Finetuning Academy

RF & Wireless matters

2-Day Workshop on RF circuit and system design

November 8-9, 2014

Bangalore

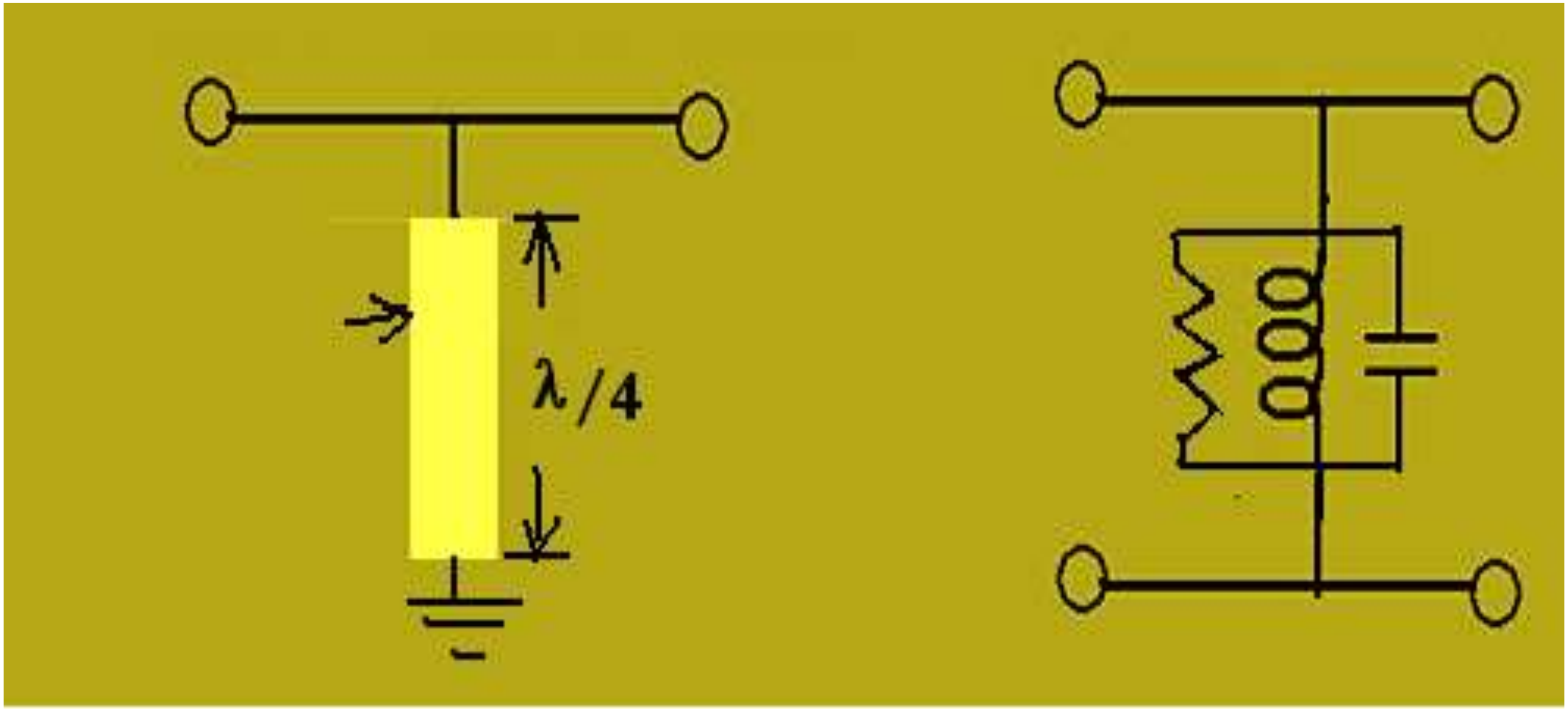
Finetuning Academy

Email: support@finetuningrf.com

