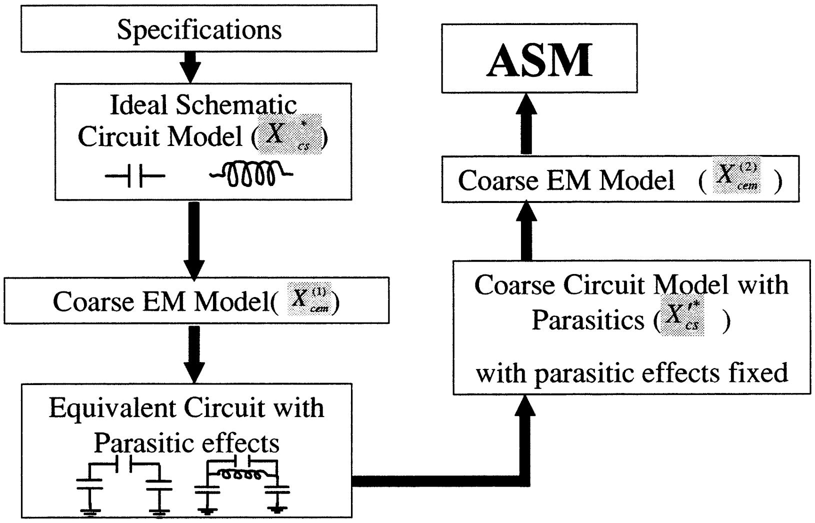
Circuit model presentation of PEEC



How to accelerate PEEC model?

- ☐ Using static Green function for small electrical size LTCC RF circuits
- ☐ Using fast circuit solver other than conventional matrix solver
- ☐ Using node deduction method that is independent to frequency

Dynamic Optimal Coarse Model



Specifications

Ideal Schematic Circuit Model (X_{cs}^{*})

Coarse EM Model($X_{cem}^{(1)}$)

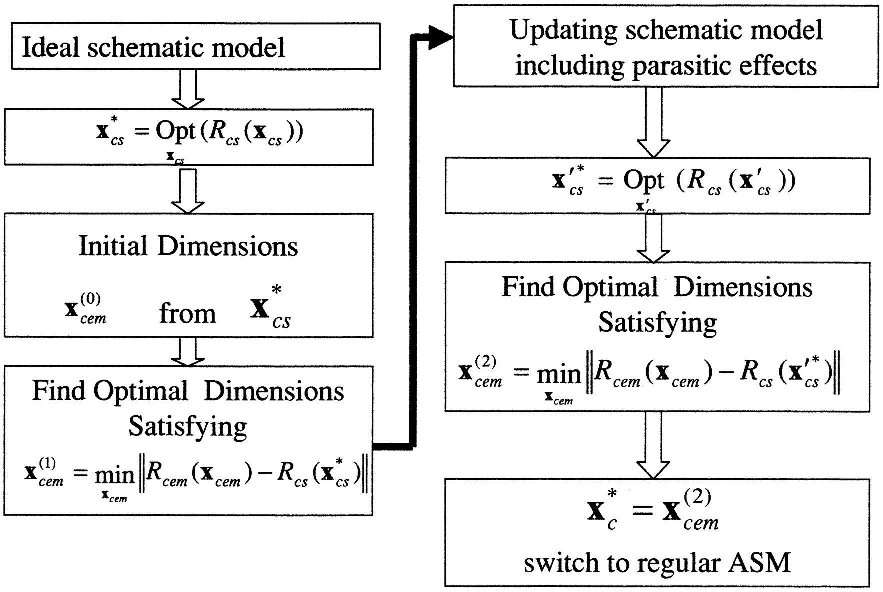
Equivalent Circuit with Parasitic effects

Coarse Circuit Model with Parasitics ($X_{cs}^{\prime *}$)

with parasitic effects fixed

Coarse EM Model ($X_{cem}^{(2)}$)

ASM



Ideal schematic model

$\mathbf{x}_{cs}^{*} = \underset{\mathbf{x}_{cs}}{\mathrm{Opt}}(R_{cs}(\mathbf{x}_{cs}))$

Initial Dimensions

$\mathbf{x}_{cem}^{(0)}$ from $\mathbf{x}_{cs}^{*}$

Find Optimal Dimensions Satisfying

$\mathbf{x}_{cem}^{(1)} = \min_{\mathbf{x}_{cem}} \left\| R_{cem}(\mathbf{x}_{cem}) - R_{cs}(\mathbf{x}_{cs}^{*}) \right\|$

Updating schematic model including parasitic effects

$\mathbf{x}_{cs}'^{*} = \underset{\mathbf{x}_{cs}'}{\mathrm{Opt}}(R_{cs}(\mathbf{x}_{cs}'))$

