High Frequency Characterization of Transistors

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Process Engineer

VP Engineering

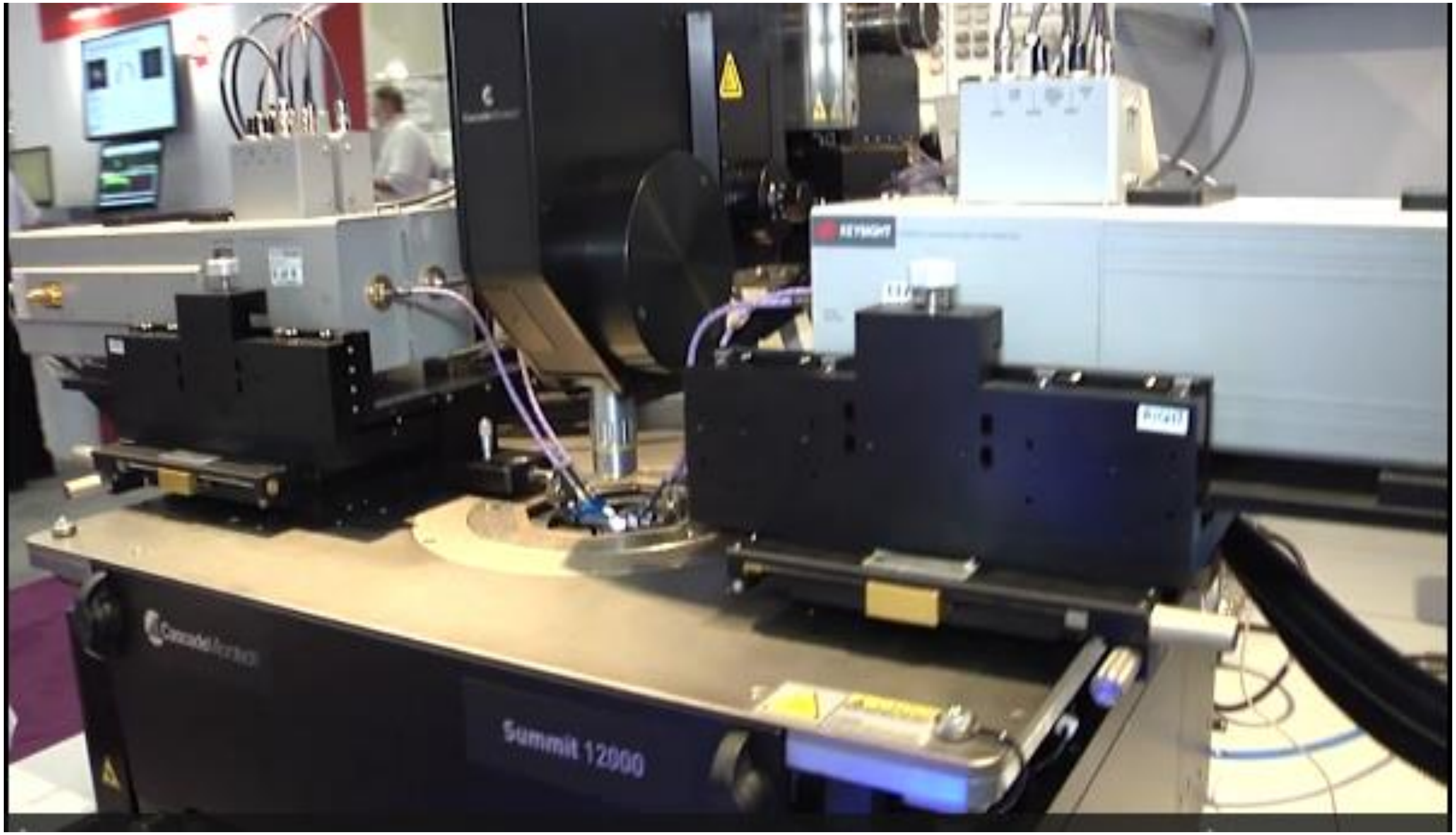
This process change should have improved Ft and Fmax !

I didn't see any change. Actually Fmax was lower!

VP Technology

Microwave Test Engineer

1.1THz Test System from Cascade Microtech/Keysight



International Microwave Symposium, San Francisco, May 22-27, 2016

What is High Frequency Characterization?

Small Signal Characterization	Large Signal Characterization
f_T	P_{in} vs P_{out}
f_{max}	PAE
NF_{min}	P_{1dB}
R_n	$OIP3$
$Zs\ opt$	$IIP3$
G_A	$Load\ Pull$
Re	$Source\ Pull$
Rb	
Rs	
Rd	
Linear Measurements	Non-Linear Measurements

This talk → f_T, f_{max}

OUTLINE

- Introduction
- S-parameters
- Smith Chart
- Vector Network Analyzer
- Calibration
- De-embedding
- Examples of measured data
- Gain and Stability
- Ft and Fmax from S-parameters
- Mason's Gain
- Transistor Specmanship
- Parameter extraction/wafer maps
- Conclusion

BIPOLAR FIGURES OF MERIT

Approximate

$$f_T = \frac{g_m}{2\pi\left(C_\pi + C_\mu\right)}$$

$$C_\pi = C_{jE} + g_m \frac{W^2}{2D_n}$$

$$C_\mu = C_{jC}$$

$$f_{\max} = \sqrt{\frac{f_T}{8\pi r_{bb'} C_\mu}}$$

Exact

$$\frac{1}{2\pi f_T} = \underbrace{\frac{N_{AB}}{N_{DE}}\left(\frac{W_E W}{2D_n}\right)}_{\tau_e} + \underbrace{\frac{W^2}{2D_n} + \frac{W}{v_m}}_{\tau_b} + \underbrace{\frac{W_{DC}}{2v_{sat}}}_{\tau_c} + r_e C_{jE} + \left(r_e + r_{ee} + r_c\right) C_{jC}$$

thermionic emission velocity

$$v_m = \sqrt{\frac{kT}{2\pi m^*}} \approx 5\times10^6 \text{ cm/sec for Si.}$$

saturation velocity

$$v_{sat} = 10^7 \text{ cm/sec for Si.}$$

$$f_{\max} = \sqrt{\frac{f_T}{8\pi r_{bb'} C_\mu}}$$

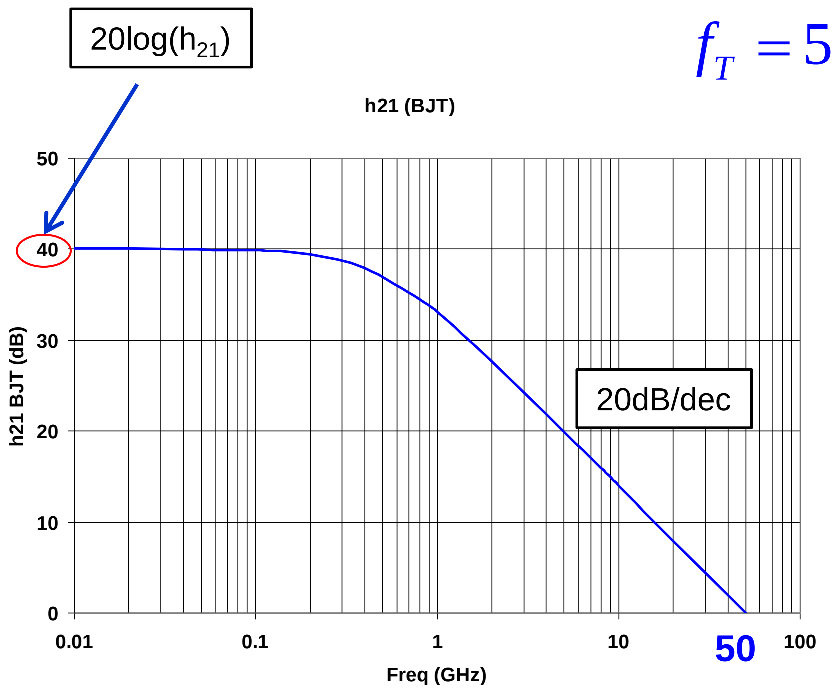
Ft VARIATION WITH FREQUENCY (Bipolar)

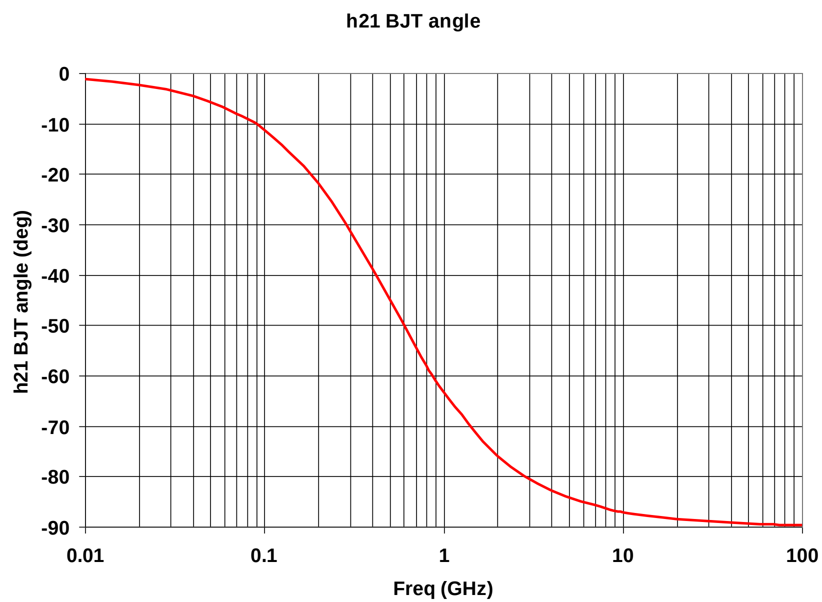
$$h_{21(BJT)} = \frac{\beta}{1 + j\beta \frac{f}{f_T}}$$

Microwave Engineers find Ft by plotting h_{21} vs frequency!

$$\beta_{DC} = 100$$

$$f_T = 50 \text{ GHz}$$





MOS FIGURES OF MERIT

Approximate

$$f_T = \frac{g_m}{2\pi\left(C_{gs} + C_{gb} + C_{gd}\right)}$$

$$f_T \approx \frac{g_m}{2\pi C_{gs}}$$

C_{gd} & C_{gb} are small in sat

$$f_{\max} = \sqrt{\frac{f_T}{8\pi R_G C_{gd}}}$$

Exact

$$f_T = \frac{g_m}{2\pi\left[(C_{gs} + C_{gd})\left(1 + \frac{R_D + R_S}{r_o}\right) + C_{gd} g_m (R_D + R_S) + C_p\right]}$$

C_p: parasitic cap

$$f_{\max} = \frac{f_T}{2\sqrt{g_{ds}(R_G + R_S) + 2\pi f_T R_G C_{gd}}}$$

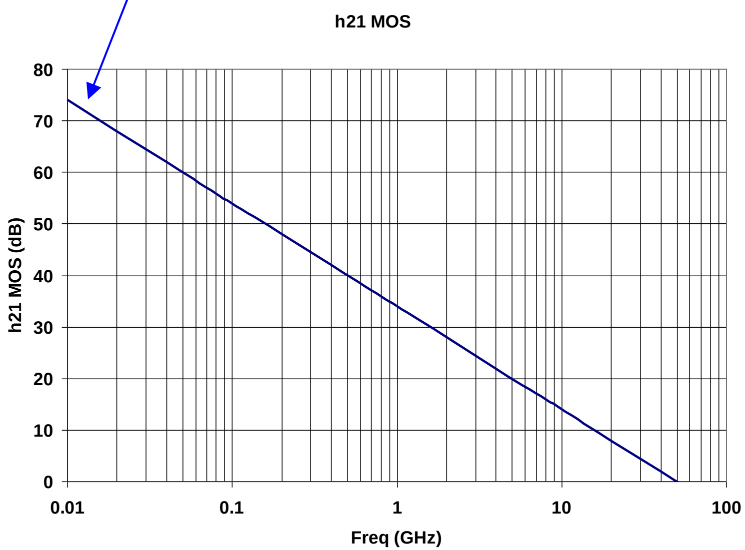
Ft VARIATION WITH FREQUENCY (MOS)

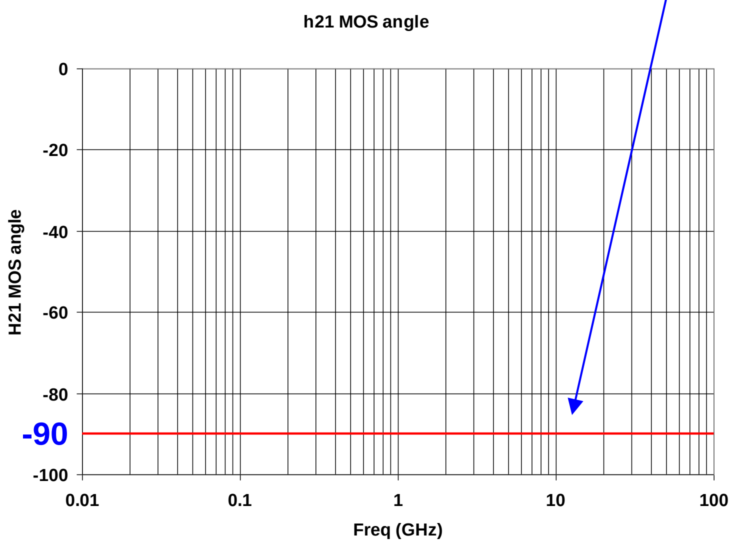
$$h_{21(MOS)} = \frac{1}{j\,\dfrac{f}{f_T}}$$

$$f_T = 50\ \text{GHz}$$

Note that this goes to infinity!

Phase is constant



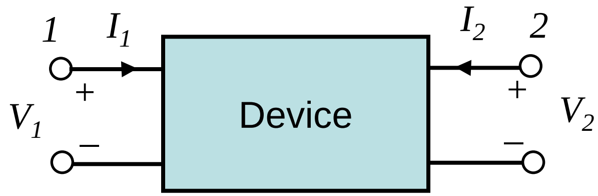


S- PARAMETERS

Why S-Parameters?

- At HF, difficult to measure currents and voltages
- Difficult to create open and shorts
- Everything behaves like Transmission lines with reflections
- S-parameters are very easy to understand and use
- S-parameters exist for any network
- Can easily relate to gain, loss, reflection and power
- Can predict the performance of cascaded networks
- From S-parameters, one can convert to Z, Y or H parameters
- Needed for SPICE model parameter extraction
- Some CAD programs need S-parameters for circuit design

Two Port Parameters



$$V_1 = z_{11}I_1 + z_{12}I_2$$
$$V_2 = z_{21}I_1 + z_{22}I_2$$

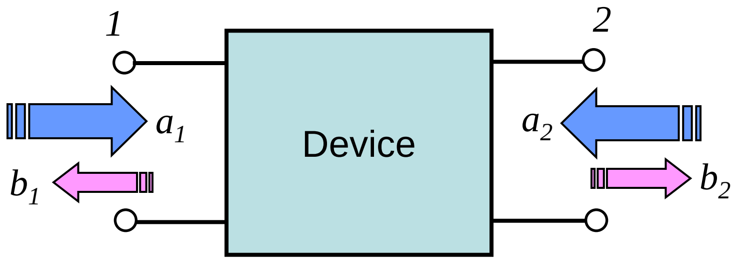
$$I_1 = y_{11}V_1 + y_{12}V_2$$
$$I_2 = y_{21}V_1 + y_{22}V_2$$

$$V_1 = h_{11}I_1 + h_{12}V_2$$
$$I_2 = h_{21}I_1 + h_{22}V_2$$

$$I_1 = g_{11}V_1 + g_{12}I_2$$
$$V_2 = g_{21}V_1 + g_{22}I_2$$

$$V_1 = AV_2 - BI_2$$
$$I_1 = CV_2 - DI_2$$

S-Parameters Defined



$$b_1 = S_{11}a_1 + S_{12}a_2$$

$$b_2 = S_{21}a_1 + S_{22}a_2$$

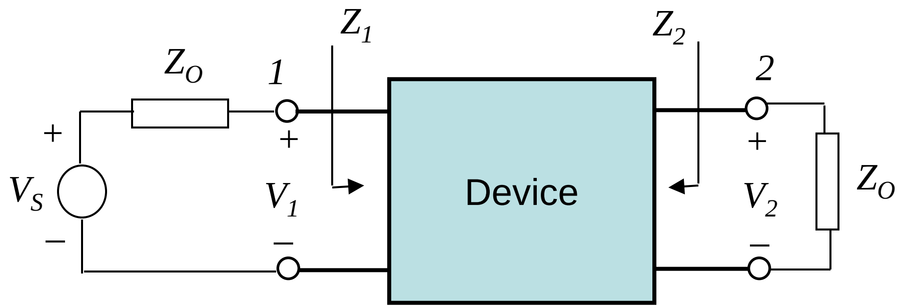
$$S_{11} = \left.\frac{b_1}{a_1}\right|_{a_2=0}$$ Input reflection coefficient with output terminated in Zo

$$S_{21} = \left.\frac{b_2}{a_1}\right|_{a_2=0}$$ Forward transmission coefficient with output terminated in Zo

$$S_{12} = \left.\frac{b_1}{a_2}\right|_{a_1=0}$$ Reverse transmission coefficient with input terminated in Zo

$$S_{22} = \left.\frac{b_2}{a_2}\right|_{a_1=0}$$ Output reflection coefficient with input terminated in Zo

S-Parameters in terms of impedances and voltages



Z_O: Characteristic impedance (50Ω)

$$S_{11} = \frac{Z_1 - Z_0}{Z_1 + Z_0}$$

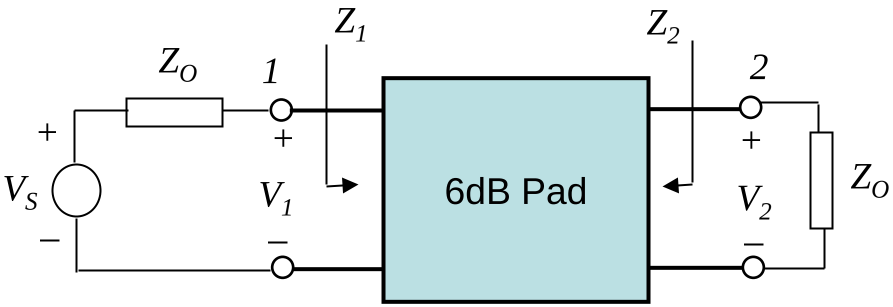
$$S_{22} = \frac{Z_2 - Z_0}{Z_2 + Z_0}$$

$$S_{21} = \frac{2V_2}{V_S}$$

$$S_{12} = \frac{2V_1}{V_S}$$

G.Gonzalez, Microwave Transistor Amplifiers, Prentice Hall 1984

S-Parameters for a 6dB pad



$$S_{11} = \frac{Z_1 - Z_0}{Z_1 + Z_0}$$

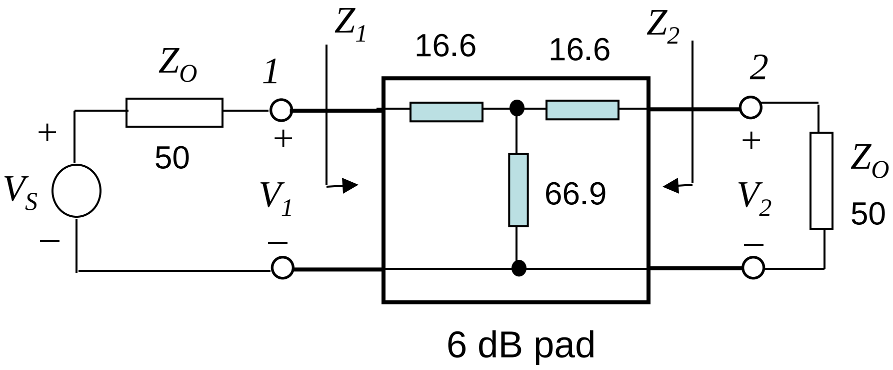
$$S_{22} = \frac{Z_2 - Z_0}{Z_2 + Z_0}$$

$$S_{21} = \frac{2V_2}{V_S}$$

$$S_{12} = \frac{2V_1}{V_S}$$

$$S = \begin{pmatrix} 0 & 0.5 \\ 0.5 & 0 \end{pmatrix}$$

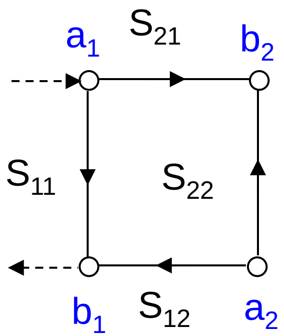
Z-Parameters for a 6dB pad

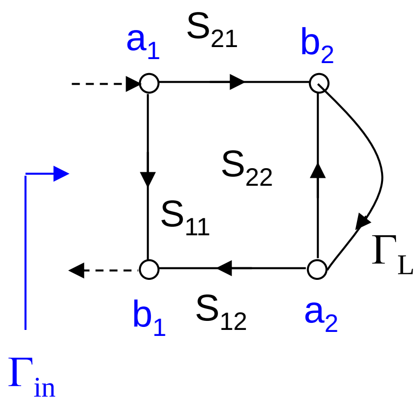


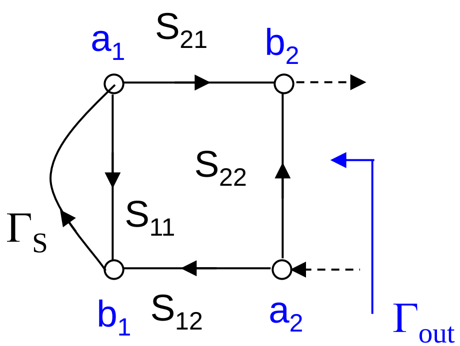
$$Z = \begin{bmatrix} 83.5 & 66.9 \\ 66.9 & 83.5 \end{bmatrix} \qquad S = \begin{bmatrix} 0 & 0.5 \\ 0.5 & 0 \end{bmatrix}$$

Looking at Z - parameters
one can not quickly infer
this is a 6dB Pad!