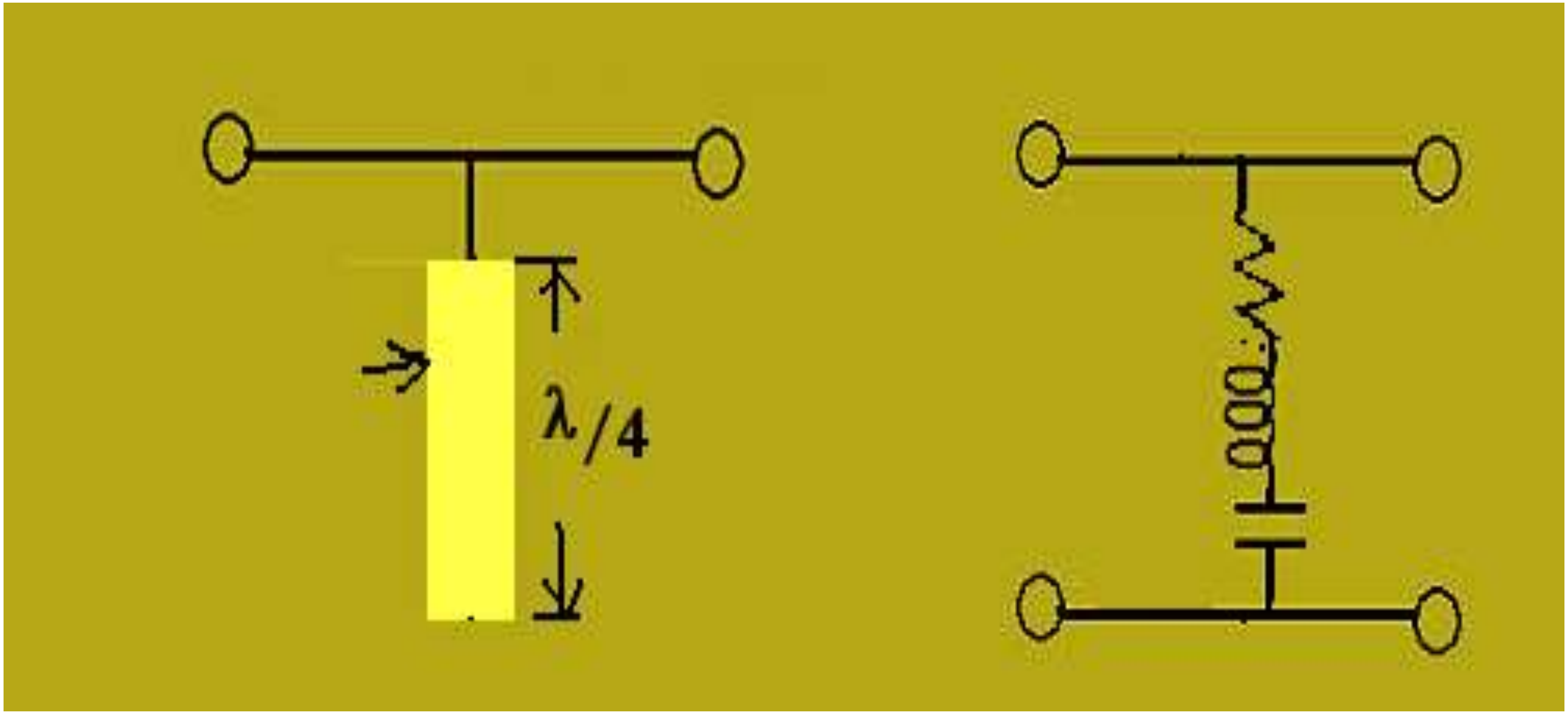
www.finetuningrf.com

Transmission lines

Short-circuited quarter-wave stub



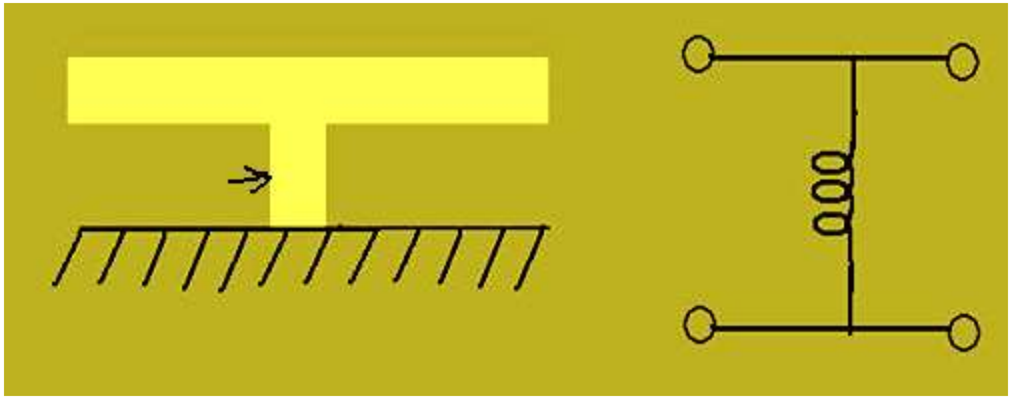
Open-circuited quarter-wave stub