Find Optimal Dimensions Satisfying

$\mathbf{x}_{cem}^{(2)} = \min_{\mathbf{x}_{cem}} \left\| R_{cem}(\mathbf{x}_{cem}) - R_{cs}(\mathbf{x}_{cs}'^{*}) \right\|$

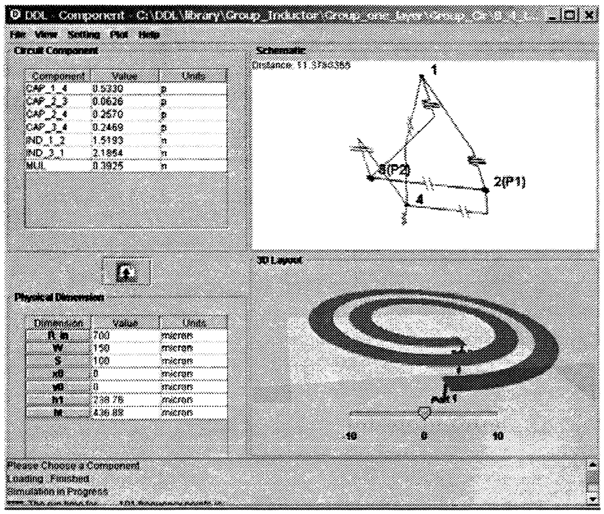
$\mathbf{x}_{c}^{*} = \mathbf{x}_{cem}^{(2)}$

switch to regular ASM

Derivation of Equivalent Circuit:

- ☐ **Equivalent Circuit can be obtained by parameter curve fitting or derivation from EM models**
- ☐ **Curve fitting model**
- ☐ **The derived equivalent circuit must be physically expressive including parasitics**

An example of derived circuit model



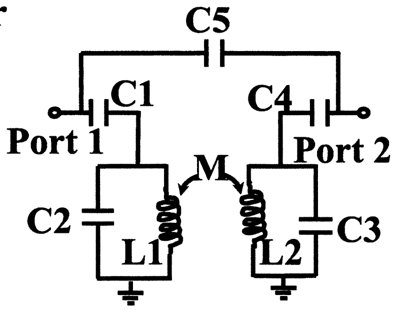
Component	Value	Units
CAP_1_4	0.5330	p
CAP_2_3	0.0626	p
CAP_2_4	0.2570	p
CAP_3_4	0.2469	p
IND_1_2	1.5193	n
IND_3_1	2.1864	n
MUL	0.3925	n

Dimension	Value	Units
R_in	700	micron
W	150	micron
S	100	micron
x0	0	micron
y0	0	micron
h1	238.78	micron
[illegible]	436.88	micron

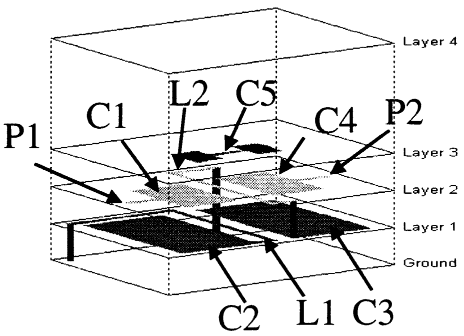
```
Please Choose a Component
Loading . Finished
Simulation in Progress
***** The run time for     101 frequency points is
```

Example 1: A Bandpass Filter

Specs of BPF:
- **Center Frequency: 2.5GHz**
- **Bandwidth: 100MHz**
- **Attenuation: 20dB (1-1.8 GHz)**
  20dB (3-4 GHz)
- **Impedance: 50Ω**



BPF: Schematic Circuit



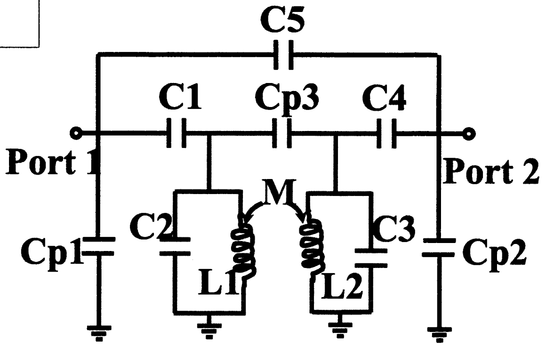
BPF: Physical Layout

With Parasitics:
C1=C4=0.624pF
C2=C3=1.731pF
C5=0.059pF
L1=L2=1.782nH
M=0.257nH
Cp1=0.075pF
Cp2=0.069pF
Cp3=0.062pF

⟺

Without Parasitics:
C1=C4=0.705pF
C2=C3=2.373pF
C5=0.068pF
L1=L2=1.396nH
M=0.164nH