SIGNAL FLOW GRAPHS







$$b_1 = S_{11}a_1 + S_{12}a_2$$
$$b_2 = S_{21}a_1 + S_{22}a_2$$

$$\Gamma_{in} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L}$$

$$\Gamma_{out} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S}$$

INTRODUCTION TO SMITH CHART

Mapping of resistances – Smith Chart-1

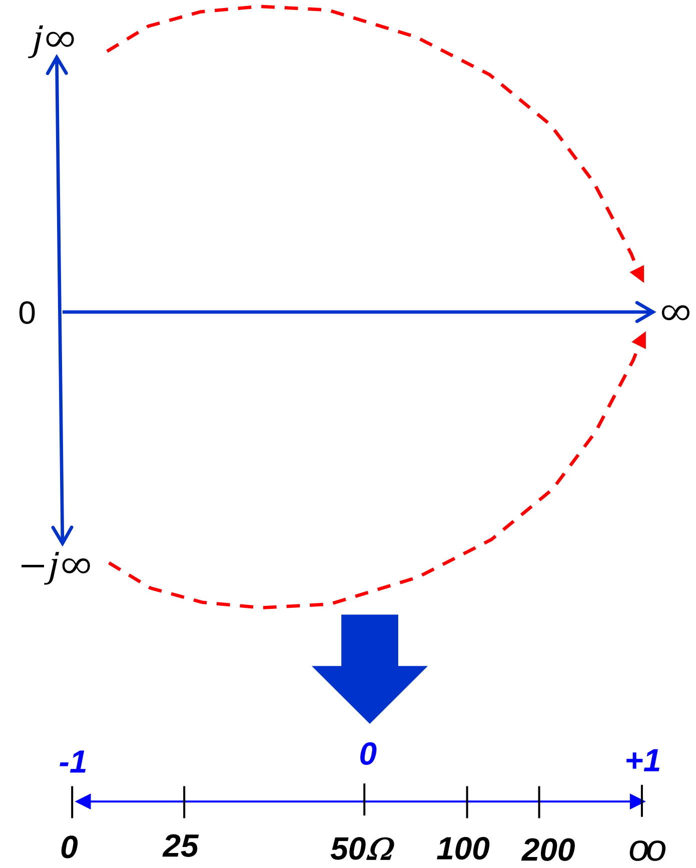
$$\Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0}$$

$$\Gamma_S = \frac{Z_S - Z_0}{Z_S + Z_0}$$

Convert all impedances to reflection coefficient and plot it. That is Smith Chart!

$Z_O = 50\Omega$

Z_L ohms	Γ_L
0	-1.00
25	-0.33
50	**0.00**
100	0.33
200	0.60
inf	1.00



Pure resistances map along the x-axis between -1 and + 1

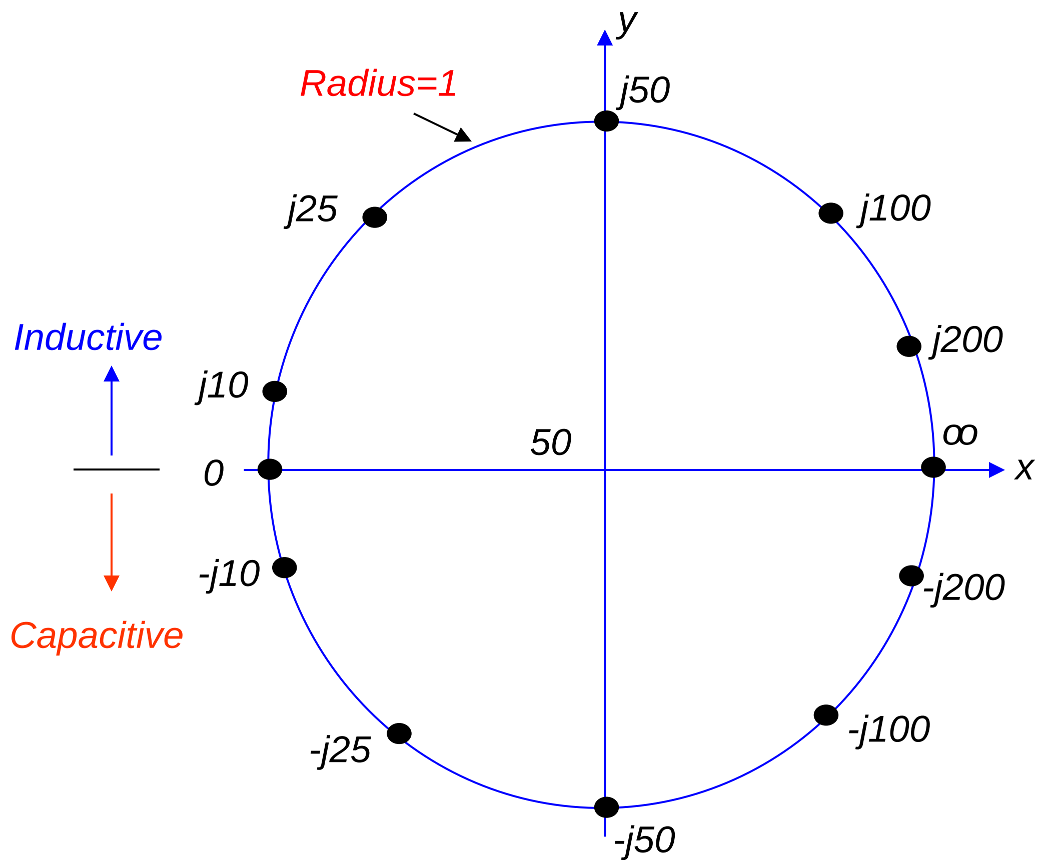
Mapping of reactances – Smith Chart-2

$$\Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0}$$

$$\Gamma_S = \frac{Z_S - Z_0}{Z_S + Z_0}$$

$Z_O = 50\Omega$

$Z_L(\Omega)$	Γ_L
0	$1\angle 180^o$
j10	$1\angle 157.4^o$
j25	$1\angle 126.9^o$
j50	$1\angle 90^o$
j100	$1\angle 53.1^o$
j200	$1\angle 28.1^o$
j∞	$1\angle 0^o$



Note that magnitude is always 1 but angle varies.
Pure reactance maps along the circumference of a unit circle.

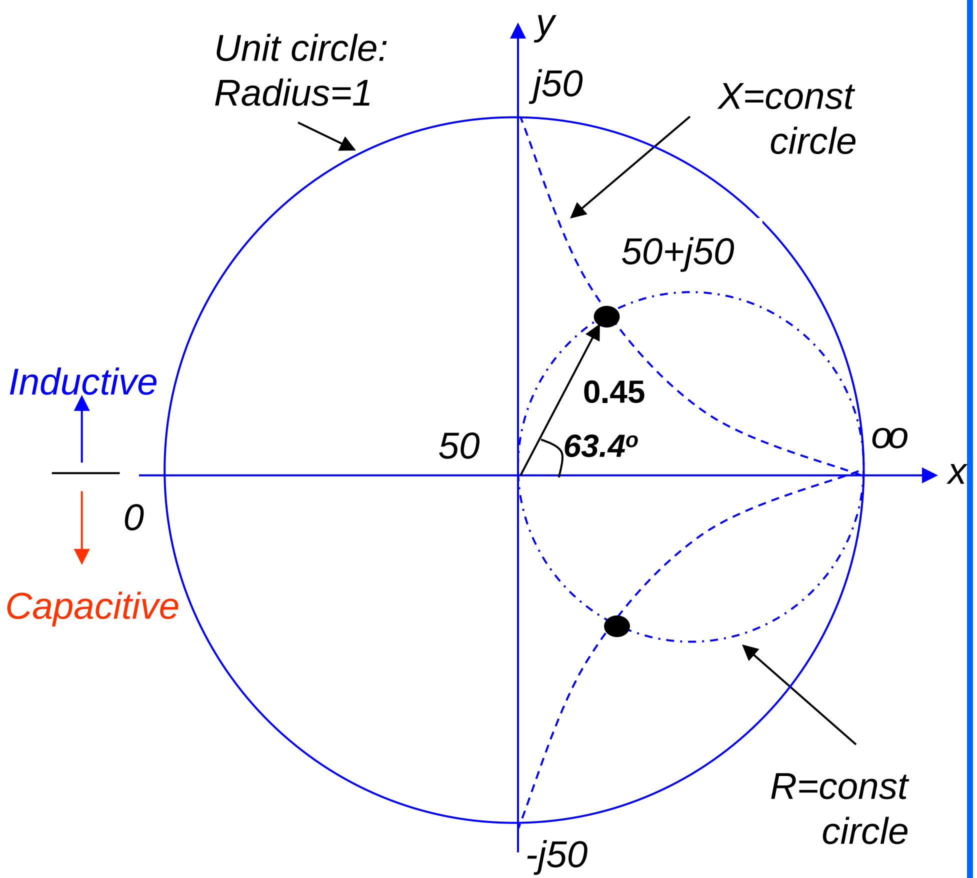
Mapping of impedances – Smith Chart-3

$$\Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0}$$

$$\Gamma_S = \frac{Z_S - Z_0}{Z_S + Z_0}$$

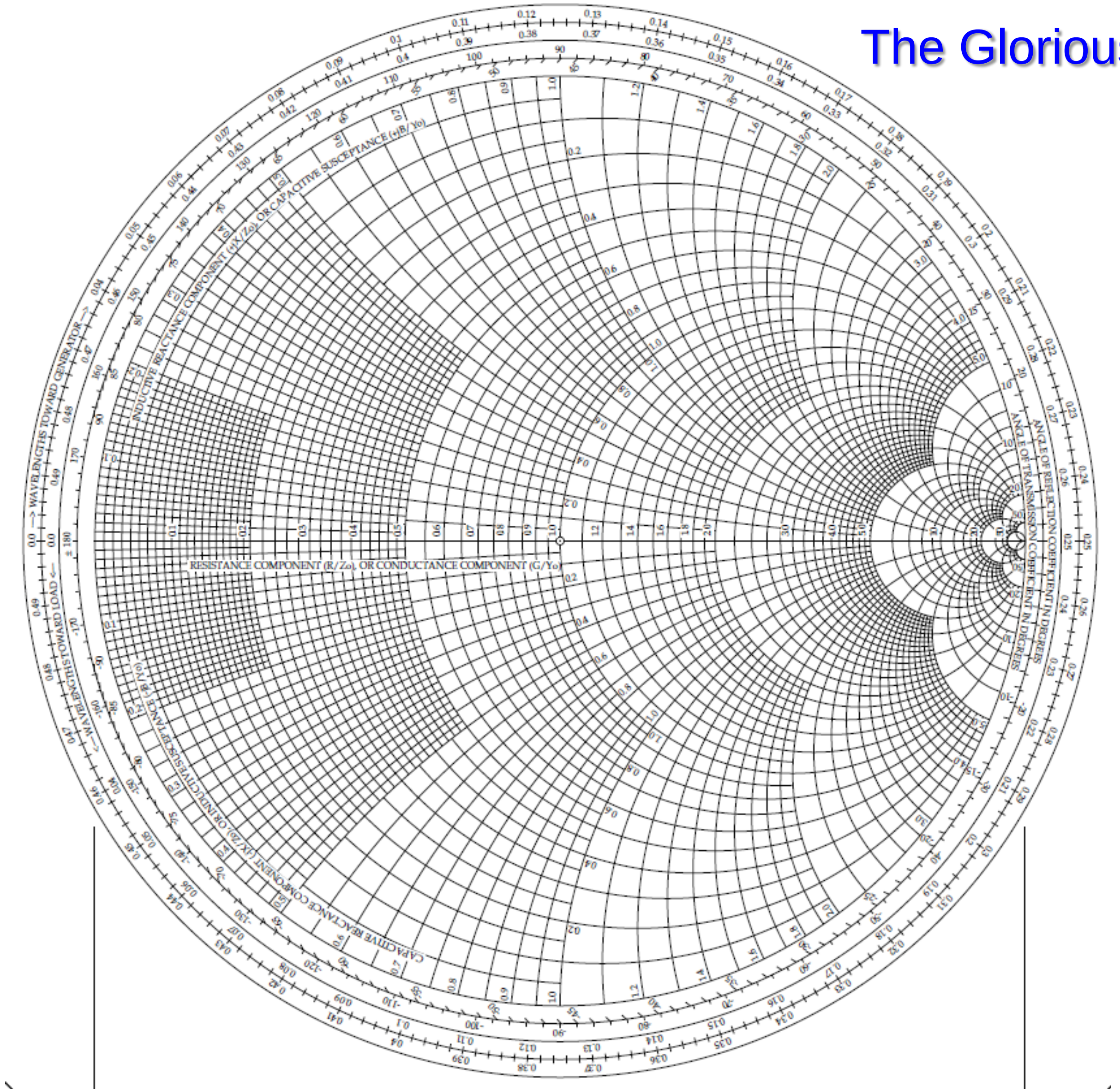
$Z_O = 50\Omega$

$Z_L(\Omega)$	Γ_L
50+j50	$0.45\angle 63.4^o$
$50 - j50$	$0.45\angle -63.4^o$



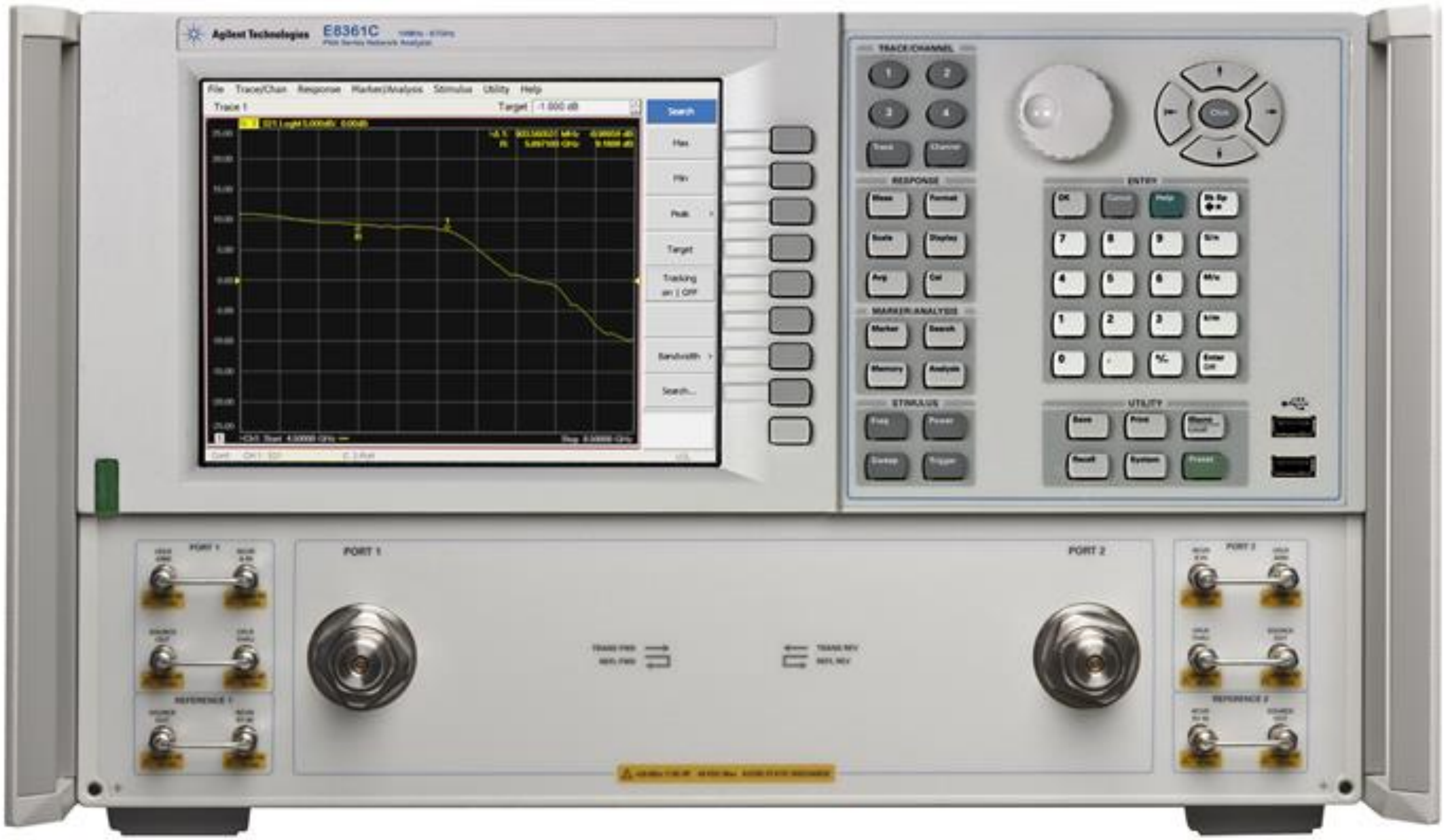
All impedances (R+jX) with R>0 will map inside a unit circle.
If R is negative (R< 0), it will map outside the unit circle.