Inductive and capacitive stubs

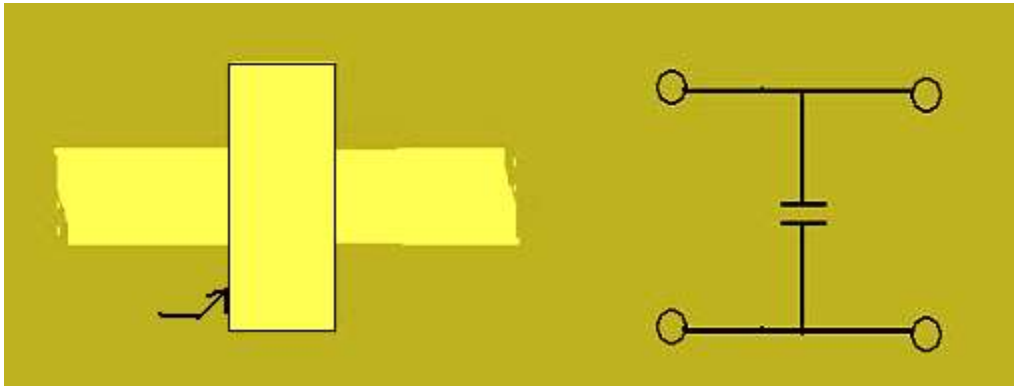
Short-circuited stub

Inductive when l<λg/4



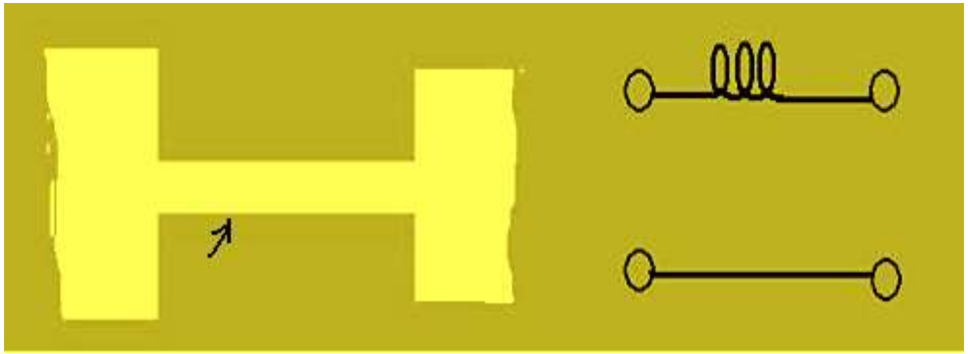
Open-circuited stub