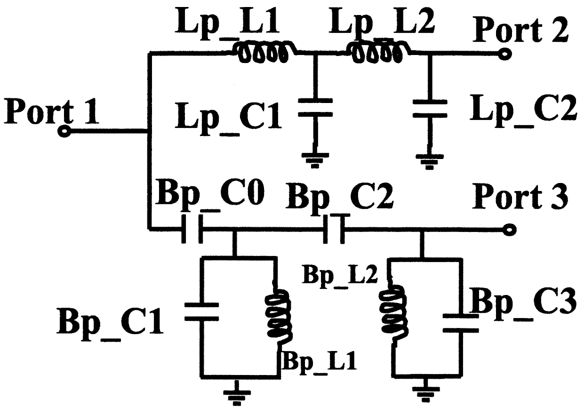
Circuit Model
with parasitics



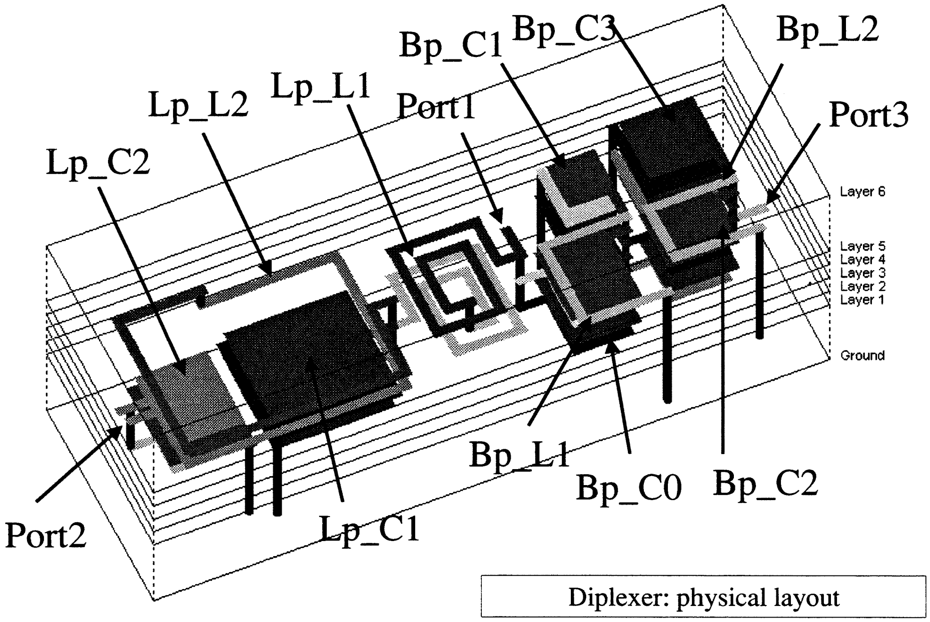
Example 2: Diplexer(LPF+BPF)

Specs of Diplexer:

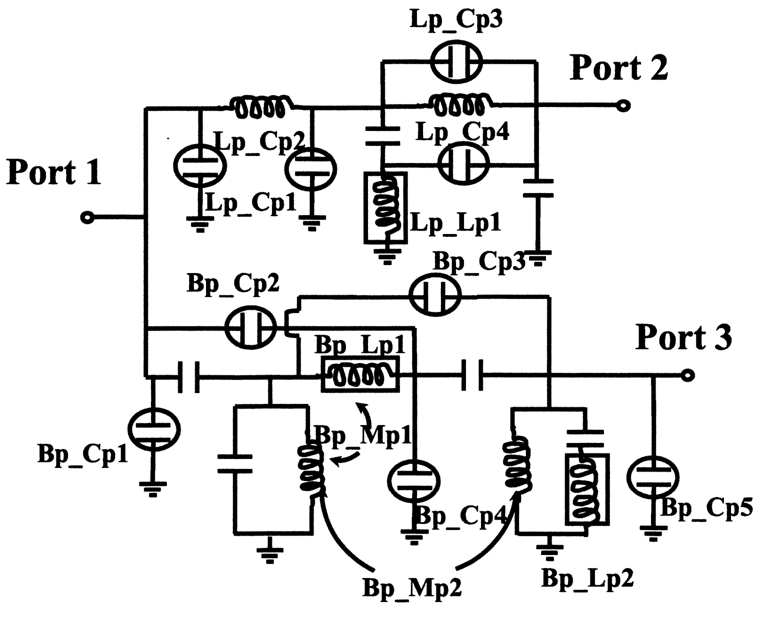
- **Return Loss: 16dB**
  880MHz-960MHz
  1710-1990MHz
- **Isolation: 30dB**
- **Impedance: 50Ω**



Diplexer: Schematic circuit



Diplexer: physical layout



Diplexer: Coarse circuit model with parasitic effects

Without Parasitics:

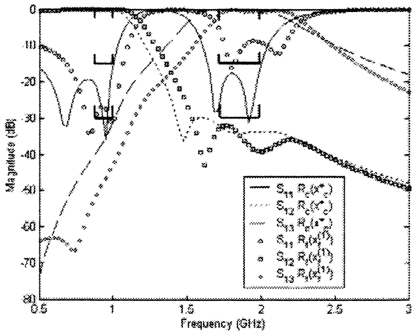
Lp_L1=10.377nH
Lp_L2=10.567nH
Lp_C1=6.893pF
Lp_C2=3.784pF
Bp_C0=3.607pF
Bp_C1=7.811pF
Bp_C2=2.327pF
Bp_C3=6.089pF
Bp_L1=0.776nH
Bp_L2=0.962nH

⇔

With Parasitics:

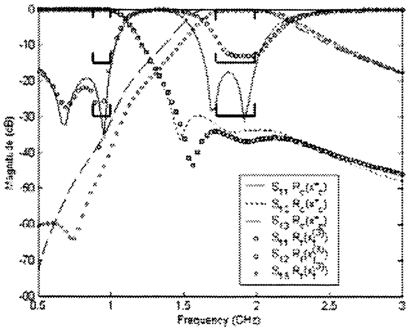
Lp_L1=10.506nH
Lp_L2=10.312nH
Lp_C1=4.489pF
Lp_C2=3.512pF
Bp_C0=2.300pF
Bp_C1=7.870pF
Bp_C2=1.455pF
Bp_C3=4.103pF
Bp_L1=1.667nH
Bp_L2=1.042nH

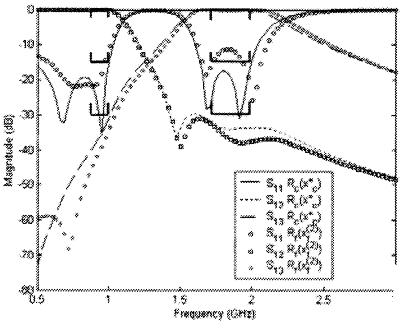
Lp_Cp1=0.404pF
Lp_Cp2=1.917pF
Lp_Cp3=0.093pF
Lp_Cp4=0.244pF
Lp_Lp1=0.398nH
Bp_Cp1=0.404pF
Bp_Cp2=0.203pF
Bp_Cp3=0.20pF
Bp_Cp4=1.477pF
Bp_Cp5=1.225pF
Bp_Lp1=0.541nH
Bp_Lp2=0.307nH
Bp_Mp1=0.089nH
Bp_Mp2=0.103nH



ASM iteration 1

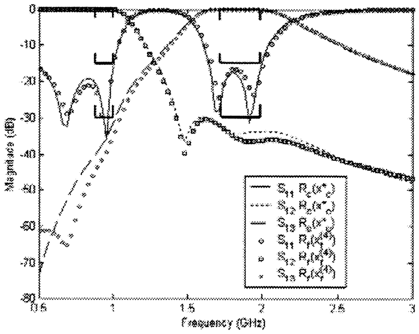
ASM iteration 2





ASM iteration 3

ASM iteration 4



Conclusions

☐ **Effective coarse model with parasitic effects is desirable in ASM**

☐ **Static PEEC model is a good choice in ASM as a coarse model for multilayer RF circuits**

☐ **Dynamic coarse model can be used when strong parasitic effects exist**

Neural Network-Based Device Modeling and Design Optimization

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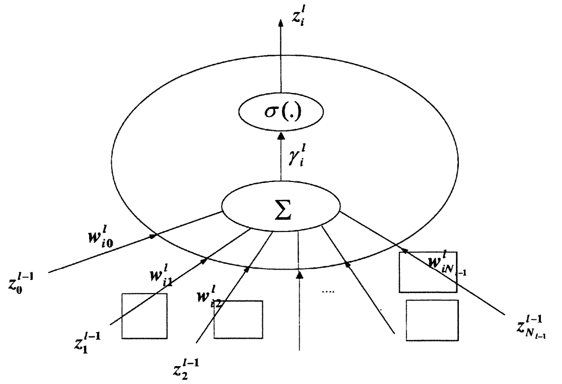
Introduction

- **Neural networks are inspired the study of human brains.**
- **They have the potential to become powerful modeling tools.**
- **They are recently applied to RF and microwave modeling**
- **They have been used to model EM problems, active devices and nonlinear circuits.**
- **These models can be used for efficient circuit design optimization**