The Glorious Smith Chart



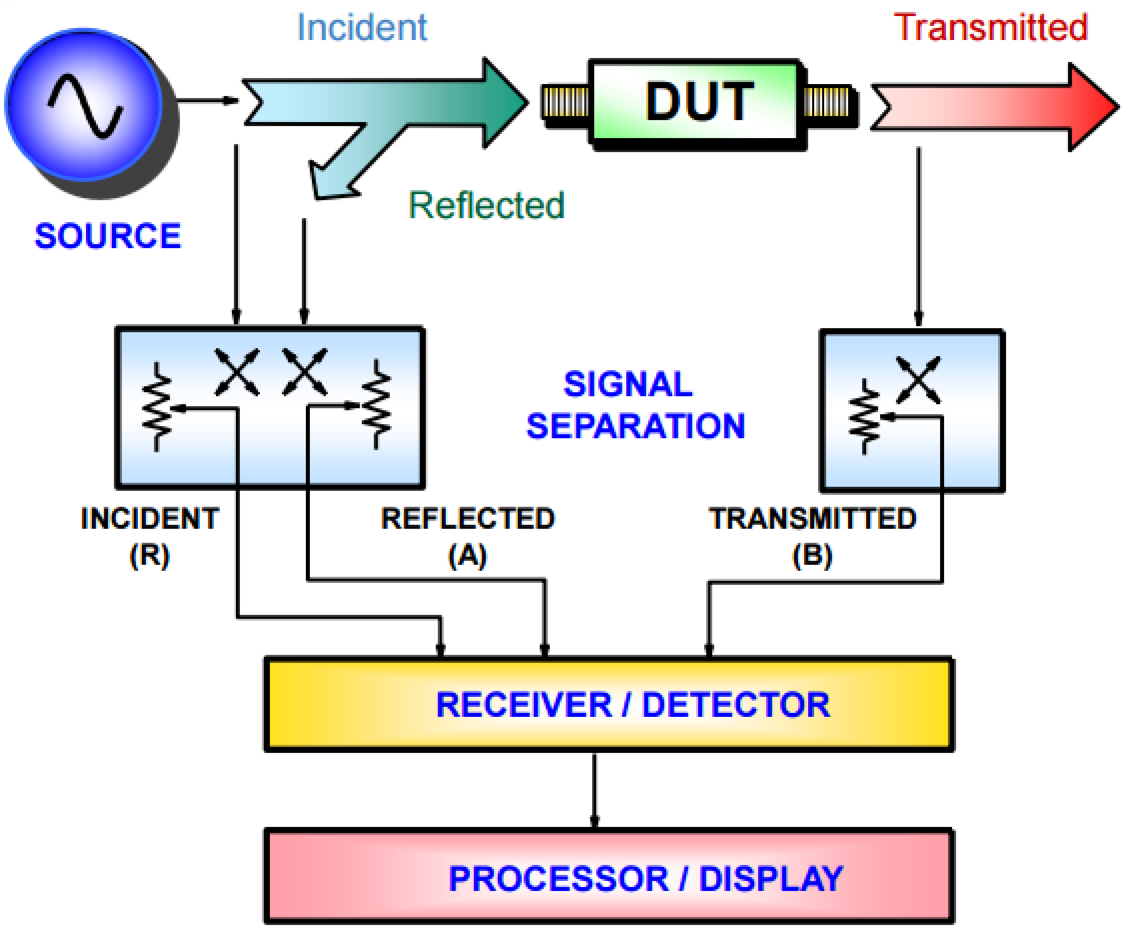
How to measure S-parameters?
- The Vector Network Analyzer

E 8361C PNA Series NETWORK ANALYZER 10MHz – 67GHz



You will also need a 4155 Semiconductor Parameter Analyzer for biasing.

VECTOR NETWORK ANALYZER BLOCK DIAGRAM



What is Calibration?

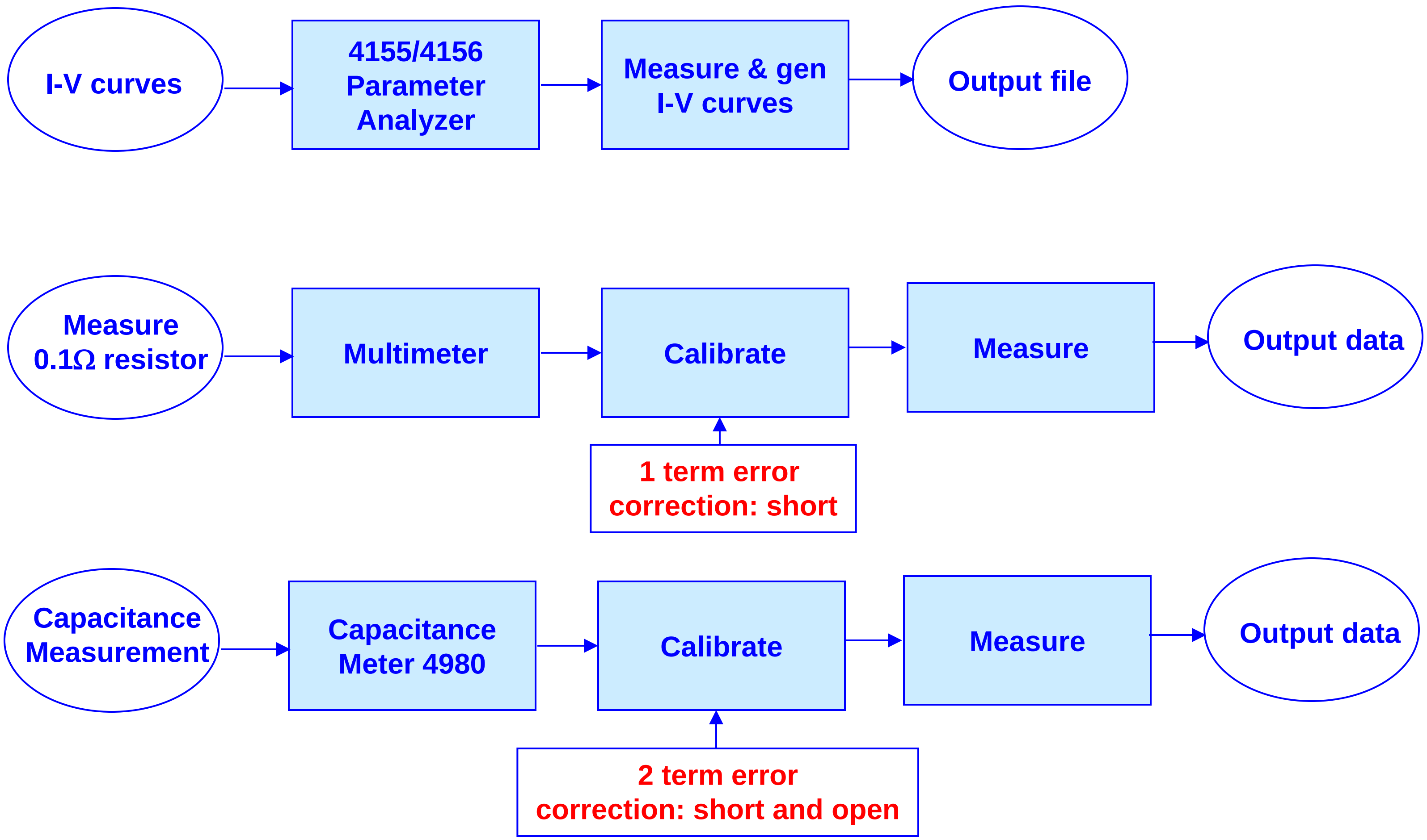
CALIBRATION

Short the leads and adjust the ZERO OHMS pot so that the meter reads zero. We have zeroed out resistance of the test leads.

CALIBRATION

Short the leads and write down the reading R1.
Connect the resistor Rx and take the reading R2.

Unknown resistor Rx = R2 – R1

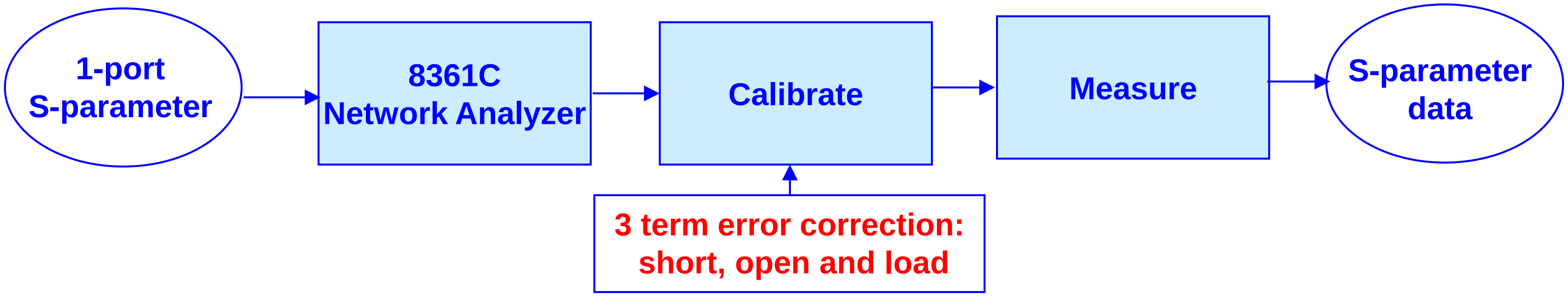

I-V curves
4155/4156 Parameter Analyzer
Measure & gen I-V curves
Output file
Measure 0.1Ω resistor
Multimeter
Calibrate
Measure
Output data
1 term error correction: short
Capacitance Measurement
Capacitance Meter 4980
Calibrate
Measure
Output data
2 term error correction: short and open

4980A CAPACITANCE METER (20Hz – 2MHz)

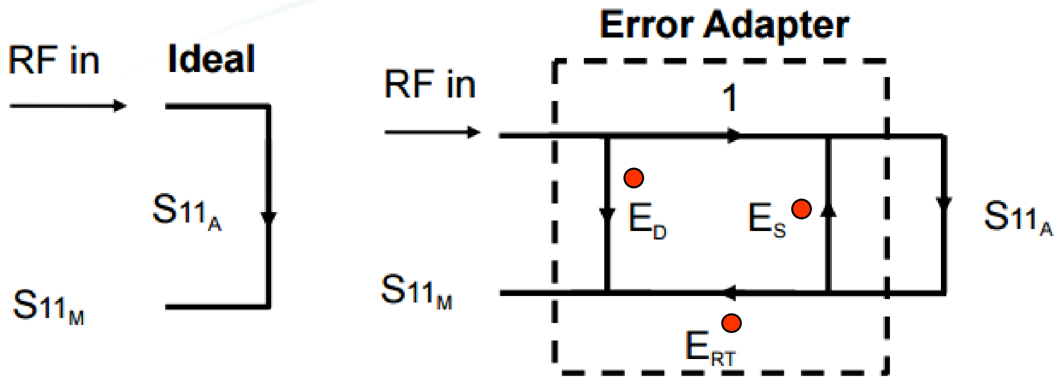
CALIBRATION

Short the leads and write down the series resistance Rs.
Open the leads and write down the stray capacitance Cp.
The instrument does the correction for the series resistance.
Then it subtracts the stray capacitance from the measured data.

One-Port Measurement using Network Analyzer



One-Port Error Correction



E_D = Directivity

E_{RT} = Reflection tracking

E_S = Source Match

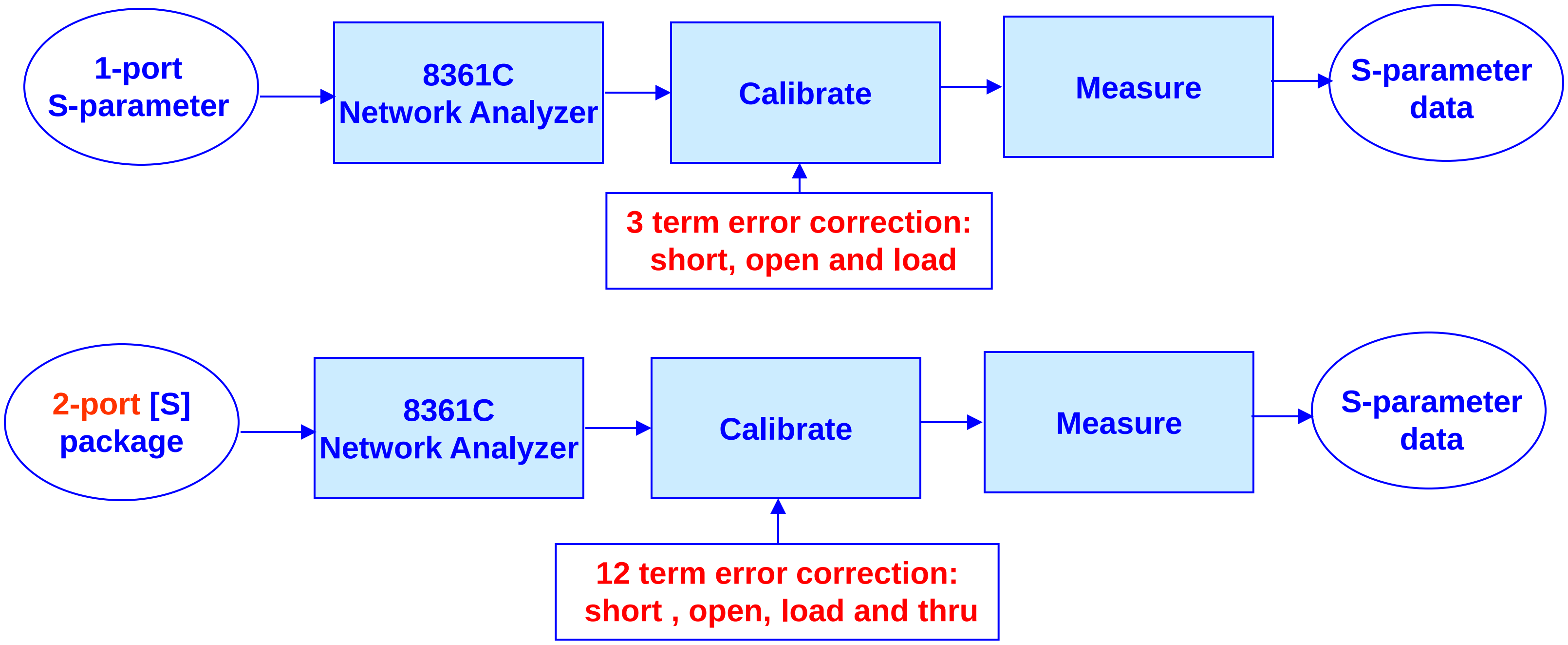
$S11_M$ = Measured

$S11_A$ = Actual

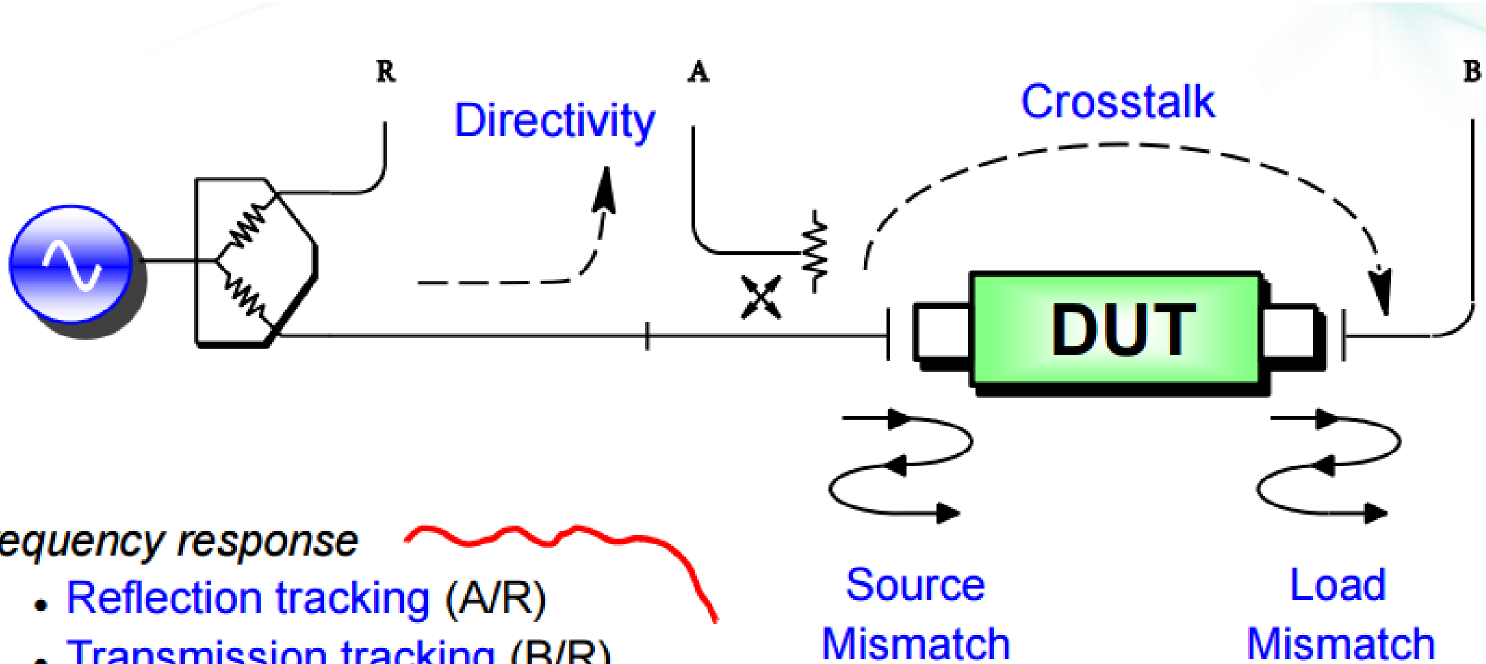
To solve for error terms, we measure 3 standards to generate 3 equations and 3 unknowns

$$S_{11M} = E_D + E_{RT}\left[\frac{S_{11A}}{1 - E_S\, S_{11A}}\right]$$

One / Two-Port Measurement using Network Analyzer



TWO PORT CALIBRATION



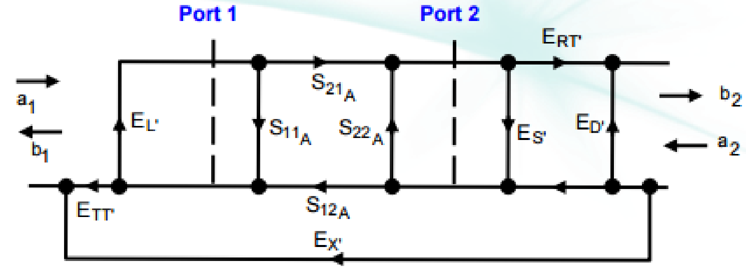
Frequency response

- Reflection tracking (A/R)
- Transmission tracking (B/R)

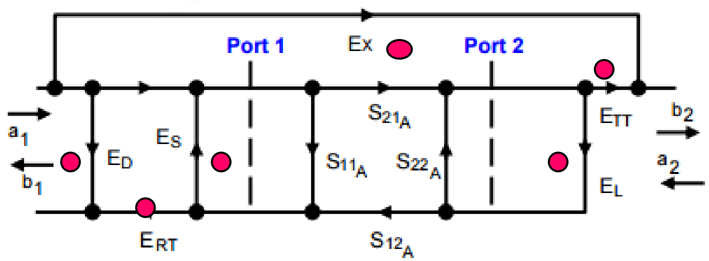
Six forward and six reverse error terms yields 12 error terms for two-port devices

Two-Port Error Correction

Reverse model



Forward model



E_D = fwd directivity
E_S = fwd source match
E_{RT} = fwd reflection tracking
$E_{D'}$ = rev directivity
$E_{S'}$ = rev source match
$E_{RT'}$ = rev reflection tracking
E_L = fwd load match
E_{TT} = fwd transmission tracking
E_X = fwd isolation
$E_{L'}$ = rev load match
$E_{TT'}$ = rev transmission tracking
$E_{X'}$ = rev isolation

- Each actual S-parameter is a function of all four measured S-parameters
- Analyzer must make forward *and* reverse sweep to update any one S-parameter
- Luckily, you don't need to know these equations to ***use*** a network analyzers!!!