Capacitive when l<λg/4

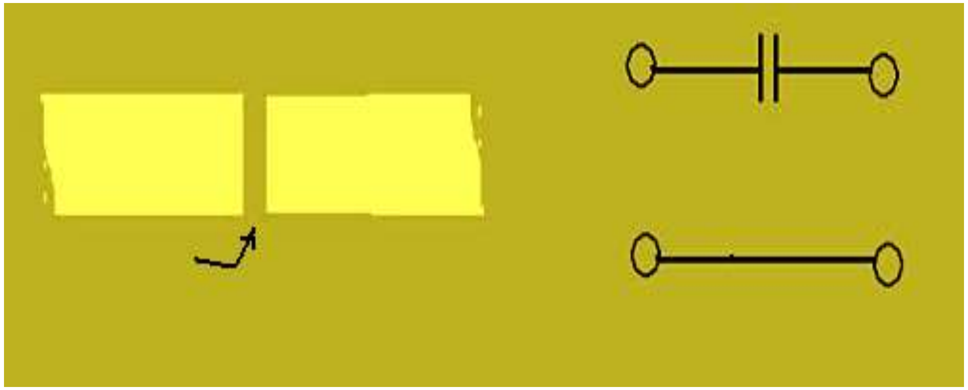


Series inductance and capacitance

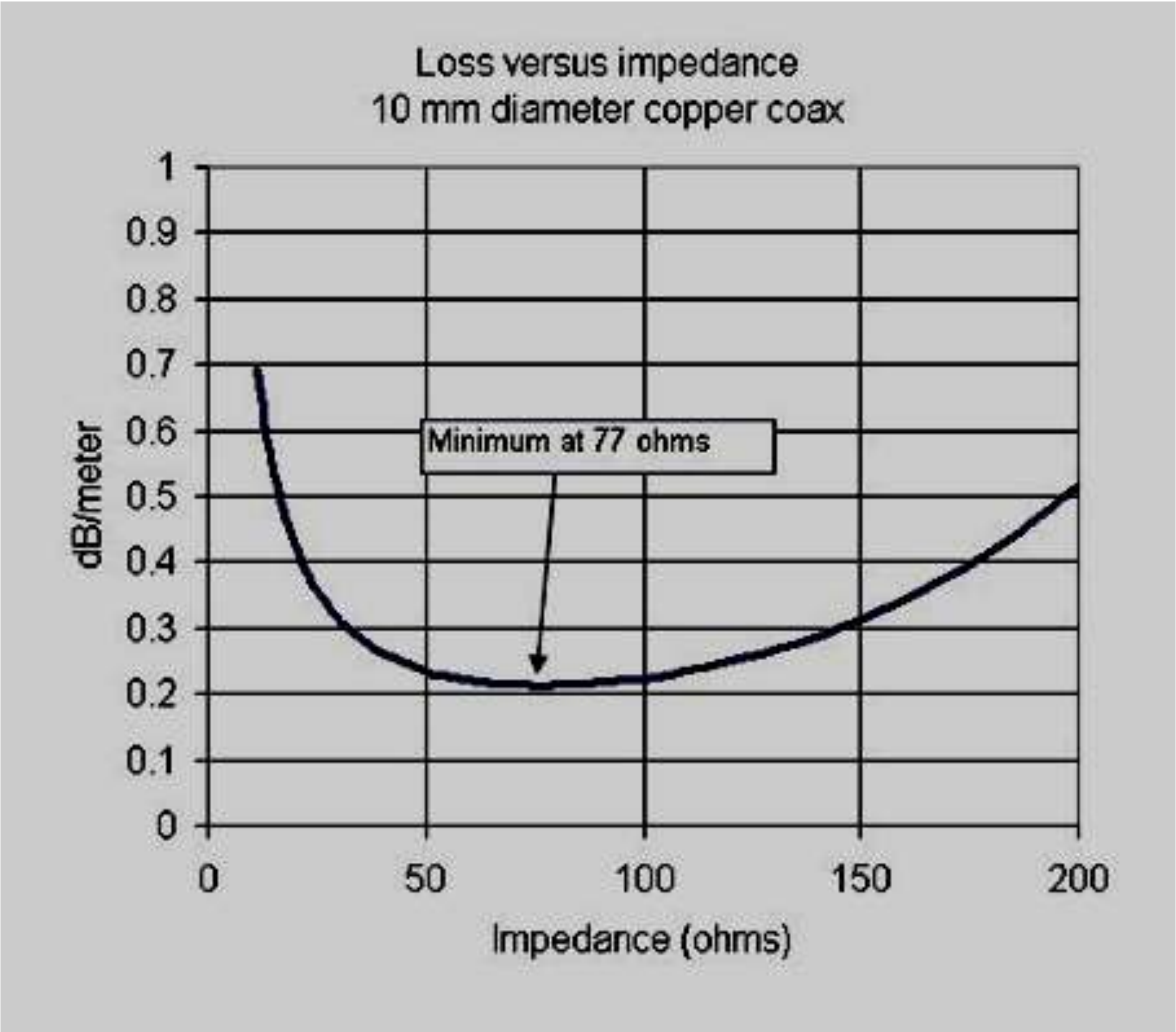
High impedance series line (Equivalent to a series inductance)