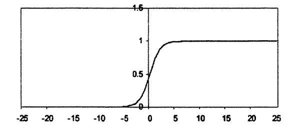
Information Processing In a Neuron



Activation Functions in Neurons

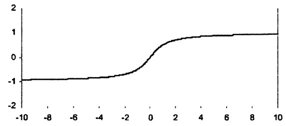
Sigmoid

$\sigma(\gamma)=1/(1+e^{-\gamma})$



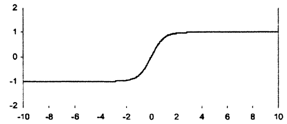
Arc-tangent

$\sigma(\gamma)=(2/\pi)arctan(\gamma)$

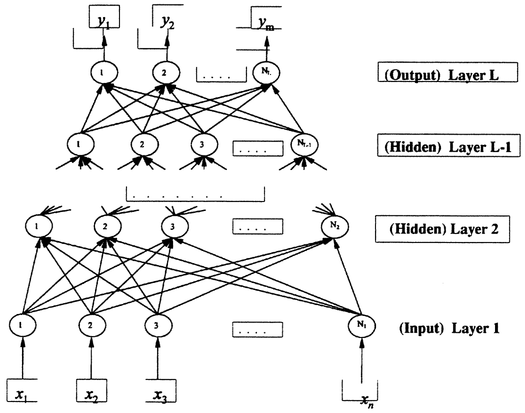


Hyperbolic-tangent

$\sigma(\gamma)=(e^{+\gamma}-e^{-\gamma})/(e^{+\gamma}+e^{-\gamma})$

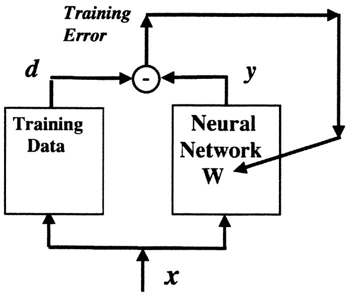


Multilayer Perceptron (MLP) Neural Network



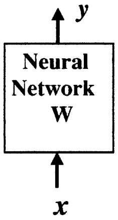
Neural Network Learning
(Train a Neural Network)

Modify the internal weights of the neural network such that the neural network input-output relationship matches that of training data.

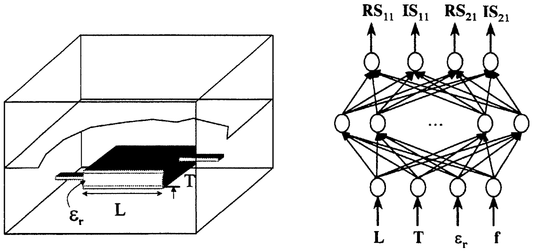


Neural Network Recall
(Use of Neural Network)

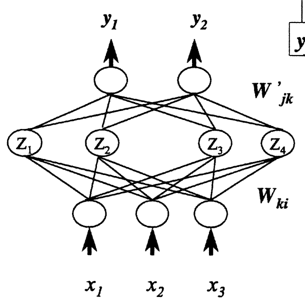
Given input information *x*, the neural network computes the output *y*. During recall, internal weights of the neural network are fixed as constants.



Embedded Capacitor Neural Model



3 Layer MLP: Feedforward Computation



Outputs

↑

$$y_j = \sum_k W'_{jk} Z_k$$

↑

Hidden Neuron Values

$$Z_k = \tanh\left(\sum_i W_{ki} x_i\right)$$

↑

Inputs

How can ANN represent an arbitrary nonlinear input-output relationship?

Universal Approximation Theorem
(Hornik, StinchCombe and White, 1989)

In plain words:

Given enough hidden layer neurons, a 3-layer MLP neural network can approximate an arbitrary continuous multidimensional function to any desired accuracy

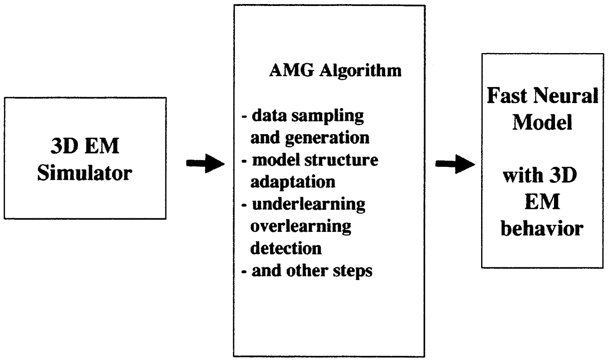
Neural Networks for RF/Microwave Design

- **Neural models are efficient alternatives to closed-form expressions, equivalent circuit models and look-up tables**
- **Neural network models can be developed from measured or simulated data**
- **Neural models can also be used to update or improve the accuracy of already existing models**
- **Neural network models have been developed for active devices, passive components, interconnect networks and nonlinear circuits**
- **These models have been used in circuit simulators for high-level simulation, design and optimization**

Features of Neural Network Models

- **Neural networks have the ability to model multi-dimensional nonlinear relationships**
- **Neural models are simple and the model computation is fast**
- **Neural networks can learn and generalize from available data thus making model development possible even when component formulae are unavailable**
- **Neural network approach is generic, i.e., the same modeling technique can be re-used for passive/active devices/circuits**
- **It is easier to update neural models whenever device or component technology changes**