$$S_{11a} = \frac{\left(\frac{S_{11m}-E_D}{E_{RT}}\right)\left(1+\frac{S_{22m}-E_D'}{E_{RT}'}E_S'\right) - E_L\left(\frac{S_{21m}-E_X}{E_{TT}}\right)\left(\frac{S_{12m}-E_X'}{E_{TT}'}\right)}{\left(1+\frac{S_{11m}-E_{D'}}{E_{RT}}E_S\right)\left(1+\frac{S_{22m}-E_D'}{E_{RT}'}E_S'\right) - E_L'E_L\left(\frac{S_{21m}-E_X}{E_{TT}}\right)\left(\frac{S_{12m}-E_X'}{E_{TT}'}\right)}$$

$$S_{21a} = \frac{\left(\frac{S_{21m}-E_X}{E_{TT}}\right)\left(1+\frac{S_{22m}-E_D'}{E_{RT}'}(E_S'-E_L)\right)}{\left(1+\frac{S_{11m}-E_D}{E_{RT}}E_S\right)\left(1+\frac{S_{22m}-E_D'}{E_{RT}'}E_S'\right) - E_L'E_L\left(\frac{S_{21m}-E_X}{E_{TT}}\right)\left(\frac{S_{12m}-E_X'}{E_{TT}'}\right)}$$

$$S_{12a} = \frac{\left(\frac{S_{12m}-E_X'}{E_{TT}'}\right)\left(1+\frac{S_{11m}-E_D}{E_{RT}}(E_S-E_L')\right)}{\left(1+\frac{S_{11m}-E_D}{E_{RT}}E_S\right)\left(1+\frac{S_{22m}-E_D'}{E_{RT}'}E_S'\right) - E_L'E_L\left(\frac{S_{21m}-E_X}{E_{TT}}\right)\left(\frac{S_{12m}-E_X'}{E_{TT}'}\right)}$$

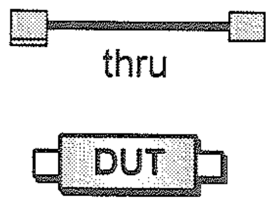
$$S_{22a} = \frac{\left(\frac{S_{22m}-E_D'}{E_{RT}'}\right)\left(1+\frac{S_{11m}-E_D}{E_{RT}}E_S\right) - E_L'\left(\frac{S_{21m}-E_X}{E_{TT}}\right)\left(\frac{S_{12m}-E_X'}{E_{TT}'}\right)}{\left(1+\frac{S_{11m}-E_D}{E_{RT}}E_S\right)\left(1+\frac{S_{22m}-E_D'}{E_{RT}'}E_S'\right) - E_L'E_L\left(\frac{S_{21m}-E_X}{E_{TT}}\right)\left(\frac{S_{12m}-E_X'}{E_{TT}'}\right)}$$

Errors and Calibration Standards

UNCORRECTED

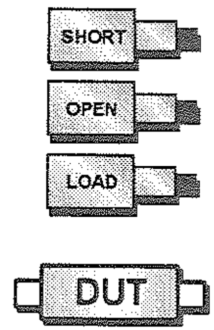
- Convenient
- Generally not accurate
- No errors removed

RESPONSE



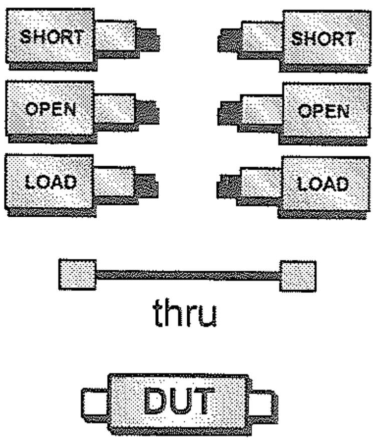
- Easy to perform
- Use when highest accuracy is not required
- Removes frequency response error

1-PORT



- For reflection measurements
- Need good termination for high accuracy with two-port devices
- Removes these errors:
  Directivity
  Source match
  Reflection tracking

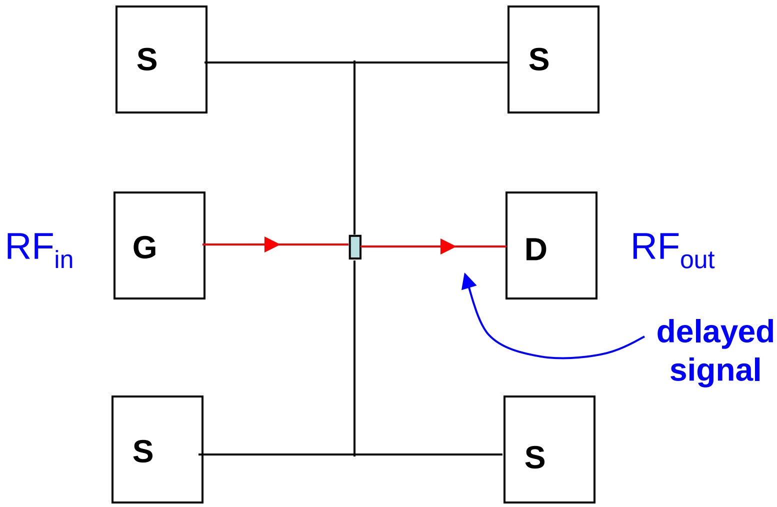
FULL 2-PORT

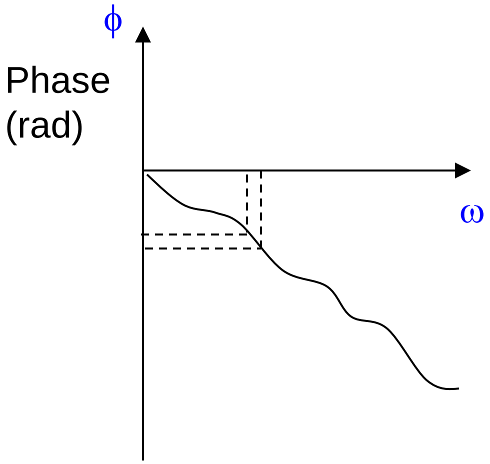


- Highest accuracy
- Removes these errors:
  Directivity
  Source, load match
  Reflection tracking
  Transmission tracking
  Crosstalk

Other errors:
Random Errors: Noise, Repeatability
Drift

THRU CALIBRATION FOR GROUP DELAY





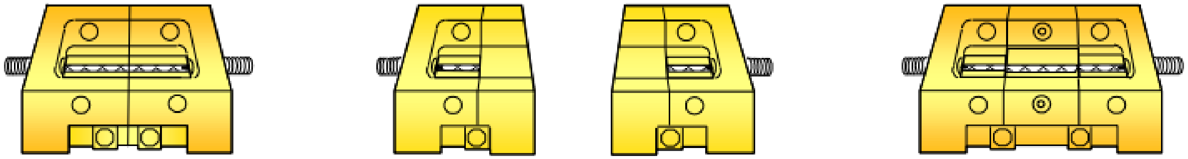
$$\text{Group delay } t_g = -\frac{\partial \phi}{\partial \omega} = -\frac{1}{360^{o}}\frac{\partial \theta}{\partial f}$$

Thru-Reflect-Line (TRL) Calibration

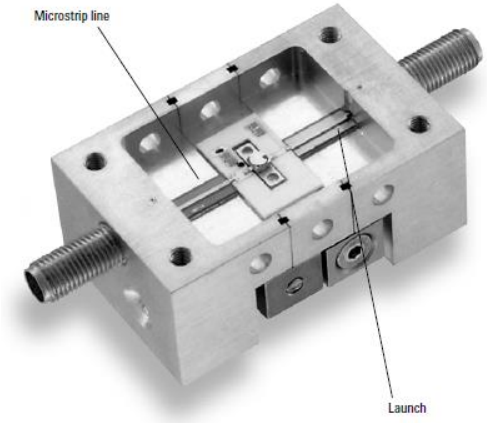
We know about Short-Open-Load-Thru (SOLT) calibration... What is TRL?

- A two-port calibration technique
- Good for non-coaxial environments (waveguide, fixtures, wafer probing)
- Characterizes same 12 systematic errors as the more common SOLT cal
- Uses practical calibration standards that are easily fabricated and characterized
- Other variations: Line-Reflect-Match (LRM), Thru-Reflect-Match (TRM), plus many others

TRL was developed for **non-coaxial microwave** measurements



TRL CALIBRATION



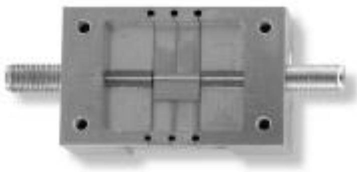
At low frequencies, the lines become long. So, we need different TRL structures for different frequency bands for wide band characterization. Use SOLT for lower freq.

THRU

$$\begin{bmatrix} 0 & 1 \\ 1 & 0 \end{bmatrix}$$

REFLECT

$$\begin{bmatrix} \Gamma & 0 \\ 0 & \Gamma \end{bmatrix}$$



LINE

$$\begin{bmatrix} 0 & e^{-\gamma l} \\ e^{-\gamma l} & 0 \end{bmatrix}$$

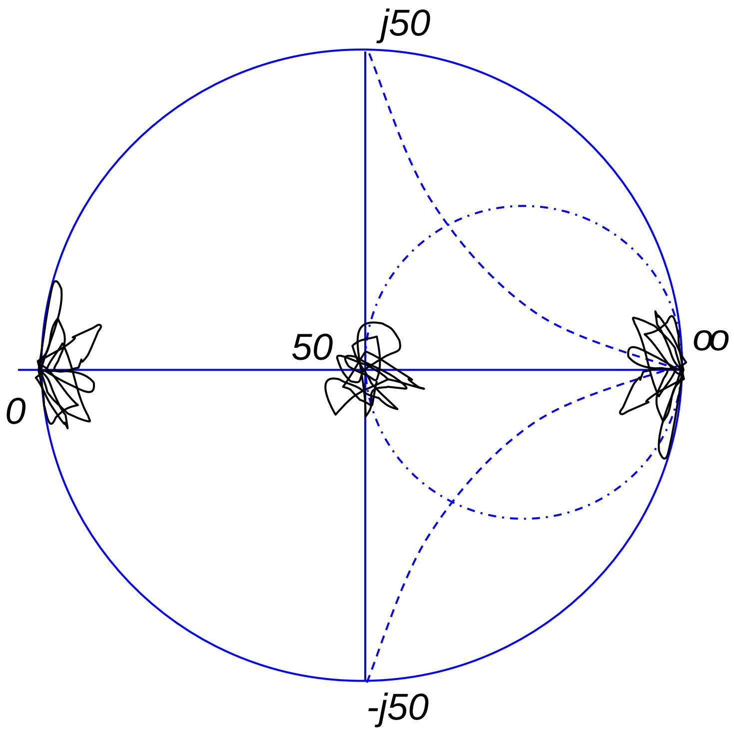
Glenn Engen, Cletus Hoer, MTT-27 (12), Dec 1979

RF MICROWAVE PROBES (CASCADE MICROTECH)

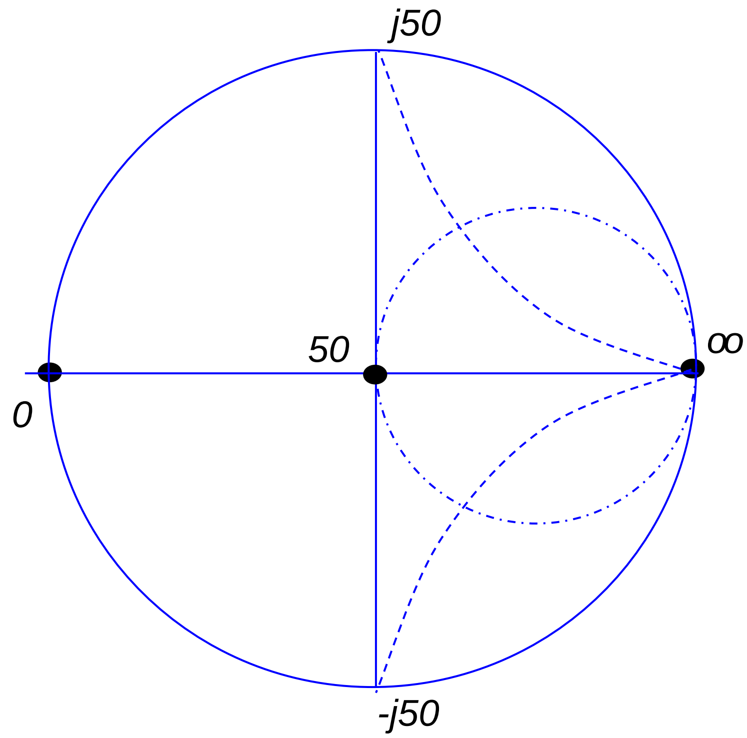
CALIBRATION USING IMPEDANCE STANDARD SUBSTRATES

The purpose of Cal is to bring the reference plane to the probe tips

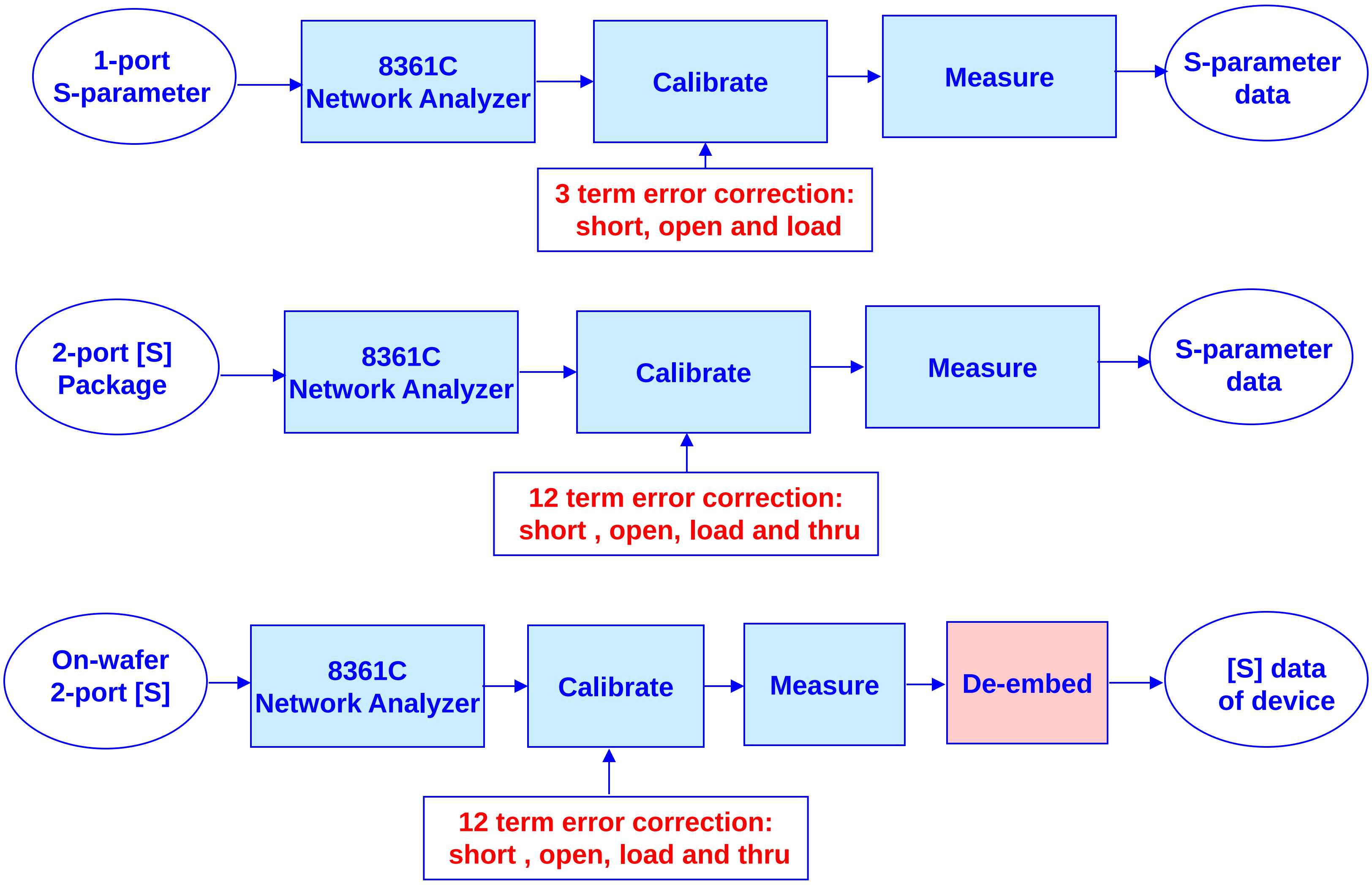
BEFORE CALIBRATION

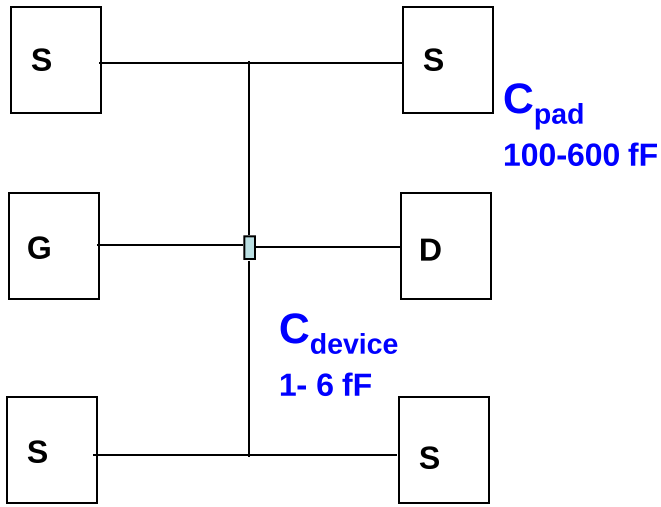


AFTER CALIBRATION

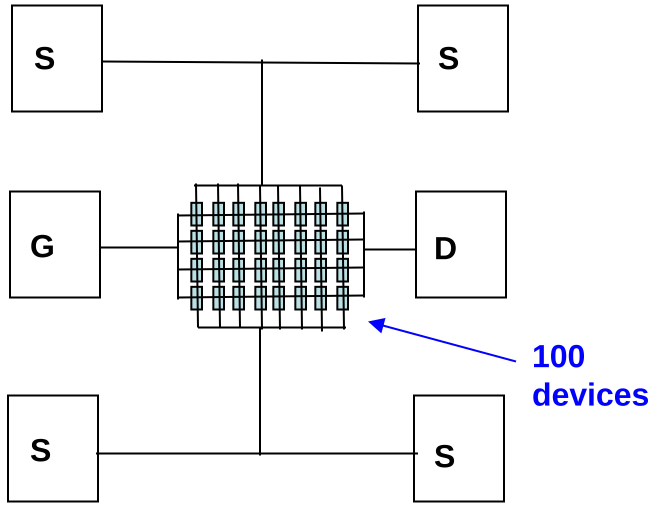


DE- EMBEDDING


1-port
S-parameter
8361C
Network Analyzer
Calibrate
Measure
S-parameter
data
3 term error correction:
short, open and load
2-port [S]
Package
8361C
Network Analyzer
Calibrate
Measure
S-parameter
data
12 term error correction:
short , open, load and thru
On-wafer
2-port [S]
8361C
Network Analyzer
Calibrate
Measure
De-embed
[S] data
of device
12 term error correction:
short , open, load and thru

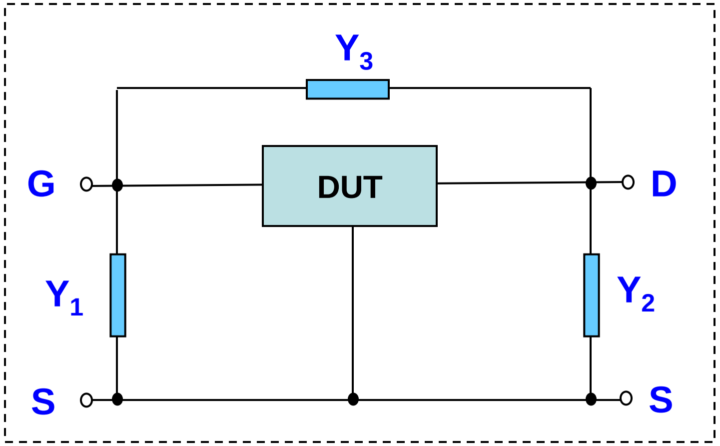


Pad capacitance far exceeds
Single device capacitance



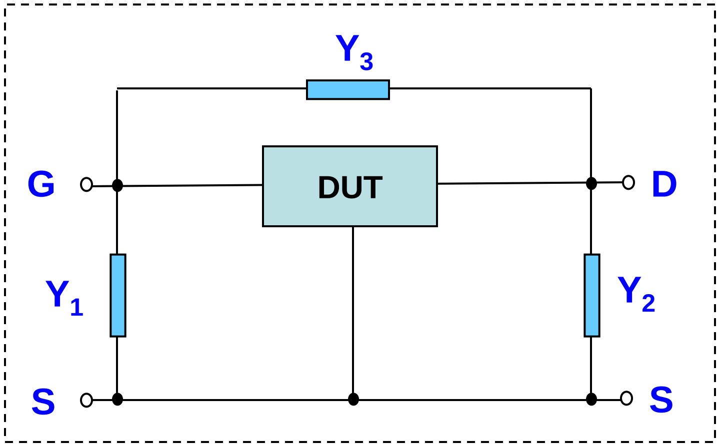
A device array alleviates
this problem to some extent