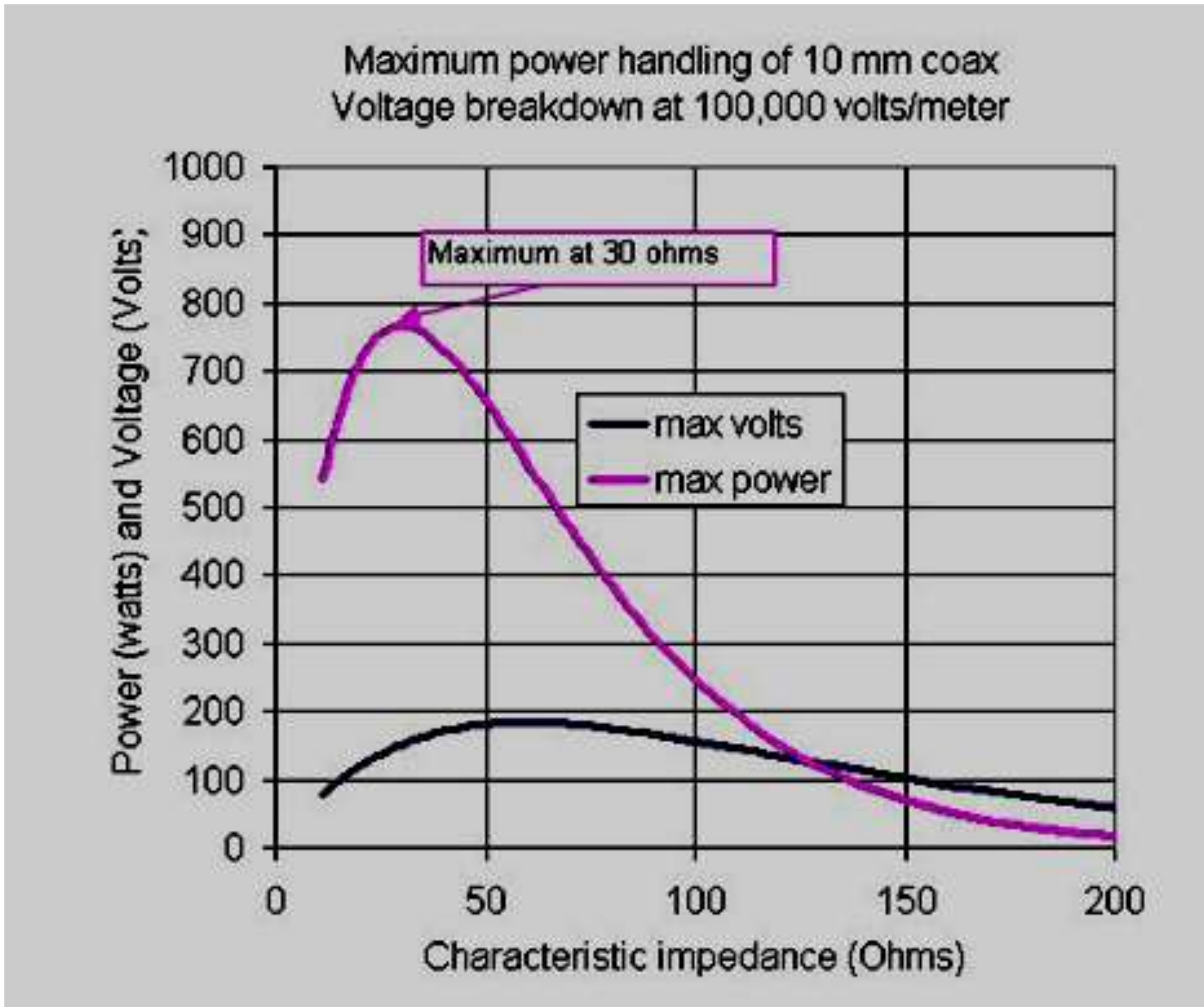
A gap left in the transmission line (Equivalent to a series capacitance)