Automatic Model Generation (AMG)



3D EM Simulator

AMG Algorithm

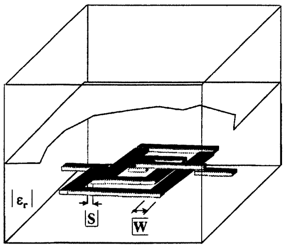
- **data sampling and generation**
- **model structure adaptation**
- **underlearning overlearning detection**
- **and other steps**

Fast Neural Model

with 3D EM behavior

Automatic Generation of Neural Model
Spiral Inductor Example

(Devabhaktuni et al. 2003)



Model Inputs: Width (w), space (s), dielectric constant (ε_r), and frequency (f).

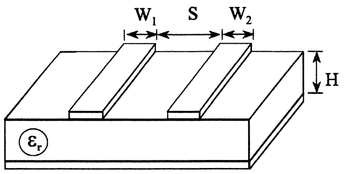
Model Outputs: Real and imaginary parts of S-parameters.

Data Generators used for training: *Sonnet EM* and *Ansoft-HFSS* simulators.

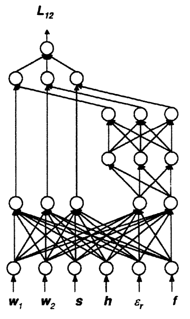
Neural Model Accuracies using 386 3D EM Data

Neural Modeling Technique	Validation Error
Conventional training	6.25%
Basic AMG algorithm	2.34%
Knowledge based AMG-DM	0.96%
Knowledge-based AMG-KBNN	1.12%
Knowledge-based AMG-PKI	0.85%
Knowledge-based AMG-SMNN	0.99%

Design Solution Space Analysis

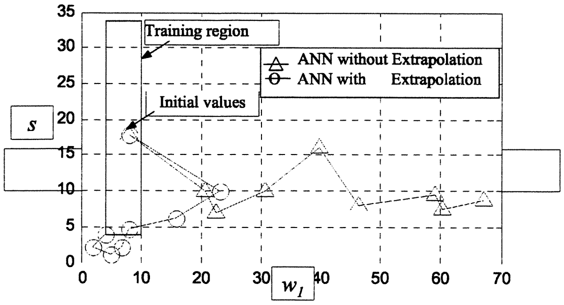


Microstrip Line



Design Solution Space Analysis of Transmission Lines Using Optimization with Neural Model

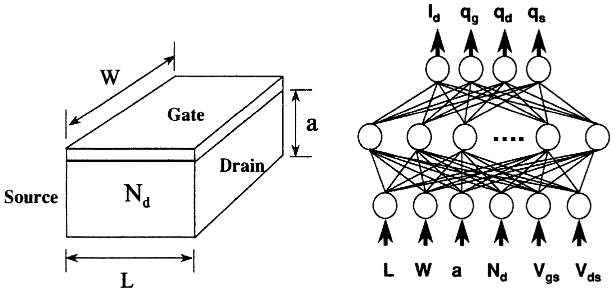
(Xu et al 2004)



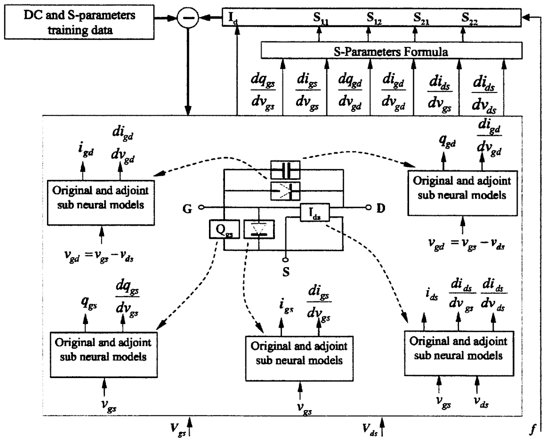
The optimization trajectory of design parameters w_1 and s

Neural Networks for Nonlinear Device/Circuit Modeling

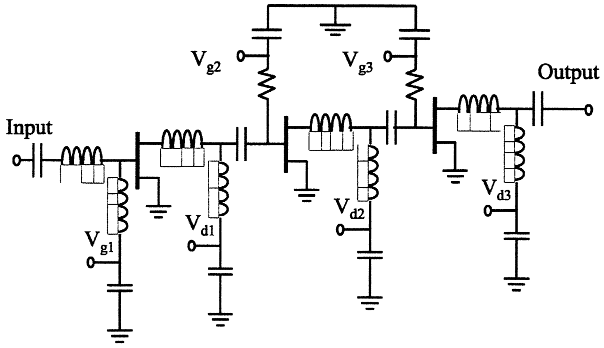
Neural Model for MESFET Modeling: I-Q Model Trained by I-Q Data