Device

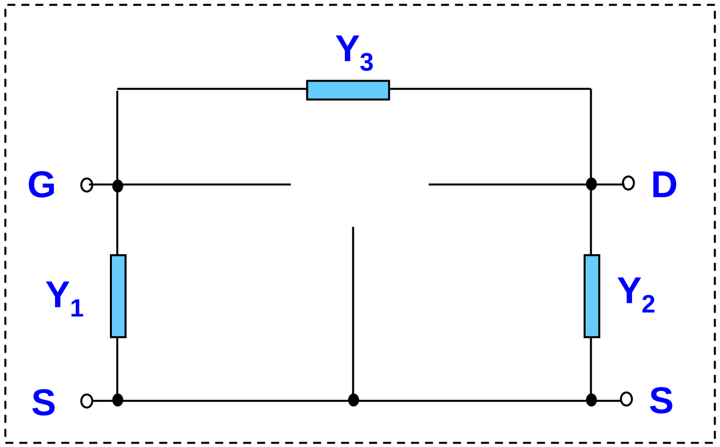


1-Step De-Embedding

Device



OPEN



$$[S]_{open} \rightarrow [y]_{open}$$

$$[S]_{device} \rightarrow [y]_{device}$$

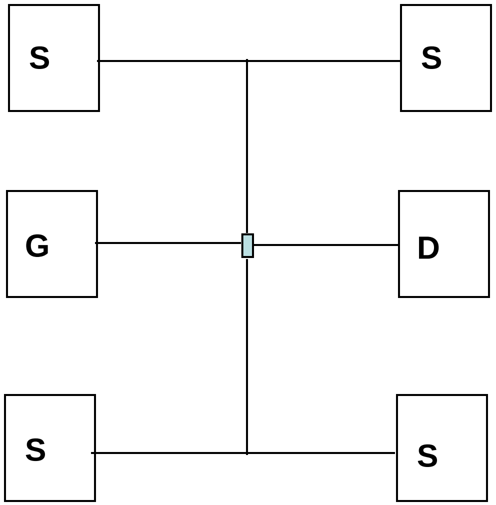
$$[y]_{DUT} = [y]_{device} - [y]_{open}$$

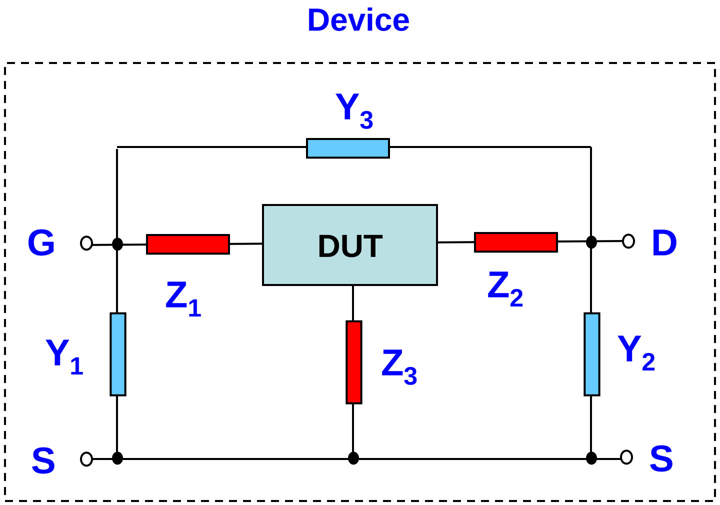
$$[y]_{DUT} \rightarrow [S]_{DUT}$$

Measure on- wafer OPEN
Measure DEVICE
Use the equations on the left

DE- EMBEDS PAD CAPACITANCE ONLY !

2- Step De - Embedding



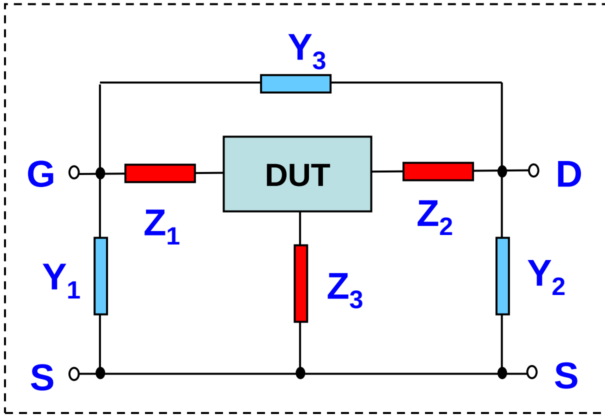


DE- EMBEDS

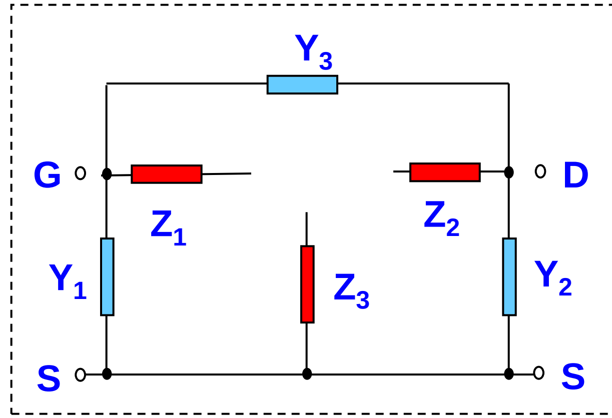
- **Pad Capacitance**
- **Series Impedance**

2- Step De - Embedding

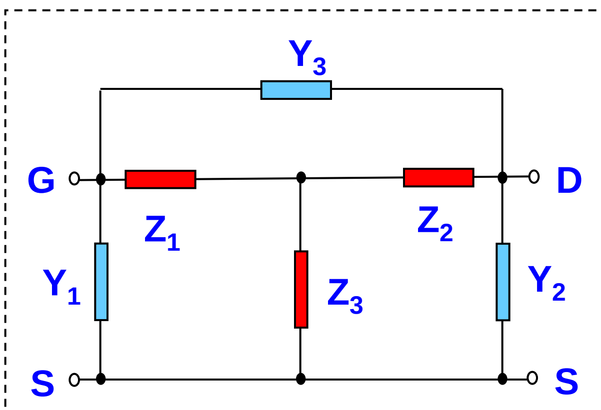
Device



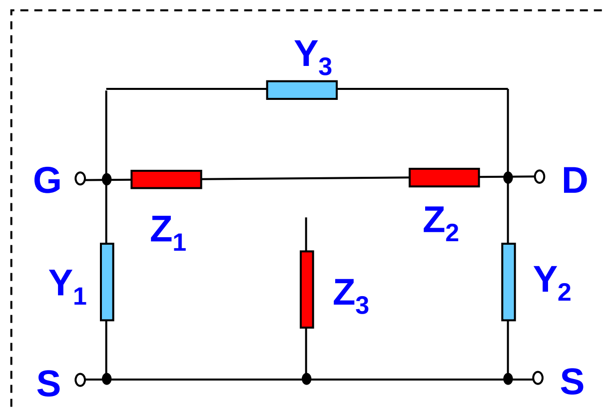
OPEN



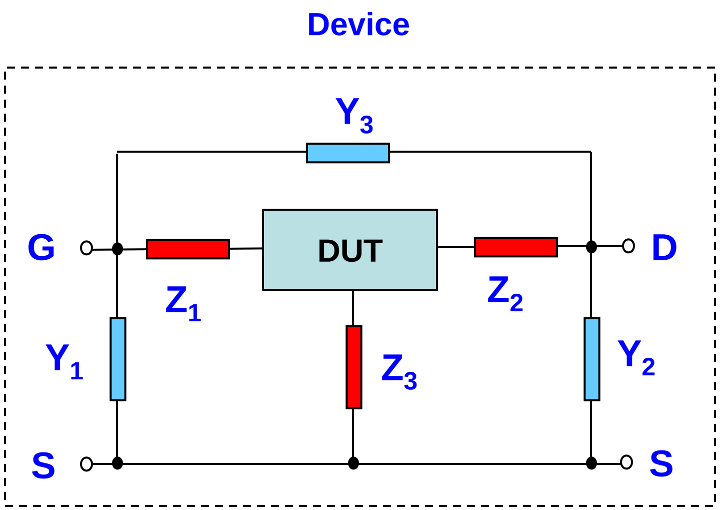
SHORT



THRU



2- Step De - Embedding



DUT with Pad capacitance and series elements

$$[S]_{open} \rightarrow [y]_{open}$$

$$[S]_{short} \rightarrow [z]_{short}$$

$$[S]_{device} \rightarrow [y]_{device}$$

$$[y]_{dev\_no\_pad} = [y]_{device} - [y]_{open}$$

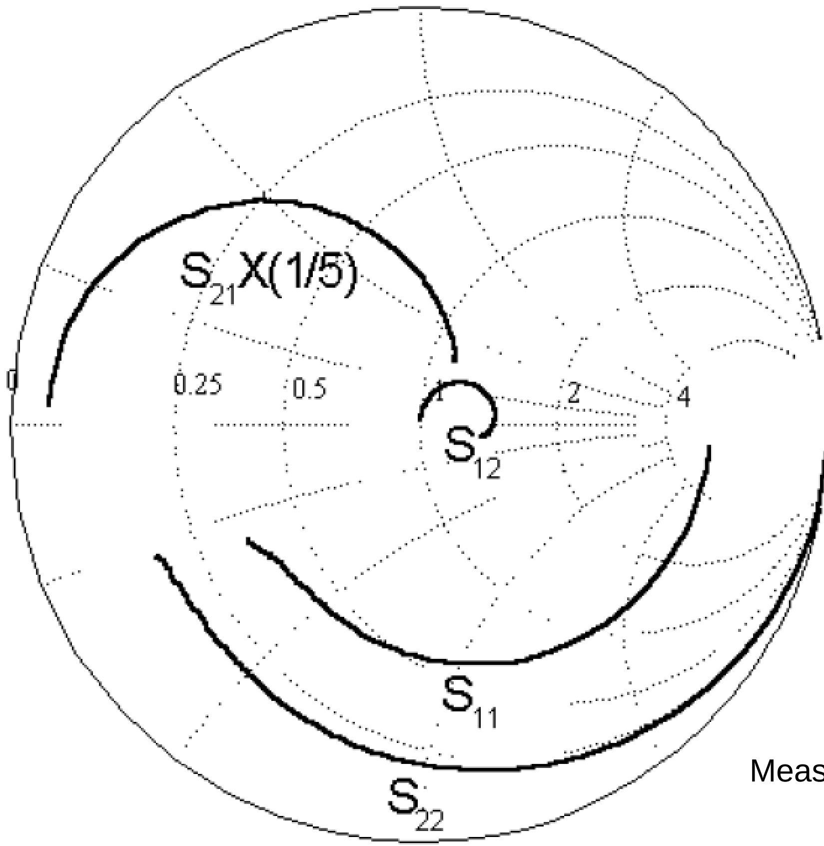
$$[y]_{dev\_no\_pad} \rightarrow [z]_{dev\_no\_pad}$$

$$[z]_{DUT} = [z]_{dev\_no\_pad} - [z]_{short}$$

$$[z]_{DUT} \rightarrow [S]_{DUT}$$

Whoa!

High Frequency Performance of 0.18um CMOS

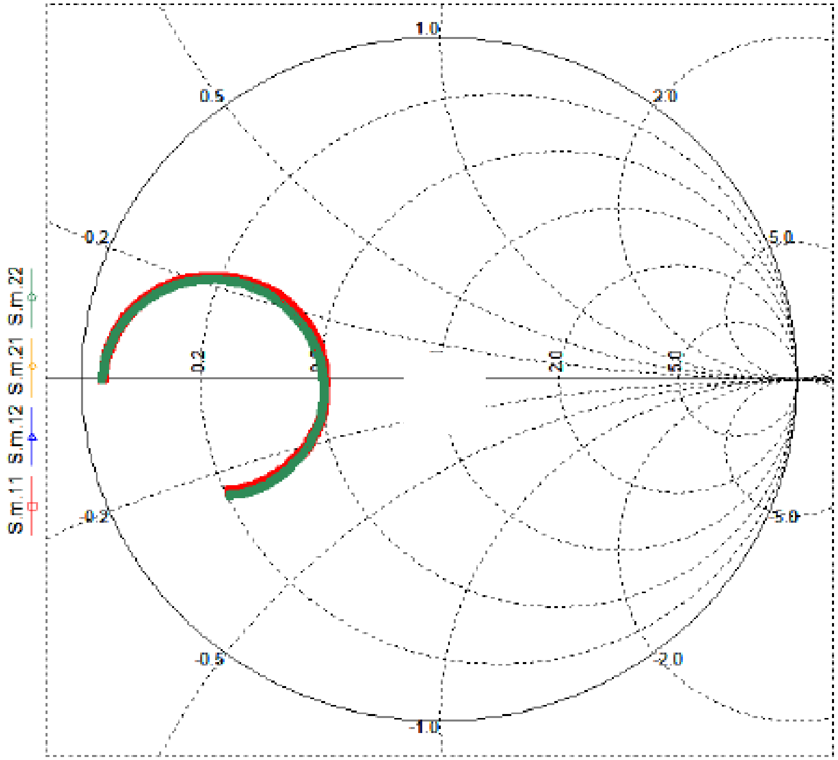


Ft=50GHz, Fmax=45GHz

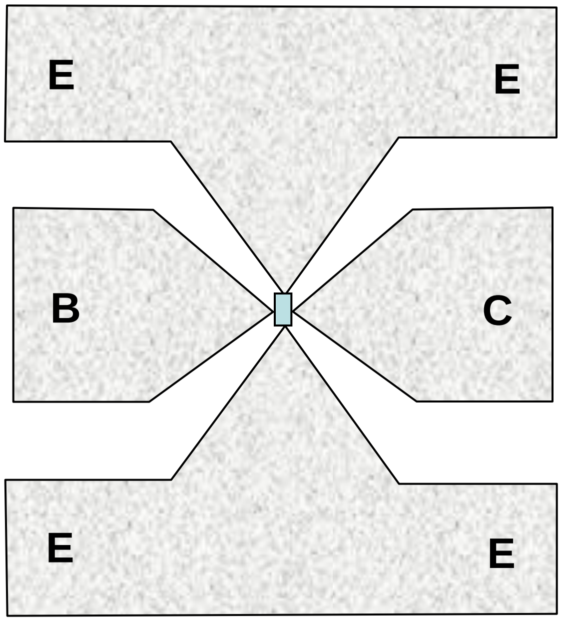
Measurement: 0.5-50GHz

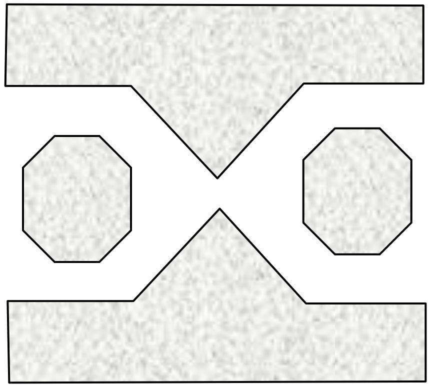
RF Characteristics of 0.18-m CMOS Transistors: Kwangseok Han, Jeong-hu Han, Minkyu Je and Hyungcheol Shin Department of Electrical Engineering and Computer Science, Korea Advanced Institute of Science and Technology, Taejon 305-701

IMPERFECT on-wafer SHORT



Layout with minimum inductance and reflections





Small octagonal pads
to reduce capacitance
and reflections

Definitions of Gain and Stability

DEFINITIONS OF GAIN

Any three terminal device will have A.C current gain. We can then define a transition frequency f_T at which the short circuit current gain becomes unity. This is one figure of merit.

$$h_{21} = \left.\frac{I_2}{I_1}\right|_{V_2=0} \qquad h_{21} = 1 \quad \text{at} \quad f = f_T$$

Any three terminal active device will have power gain $G_p = \frac{P_L}{P_{IN}}$

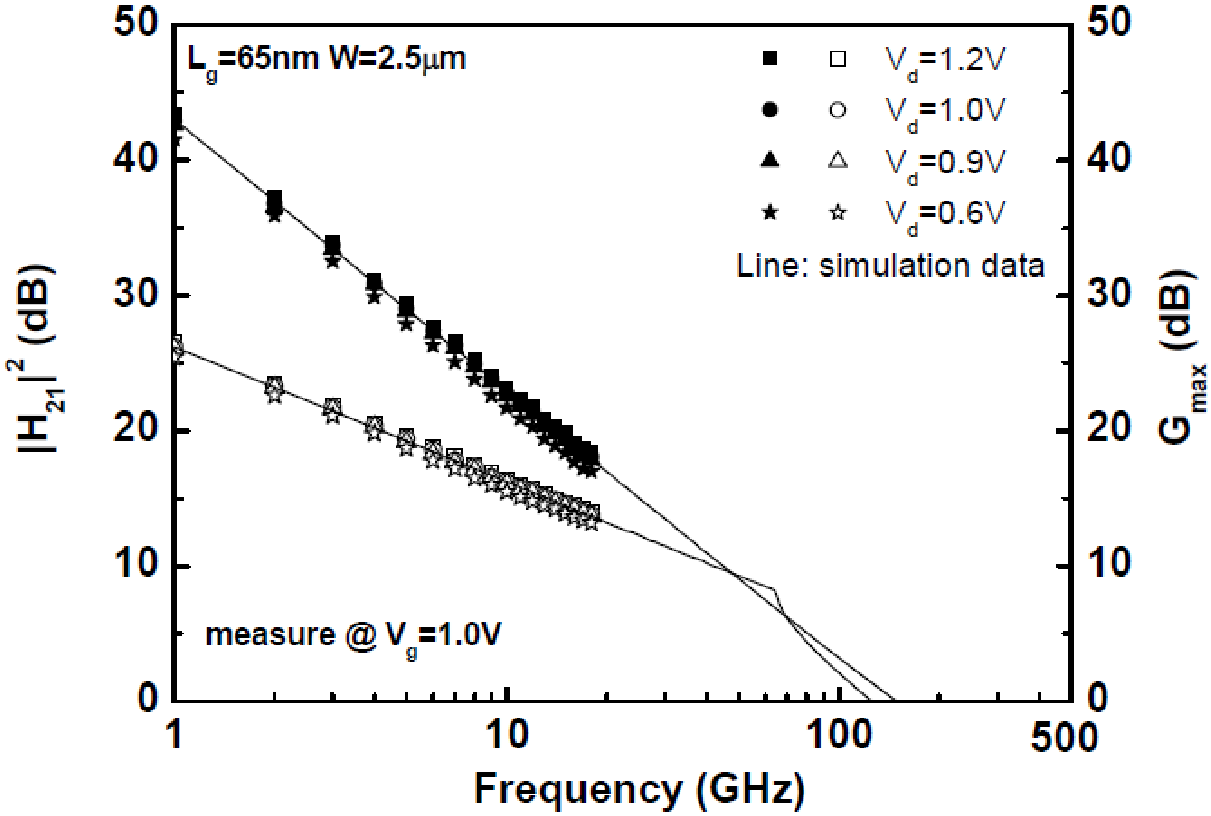
Since P_L depends on load impedance and P_{IN} depends on source impedance it is meaningful to define Available Power Gain as:

$$G_A = \frac{P_{AVN}}{P_{AVS}} = \frac{\text{Power available from the network}}{\text{Power available from the source}}$$

The frequency at which the Maximum Available Gain (MAG) becomes unity is defined as f_{max} . This is another figure of merit.