Choice of characteristic impedance



Choice of characteristic impedance

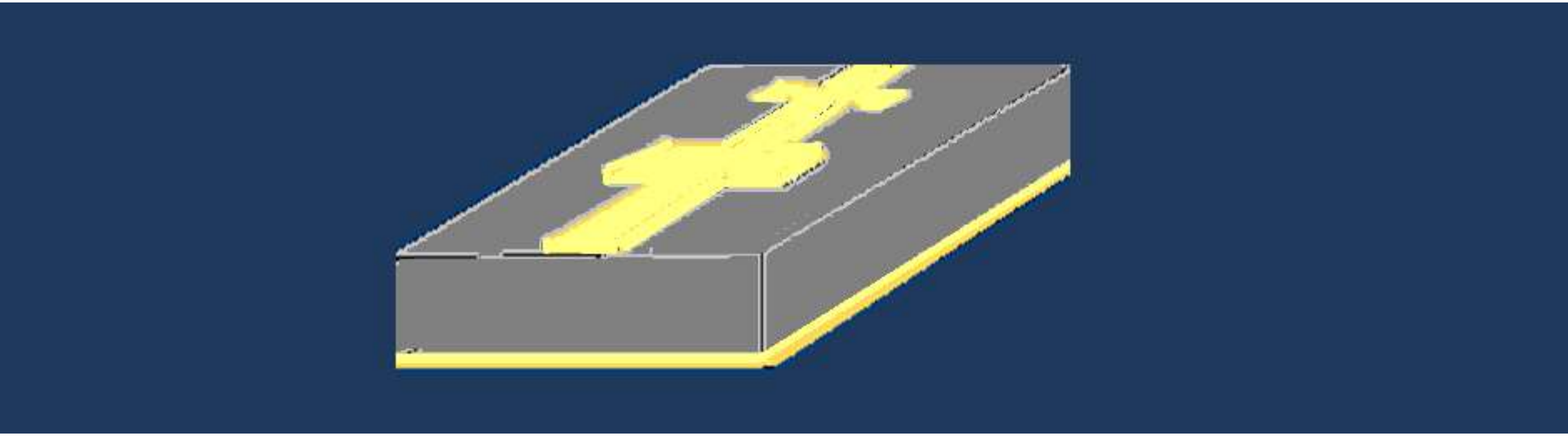


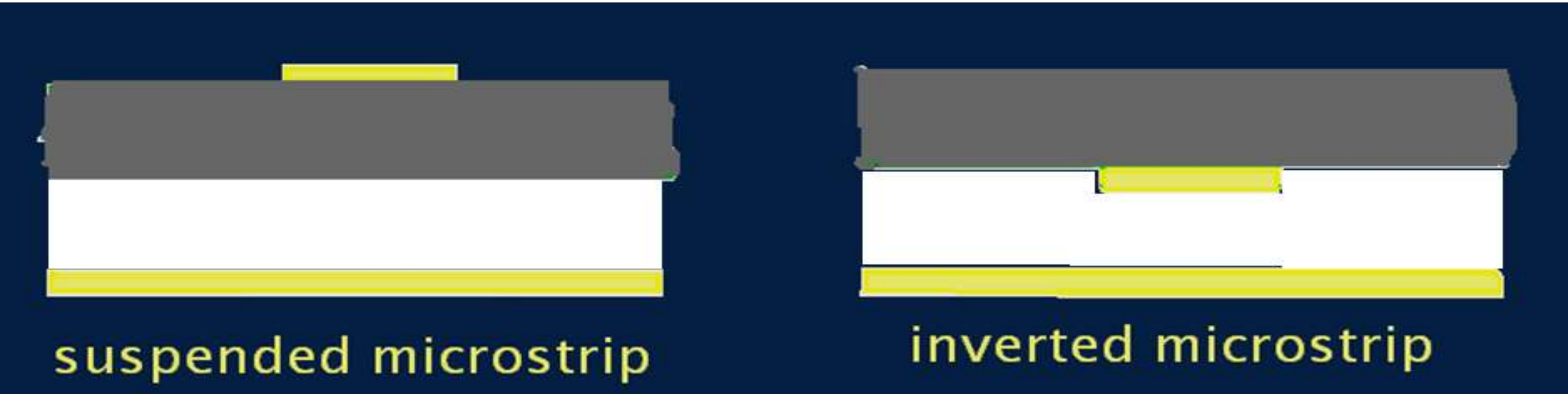
Coaxial cable

L = 460.6xlog (D/d) nH/m
C = 24.13 x εr / log (D/d) pF/m

50 ohm PTFE-filled cable will have 94.8 pF/meter capacitance and 237 nH/meter inductance