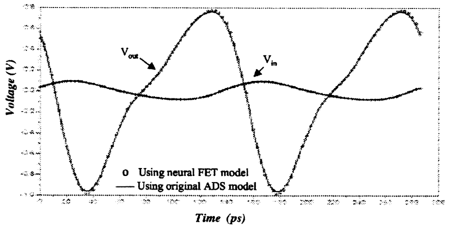
Nonlinear Modeling Using Adjoint Neural Network (Xu et al. 2001)

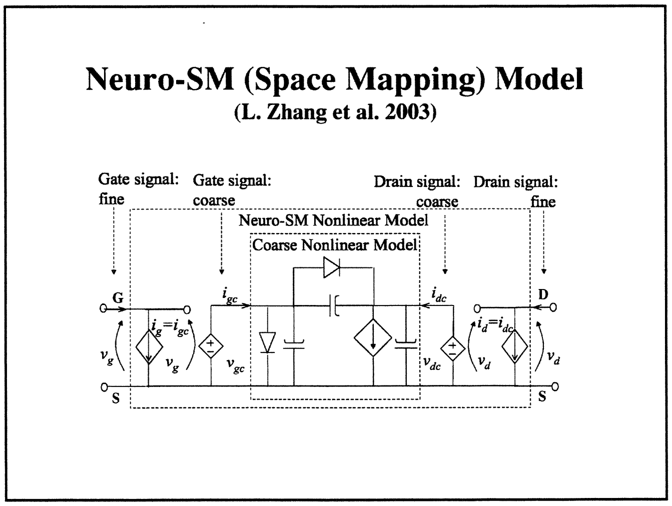


Three-Stage X-Band Amplifier Circuit with 3 FETs Represented By Neural Models



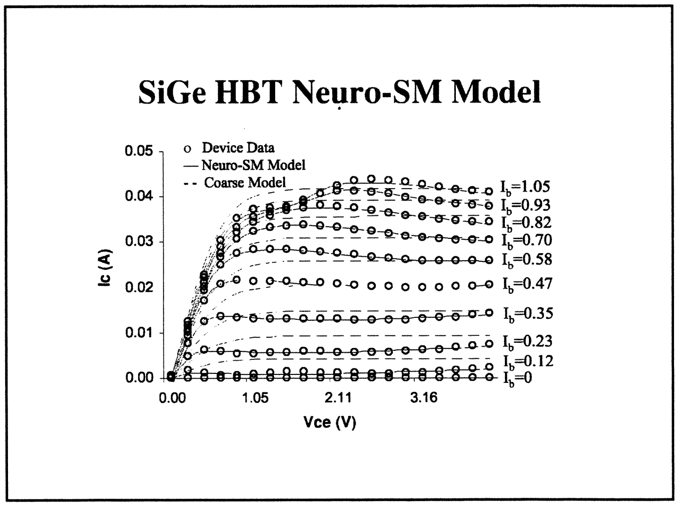
Large-Signal Simulation of a 3-Stage Amplifier using Neural based FET Model





Neuro-SM (Space Mapping) Model

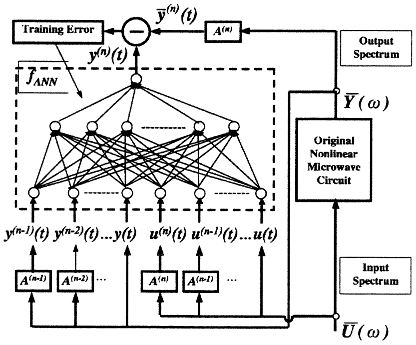
(L. Zhang et al. 2003)



SiGe HBT Neuro-SM Model

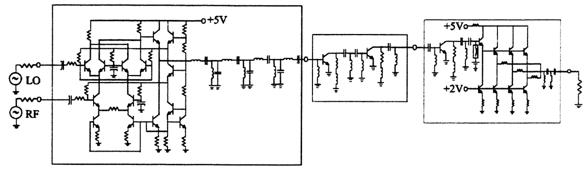
Time-Domain Model Trained by Time- or Frequency-Domain Data (DNN)

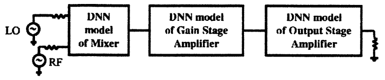
(Xu et al. 2002)