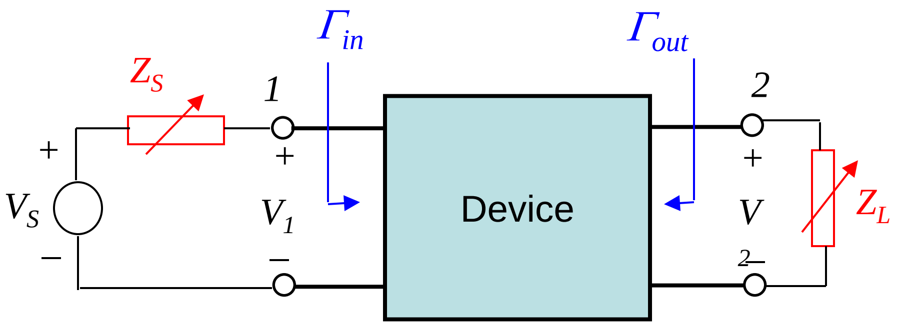
$$MAG = 1 \quad \text{at} \quad f = f_{\max}$$

Ft and Fmax of 65nm CMOS



Limiting Factors of RF Performance Improvement as Down-scaling to 65-nm Node MOSFETs
H. L. Kaoa*, B. S. Lina, C. C. Liaob, M. H. Chenc, C. H. Wuc, and Albert Chinb
a Dept. of Electronic Engineering, Chang Gung Univ., Tao-Yuan, Taiwan, ROC
b Nano-Sci. Tech. Ctr, EE. Dept., Nat'l Chiao-Tung Univ., UST, Hsinchu, Taiwan, ROC
c Dept. of MicroElectronics Engineering, Chung Hua Univ., Hsinchu, Taiwan, ROC

CIRCUIT FOR MEASURING MAXIMUM AVAILABLE GAIN & Fmax



We vary Z_S and Z_L so as to provide a simultaneous conjugate match.
This maximizes the input power and delivers maximum output power to the load.
This will give us MAG.
If we do this at each frequency, we can generate a plot of MAG vs frequency.
From this plot, we can determine the frequency at which the power gain will become unity. This is Fmax.

THE EFFECT OF LOAD IMPEDANCE

$$S_{11} = 0.65\angle -95^\circ$$

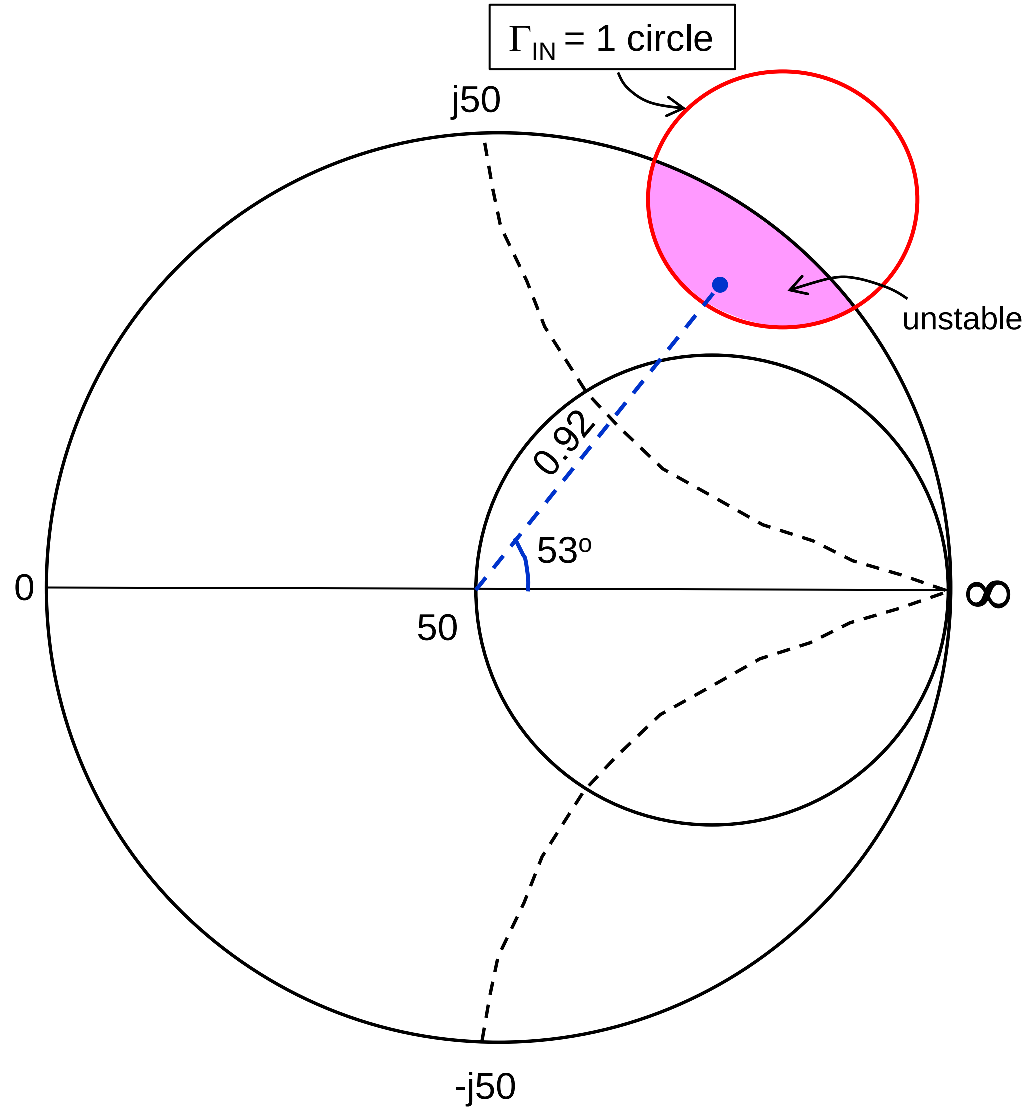
$$S_{12} = 0.04\angle 40^\circ$$

$$S_{21} = 5.00\angle 115^\circ$$

$$S_{22} = 0.80\angle -35^\circ$$

$$\Gamma_{in} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L}$$

Z_L (ohms)	Γ_L	Γ_{IN}	Stable ?
50+j 0	0	$0.65\angle -95^\circ$	Yes
35+j 100	$0.77\angle 49^\circ$	$0.91\angle -108^\circ$	Yes
10+j 100	$0.92\angle 53^\circ$	$1.07\angle -103^\circ$	Unstable



THE EFFECT OF SOURCE IMPEDANCE

$S_{11} = 0.65\angle -95^{\circ}$

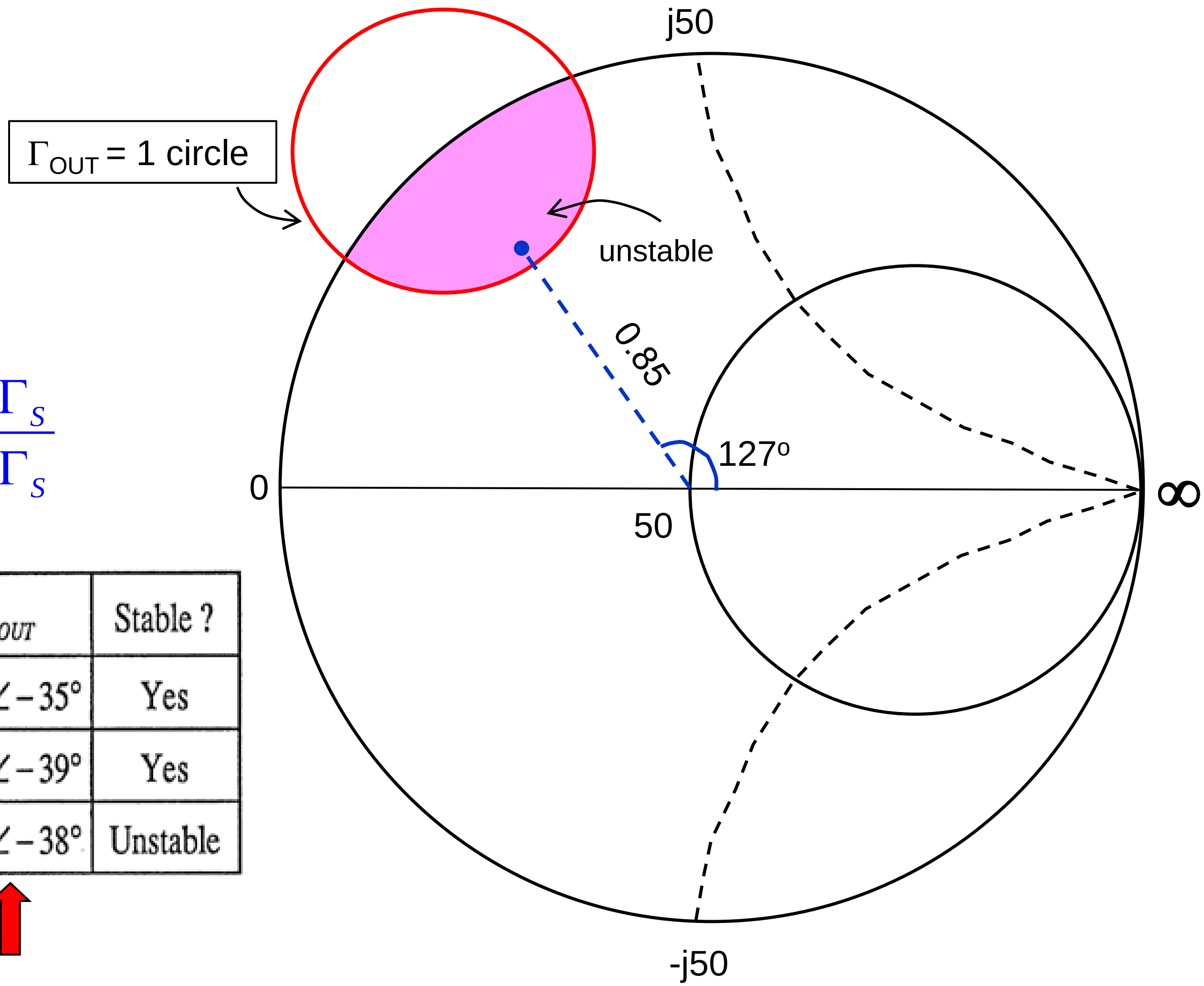
$S_{12} = 0.04\angle 40^{\circ}$

$S_{21} = 5.00\angle 115^{\circ}$

$S_{22} = 0.80\angle -35^{\circ}$

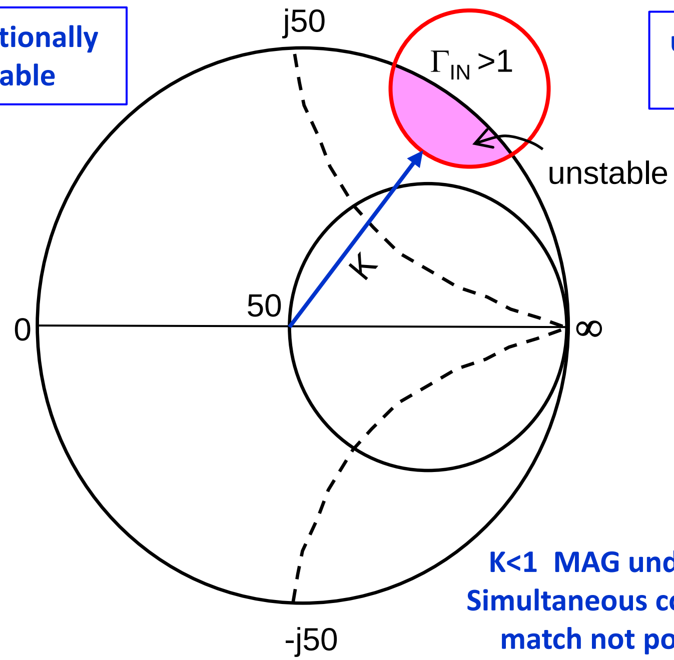
$$\Gamma_{out} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S}$$

Z_S (ohms)	Γ_S	Γ_{OUT}	Stable ?
$50+j\,0$	0	$0.80\angle -35^{\circ}$	Yes
$10+j\,25$	$0.73\angle 125^{\circ}$	$0.99\angle -39^{\circ}$	Yes
$5+j\,25$	$0.85\angle 127^{\circ}$	$1.04\angle -38^{\circ}$	Unstable



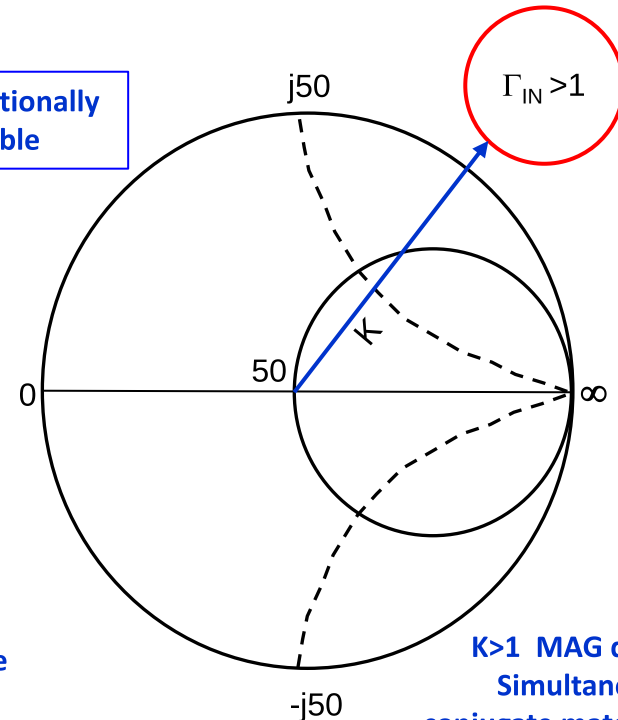
OUTPUT STABILITY CIRCLE & UNCONDITIONAL STABILITY

conditionally stable



K<1 MAG undefined
Simultaneous conjugate match not possible

unconditionally stable



K>1 MAG defined
Simultaneous conjugate match possible

$$K = \frac{1 + |S_{11}S_{22} - S_{12}S_{21}|^2 - |S_{11}|^2 - |S_{22}|^2}{2|S_{12}||S_{21}|}$$ **= Stability factor**

$$\Delta = |S_{11}S_{22} - S_{12}S_{21}|$$

For unconditional stability

$$K > 1 \; and \; \Delta < 1$$

J.M.Rolett, IRE Trans CT, CT-9(1), pp 29-32, Mar 1962,   W.Ku, Proc IEEE 54(11), pp 1617-1618, Nov 1966

GAIN EQUATIONS in S- Domain

Maximum Available Gain

$$\mathrm{MAG} = \frac{|S_{21}|}{|S_{12}|}\left(K - \sqrt{K^2 - 1}\right)$$

Use only for K>1

Maximum Stable Gain

$$\mathrm{MSG} = \frac{|S_{21}|}{|S_{12}|}$$

Use only for K<1

Mason's Gain

$$U = \frac{|S_{21}/S_{12} - 1|^2}{2K\,|S_{21}/S_{12}| - 2\mathrm{Re}(S_{21}/S_{12})}$$

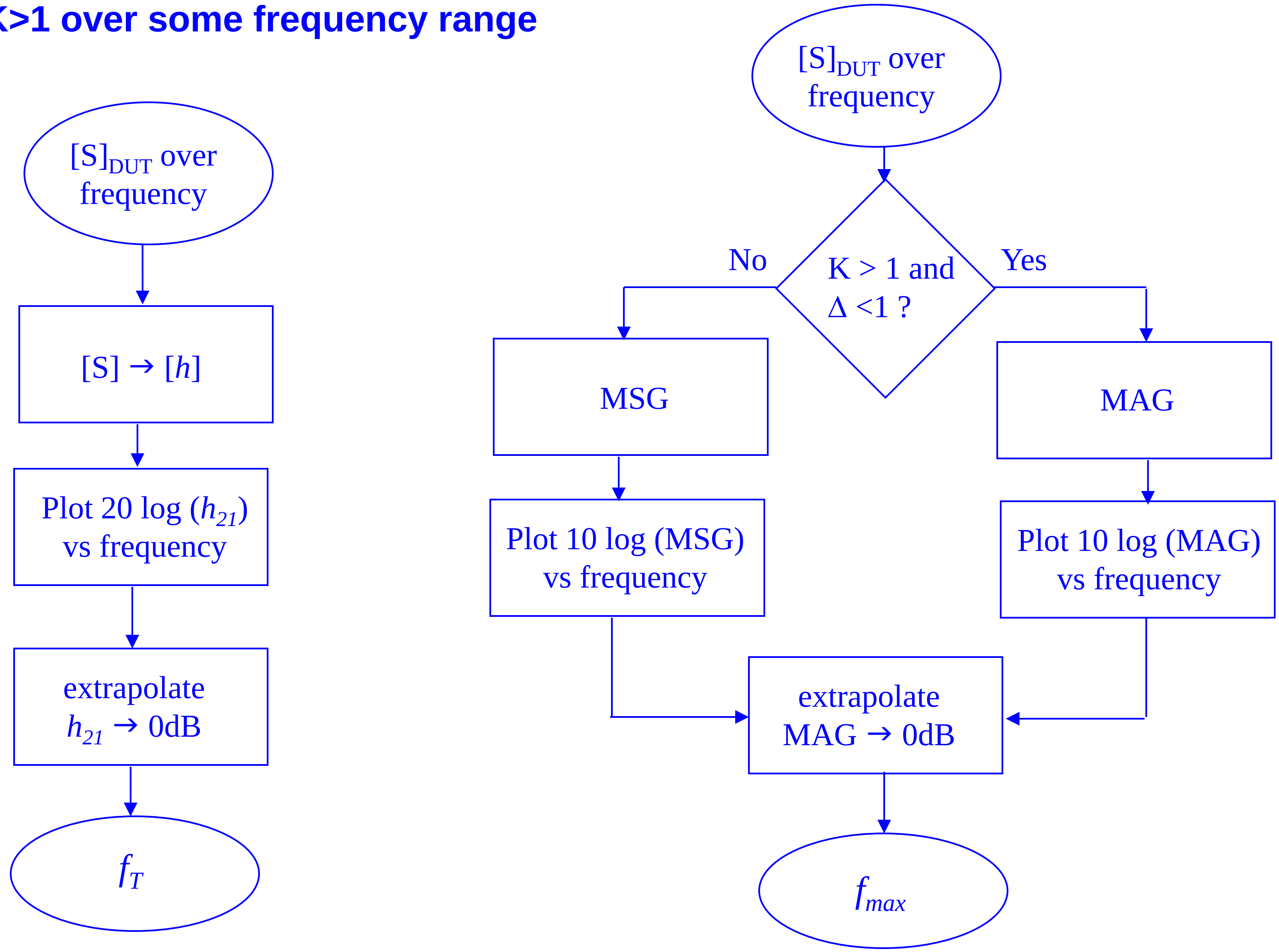
Use for all K
Does not depend on K

Current Gain

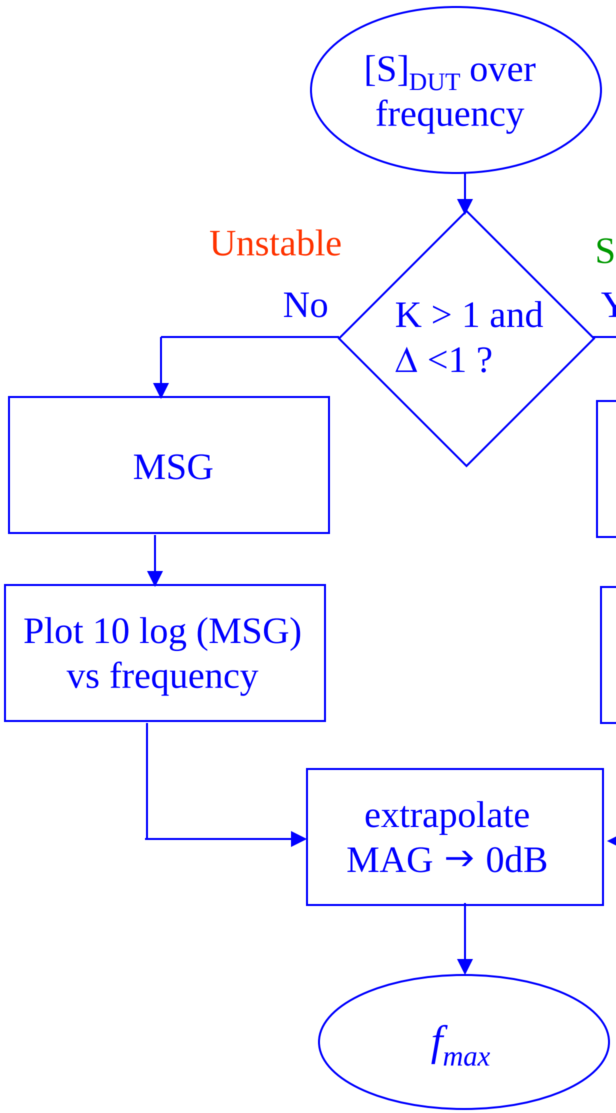
$$h_{21} = \frac{-2S_{21}}{(1 - S_{11})(1 + S_{22}) + S_{12}S_{21}}$$