Microstrip

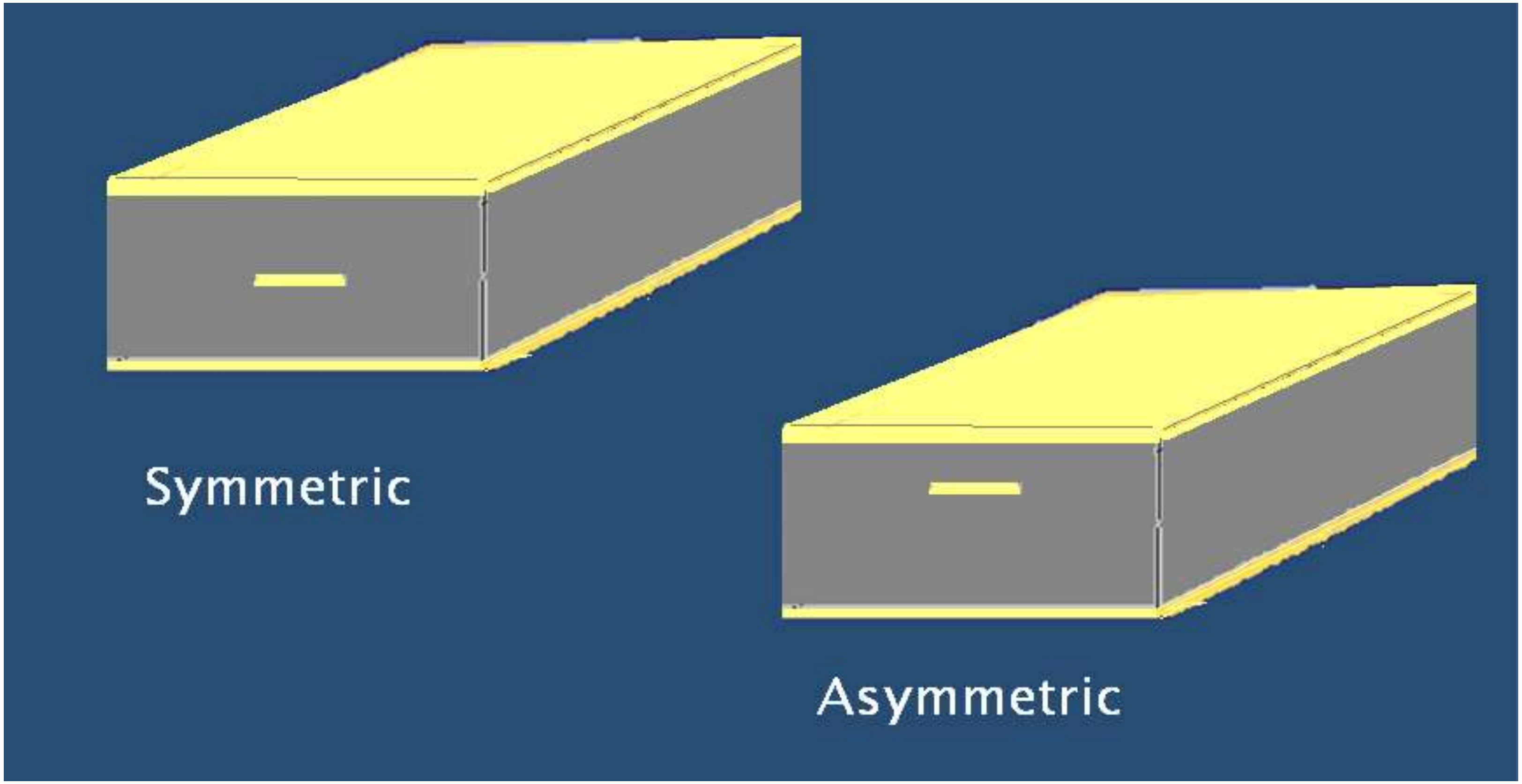




Surface and embedded microstrip

Most common type of transmission line
suitable for both hybrid and monolithic circuits.
Moderately dispersive at high frequencies.