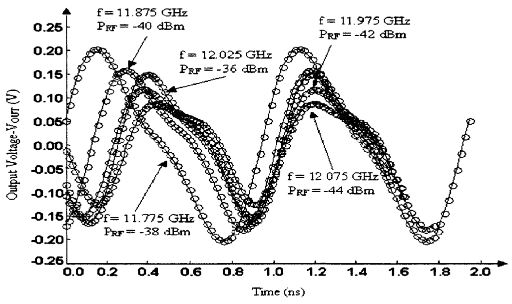
* where A is Inverse Fourier Transform

Nonlinear Simulation of DBS Receiver System Using DNN



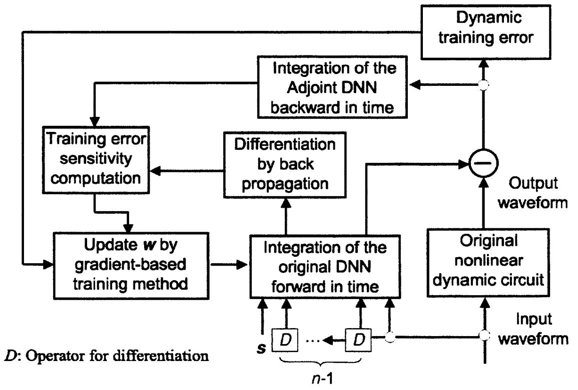


DBS Receiver System Simulation using Original Circuits and DNN Models



Adjoint DNN Method for DNN Training

(Cao et al 2003)



Dynamic training error

Integration of the Adjoint DNN backward in time

Training error sensitivity computation

Differentiation by back propagation

Output waveform

Update ***w*** by gradient-based training method

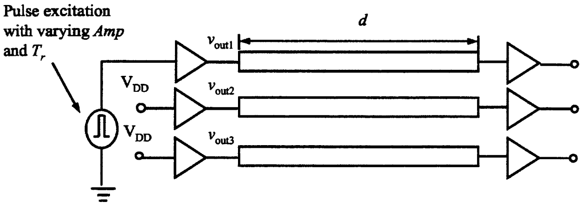
Integration of the original DNN forward in time

Original nonlinear dynamic circuit

Input waveform

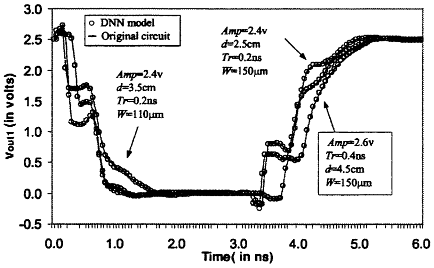
D: Operator for differentiation

Coupled 3-conductor Interconnect Circuit Terminated with Nonlinear Buffers

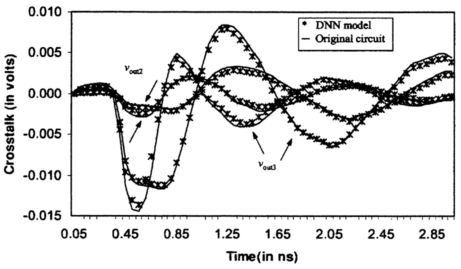


Amp: Pulse amplitude
T_r: Pulse rise-time
d: Interconnect length

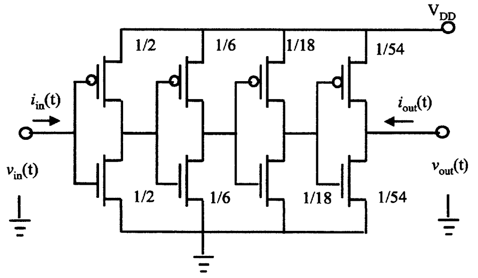
Comparison of Signal Propagation Waveforms v_{out1} of 3-Coupled Interconnect Circuit



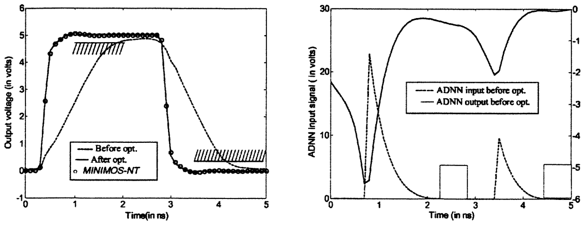
Comparison of Crosstalks v_{out2} and v_{out3} of 3-Coupled Interconnect Circuit



Transient DNN Modeling of a 4-stage CMOS Driver Example

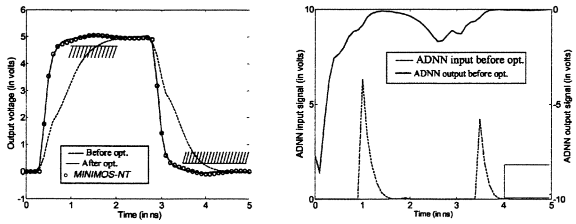


Responses of the 4-Stage CMOS Driver Models Before and After Optimization



	CPU Time for Optimization
ADNN-Based Optimization	8 seconds
Direct Optimization by Driving MINIMOS-NT	10.5 hours

Optimization of the 4-Stage CMOS Driver with Transmission Line Load, Signal in Original and Adjoint DNN



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