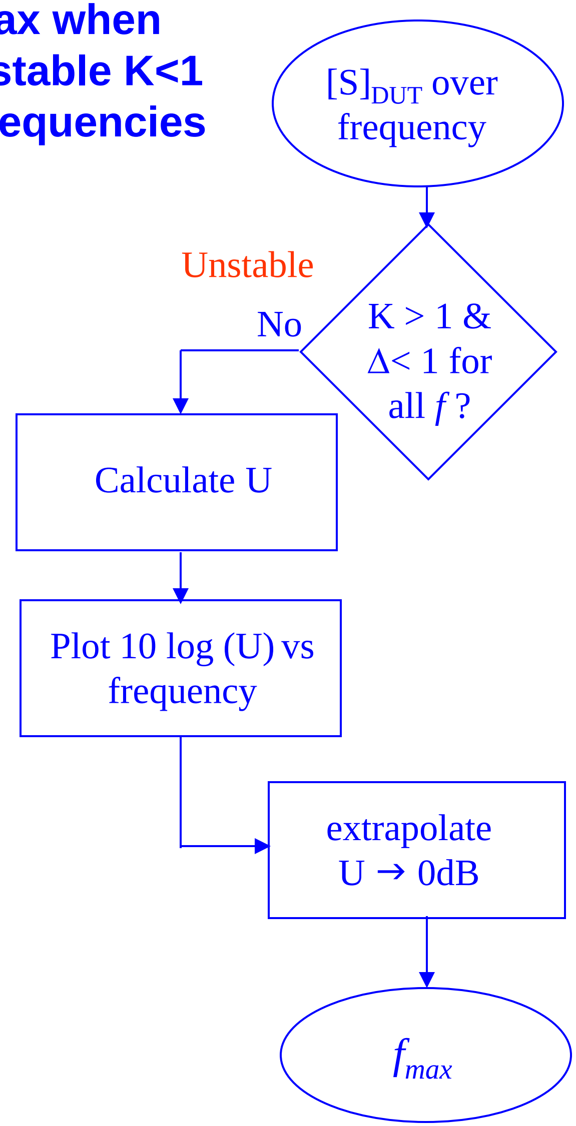
Determining Ft and Fmax from Measured S-parameter Data

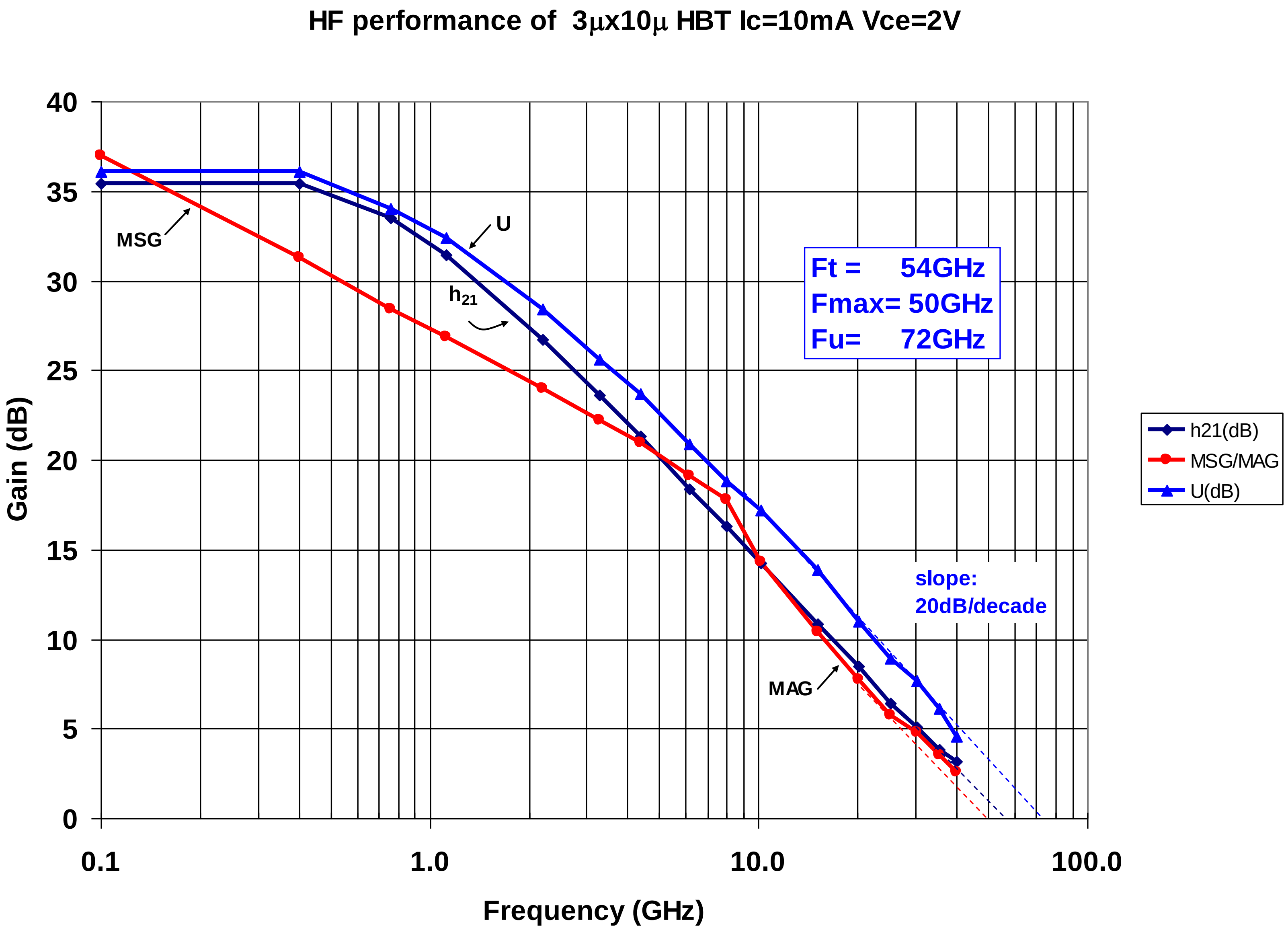
Determining Fmax when the device is stable K>1 over some frequency range



Determining Fmax when the device is unstable K<1 throughout all frequencies

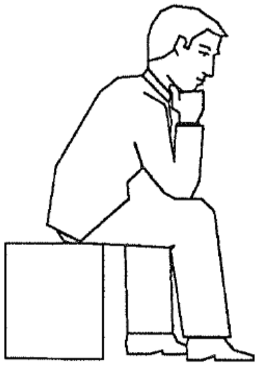





HF performance of 3μx10μ HBT Ic=10mA Vce=2V
40
35
30
25
20
15
10
5
0
Gain (dB)
0.1
1.0
10.0
100.0
Frequency (GHz)
MSG
U
h21
MAG
Ft = 54GHz
Fmax= 50GHz
Fu= 72GHz
slope:
20dB/decade
h21(dB)
MSG/MAG
U(dB)

Mason’s Gain Explained

If Fmax is a Figure of Merit, is it Unique?



MAG will be different in different configurations (CE,CB,CC).
Then, when I extropolate MAG will I get different Fmax
for different configurations ?

If the device is unstable throughout the band of frequencies,
should I extrapolate MSG to find Fmax ?

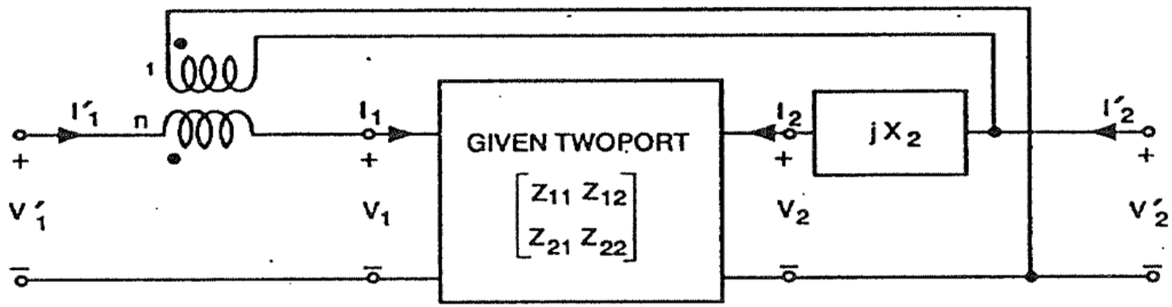
The device becomes unstable because of S12.
Is there a way to make S12=0 so that I can always
compute stable gain and find Fmax ?

Is there a way gain can be defined so that it is
independent of the configuration ?

Can the new definition of gain be consistent
with the way MAG is extrapolated ?

How can I find out whether a device is Active or Passive?

Method of unilateralizing a Two Port Network



$$\begin{bmatrix} Z'_{11} & 0 \\ Z'_{21} & Z'_{22} \end{bmatrix}$$

Add a reactance jX2 at the output such that the voltage V2 is exactly out of phase with V1 when I1=0. Choose a turns ratio such that the voltage introduced cancels out the reverse voltage transfer through the device.

POWER GAIN CALCULATIONS IN DIFFERENT CONFIGURATIONS

Parameter	CE	CC	CB
S11	$0.33\angle -111°$	$0.91\angle -29.03°$	$0.71\angle 159.32°$
S12	$0.09\angle 52°$	$0.27\angle 59.44°$	$0.04\angle 84.04°$
S21	$3.79\angle 96°$	$1.58\angle -22.71°$	$1.65\angle -25.32°$
S22	$0.51\angle -29°$	$0.67\angle 143.02°$	$0.99\angle -15.64°$
K	1.09	0.19	0.03
h_{fe} dB	14.38	14.31	-0.11
MAG dB	14.47	***	***
MSG dB	16.24	7.59	16.00
U dB	17.45	17.45	17.45

MASON'S INVARIANT or UNILATERAL GAIN U

$$U = \frac{|y_{12} - y_{21}|^2}{4[\mathrm{Re}(y_{11})\mathrm{Re}(y_{22}) - \mathrm{Re}(y_{12})\mathrm{Re}(y_{21})]}$$

$$U = \frac{|z_{12} - z_{21}|^2}{4[\mathrm{Re}(z_{11})\mathrm{Re}(z_{22}) - \mathrm{Re}(z_{12})\mathrm{Re}(z_{21})]}$$

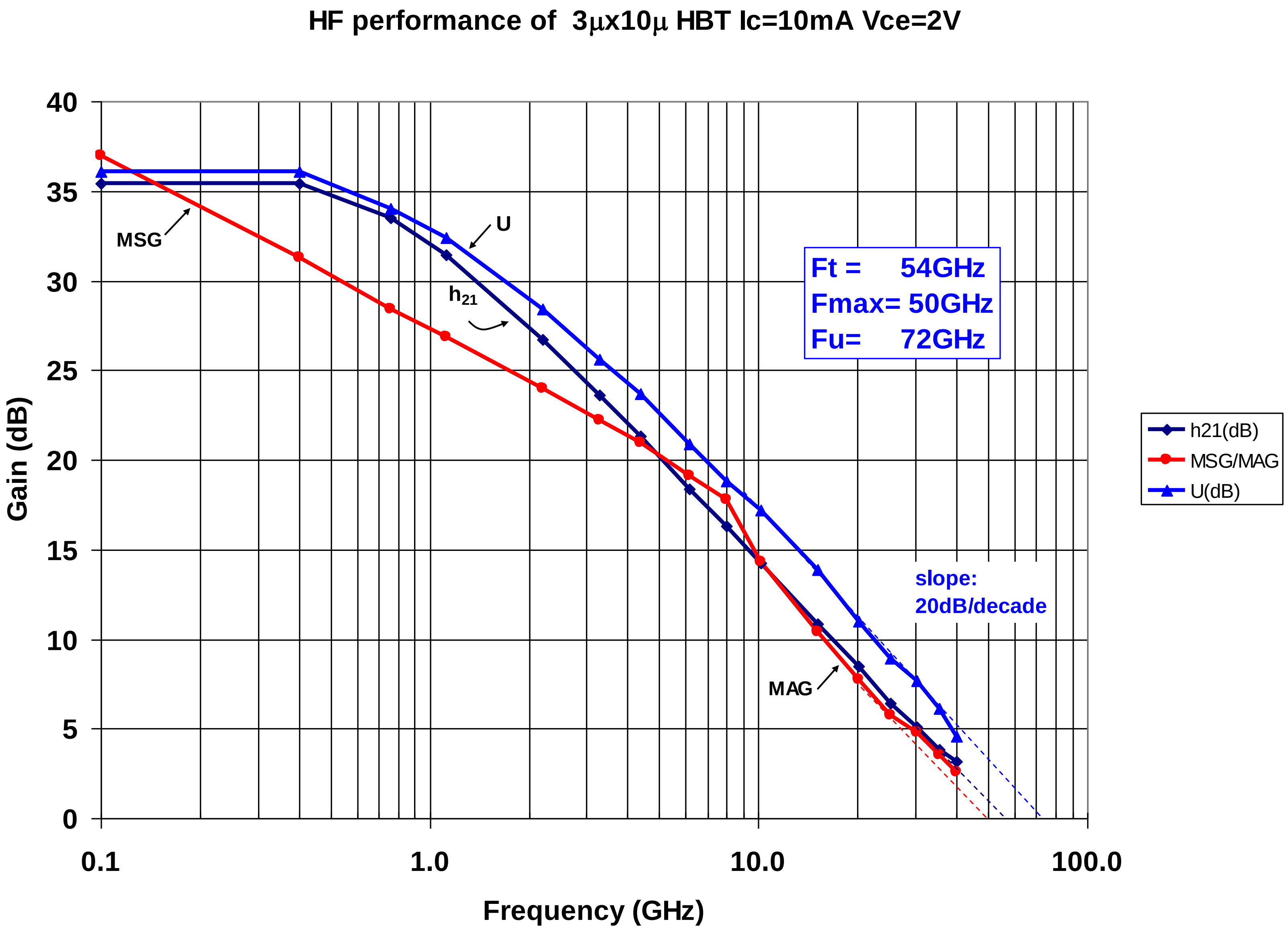
$$U = \frac{|S_{12} - S_{21}|^2}{\det[\mathbf{1} - \mathbf{SS^*}]}$$

If $U > 1$, DEVICE IS ACTIVE

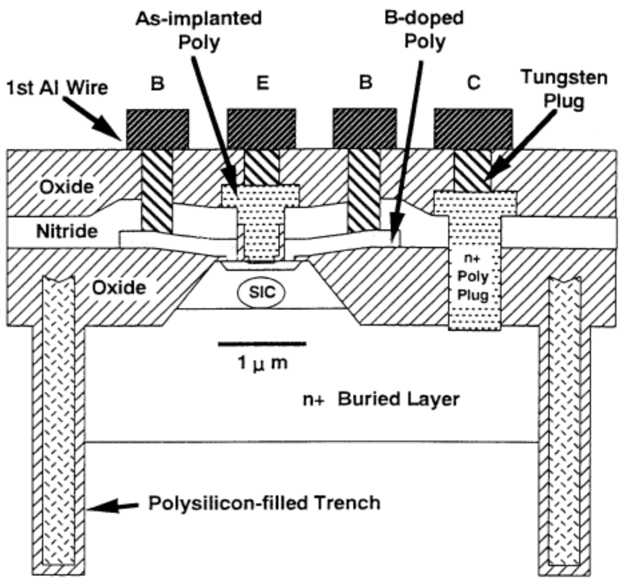
If $U < 1$, DEVICE IS PASSIVE

S.J.Mason, IRE Tran CT, CT-1 (2), pp 20-25, June 1954
M.S.Gupta, T-MTT, 40(5), pp 864-879, 1992

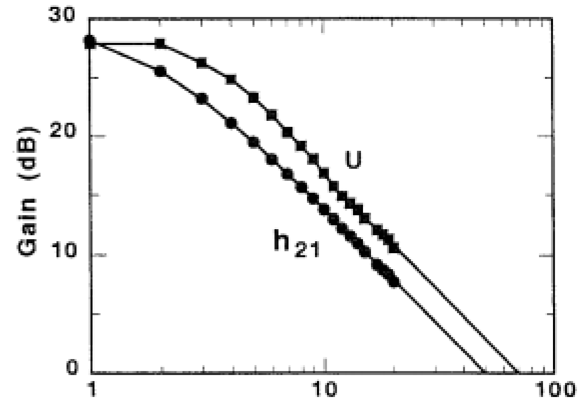
Transistor Specmanship!


HF performance of 3μx10μ HBT Ic=10mA Vce=2V
Gain (dB)
40
35
30
25
20
15
10
5
0
0.1
1.0
10.0
100.0
Frequency (GHz)
MSG
U
h21
MAG
Ft = 54GHz
Fmax= 50GHz
Fu= 72GHz
slope:
20dB/decade
h21(dB)
MSG/MAG
U(dB)

70GHz Fmax Silicon Bipolar Transistor



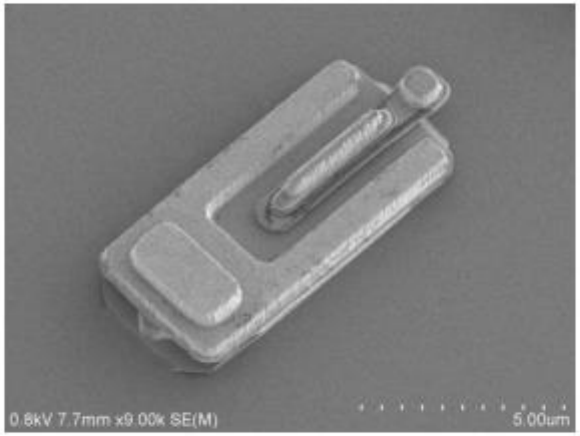
SIC = Selectively Ion-implanted Collector



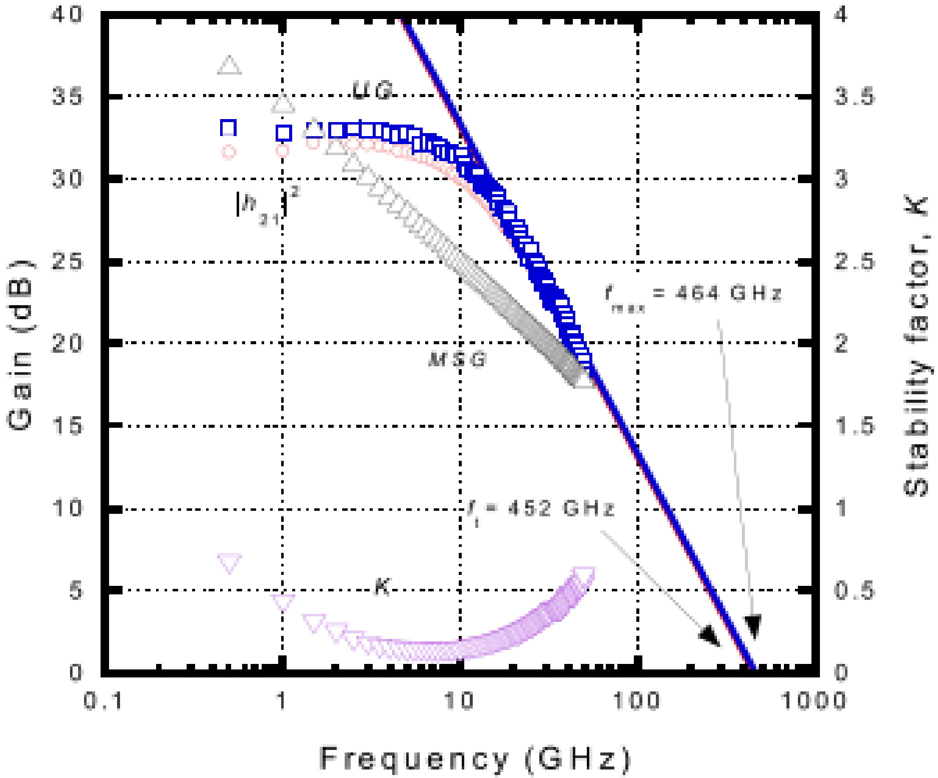
Emitter Area (μm^2)	1.5×0.3	0.6×0.3
β_{MAX}	60	60
R_B (Ω)	156	230
R_E (Ω)	60	120
R_C (Ω)	167	336
BV_{CEO} (V)	3	3
BV_{CBO} (V)	10	10
BV_{EBO} (V)	4.5	4.5
Early Voltage (V)	9	9
C_{JC} (f F)	2.7	1.8
C_{JE} (f F)	2.8	1.3
peak f_T (GHz)		
@ V_{CE} = 1.0 V	44	42
@ V_{CE} = 2.5 V	50	47
peak f_{MAX} (GHz)		
@ V_{CE} = 1.0 V	55	53
@ V_{CE} = 2.5 V	70	65

Mamoru Ugajin, Jun-ichi Kodate, Yoshiji Kobay ashi, Shinsuke Konaka, and Tetsushi Sakai NTT LSI Laboratories 3-1, Morinosato Wakamiya, Atsugi-Shi, Kanagawa, 243-01 Japan
IEDM95

450GHz Ft and Fmax InP/InGaAs HBT



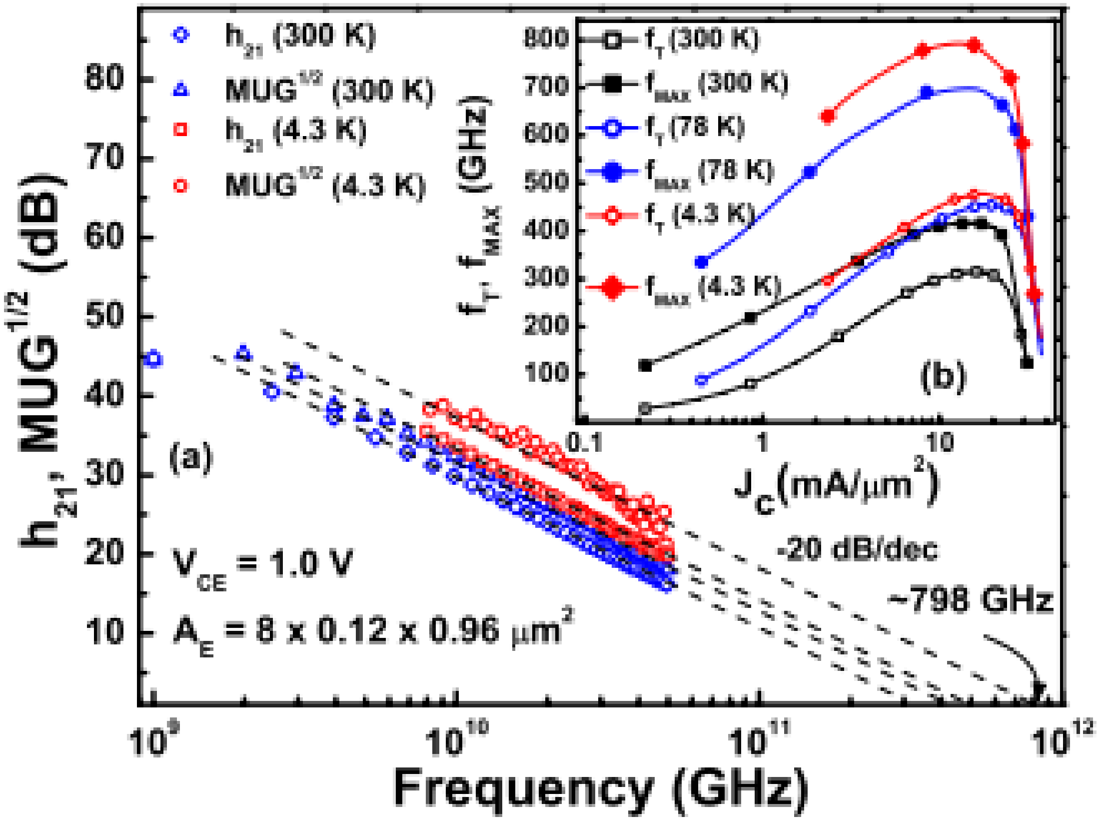
3. SEM image of a 0.25-μm emitter InP/InGaAs DHBT.



N. Kashio et al., TED 61(10), Oct 2014 p 3423

High Frequency Performance of 0.12um SiGe HBT

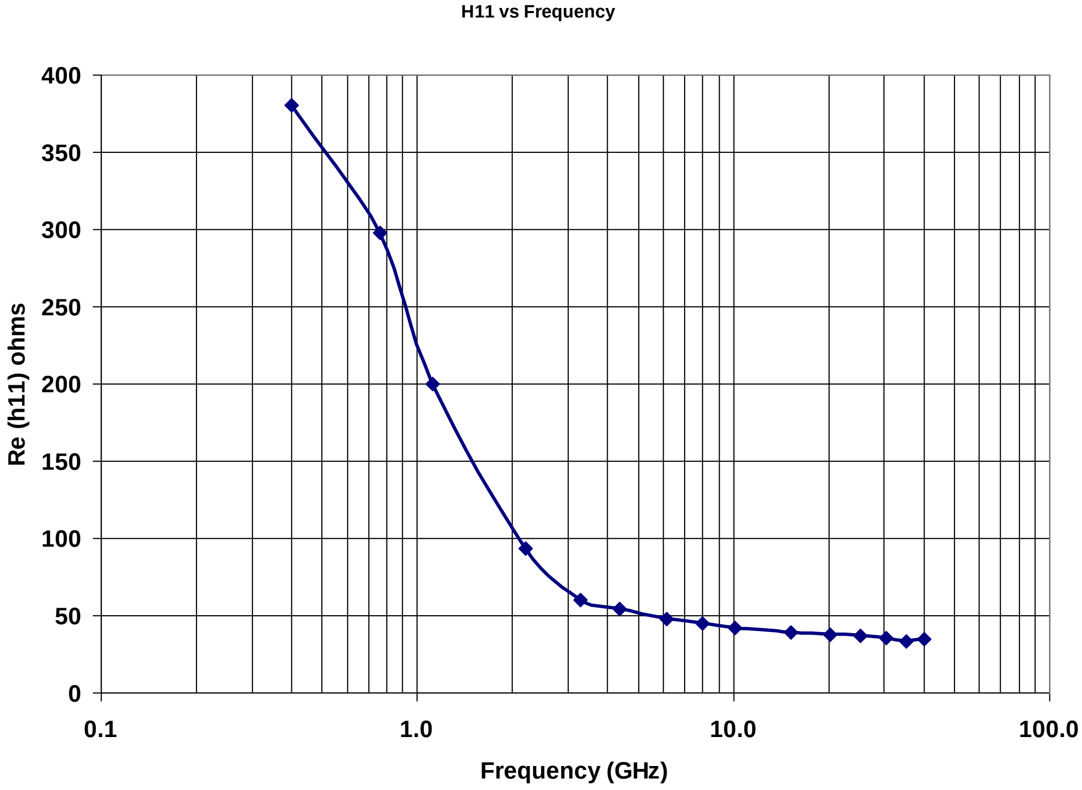
IEEE ELECTRON DEVICE LETTERS, VOL. 35, NO. 2, FEBRUARY 2014

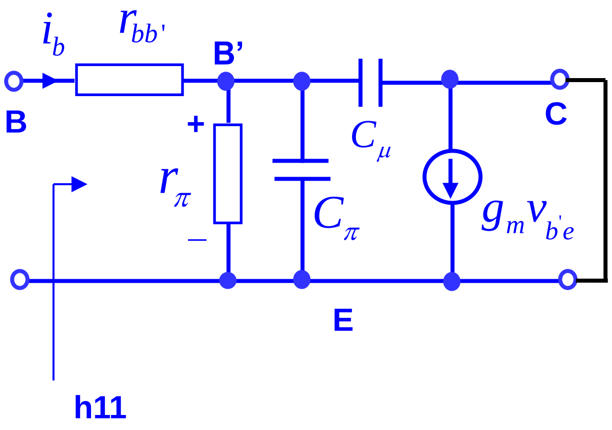


Other uses of S-parameter Measurements:

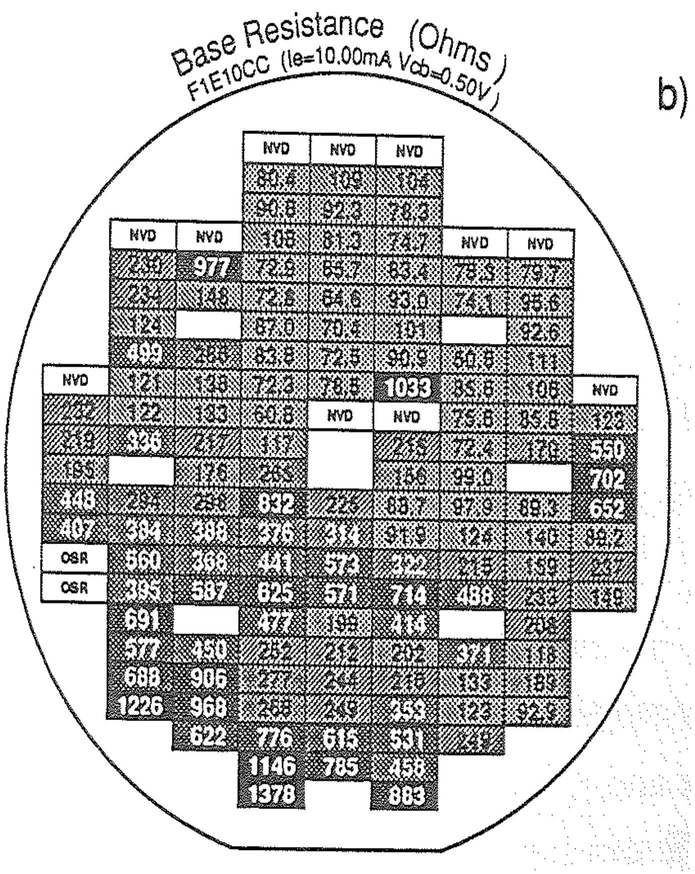
Parameter Extraction for SPICE

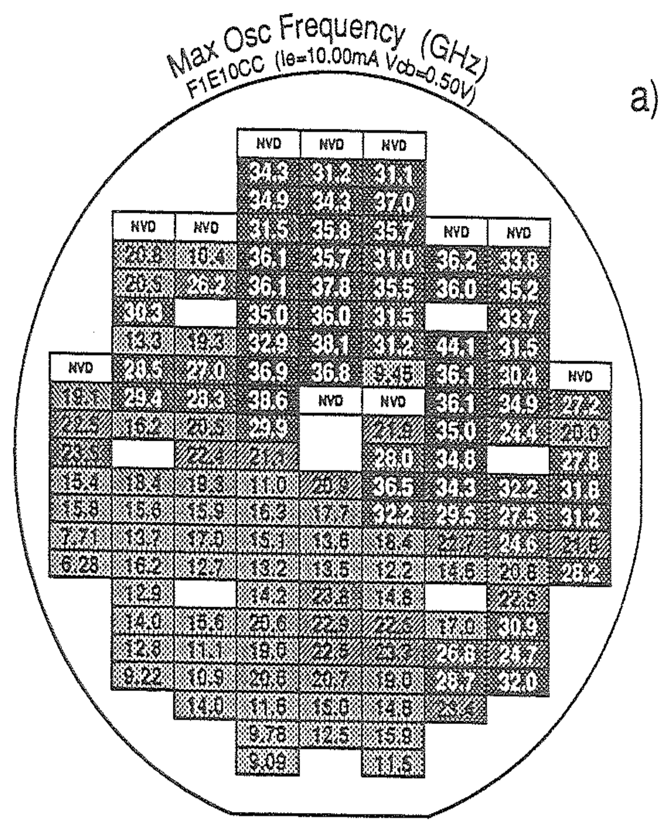
Base resistance extraction from h_{11}





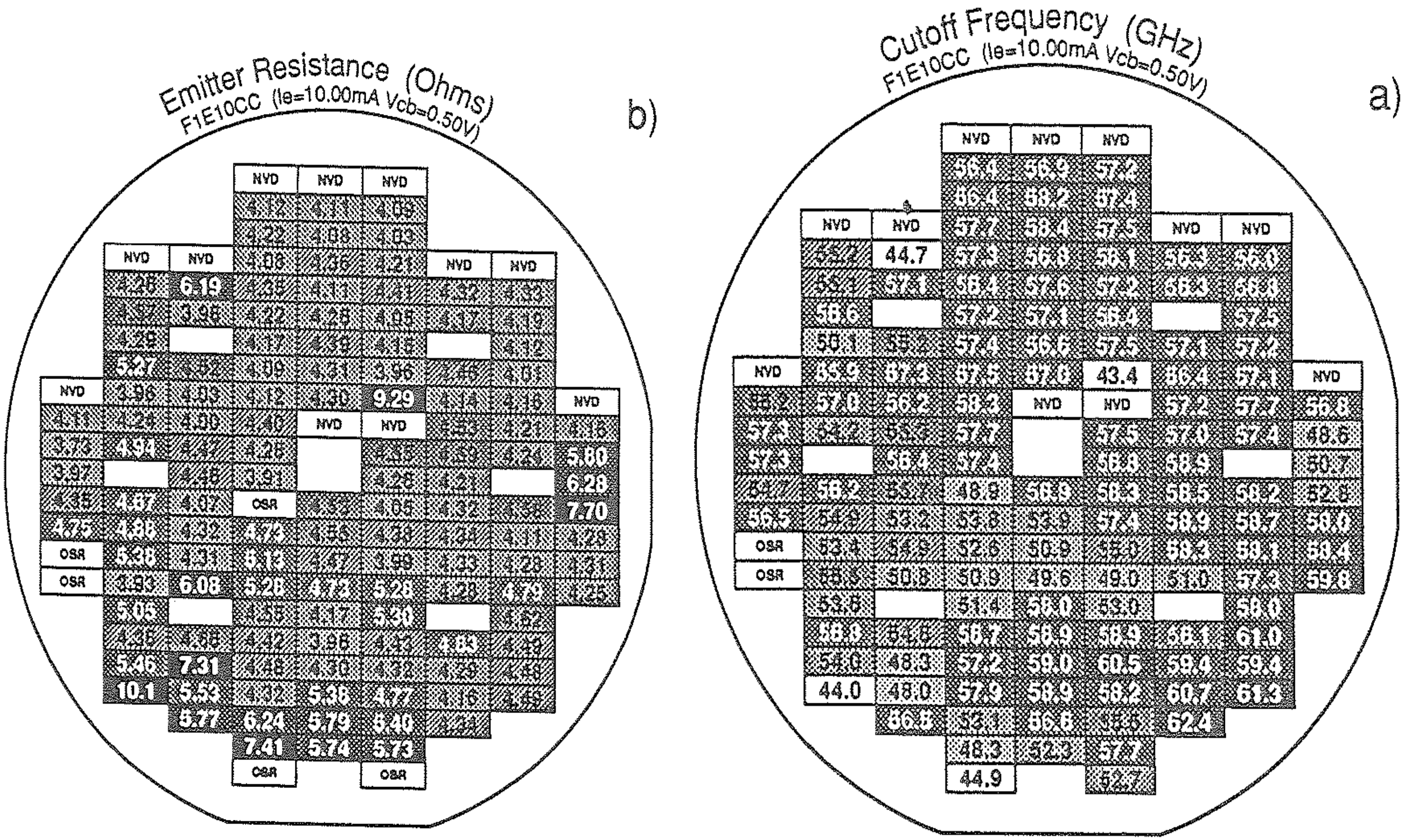
Wafer Map of Base Resistance and Fmax





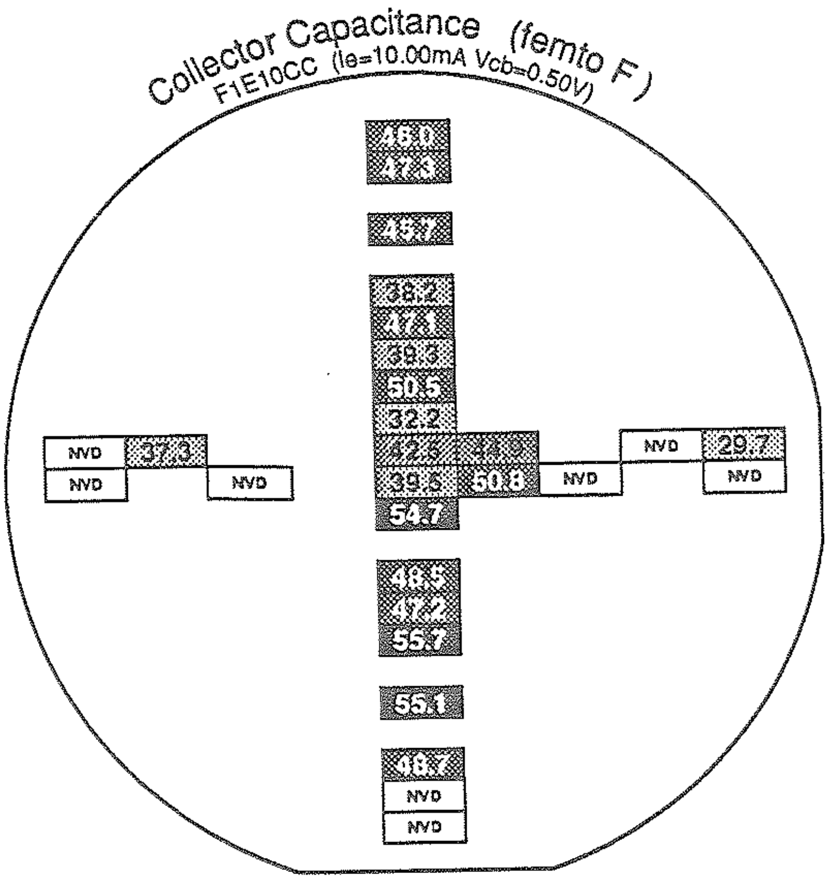
S.J. Prasad, BCTM 1992, pp 204-207

Wafer Map of Emitter resistance and Ft



S.J. Prasad, BCTM 1992, pp 204-207

Wafer Map of Collector Capacitance



S.J. Prasad, GaAs IC Symposium 1992, pp 271- 274

ENGINEERING

It is a great profession.
There is the fascination of watching a figment of imagination emerge through the aid of science to a plan on paper.
Then it moves to realization in stone or metal or energy.
Then it brings jobs and homes to men.
Then it elevates the standards of living and adds comforts to life.
That is the engineer's high privilege.

ENGINEERING - II

The great ability of the engineer compared to men of other professions
Is that his works are out in the open where all can see them.
His acts, step by step are in hard substances.
He cannot bury his mistakes in the grave like doctors;
He cannot argue them into thin air like the lawyers;
He cannot cover his failures with trees and wines like the architects;
He cannot screen his shortcomings by blaming the opponents
like the politicians ;
The engineer simply cannot deny he did it.
If his works does not work he is damned.

-- Herbert Hoover