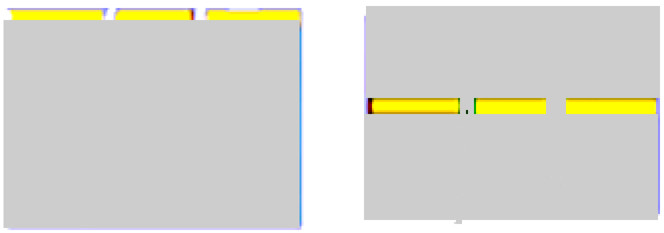
Stripline



Does not allow convenient mounting of discrete circuit elements. Best for passive components. Low loss TEM. Good transition to coax. Better immunity to crosstalk.

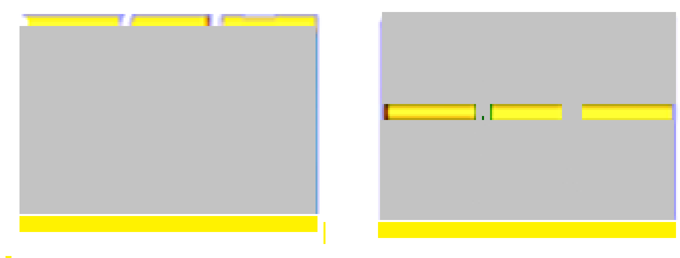
Coplanar Waveguide (CPW)

CPW



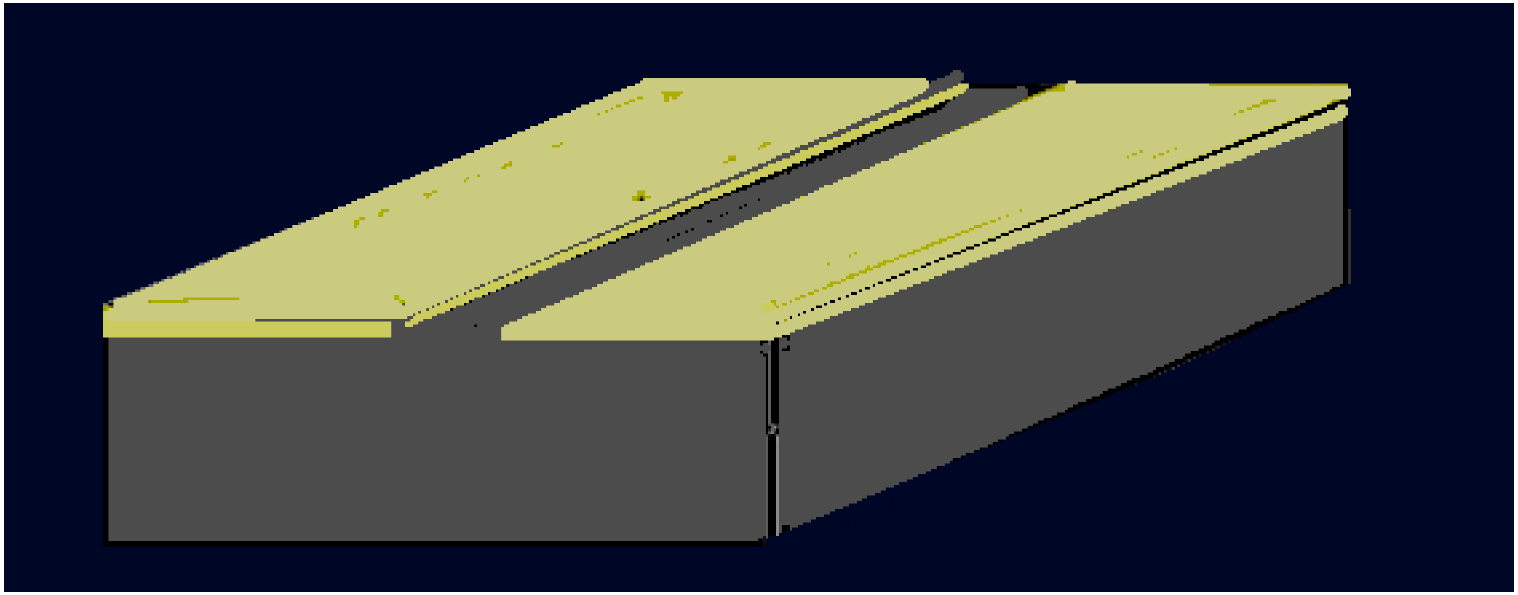
Produces smaller trace per given Zo than microstrip. Circuits can be made denser than microstrip circuit.

CPW with ground plane

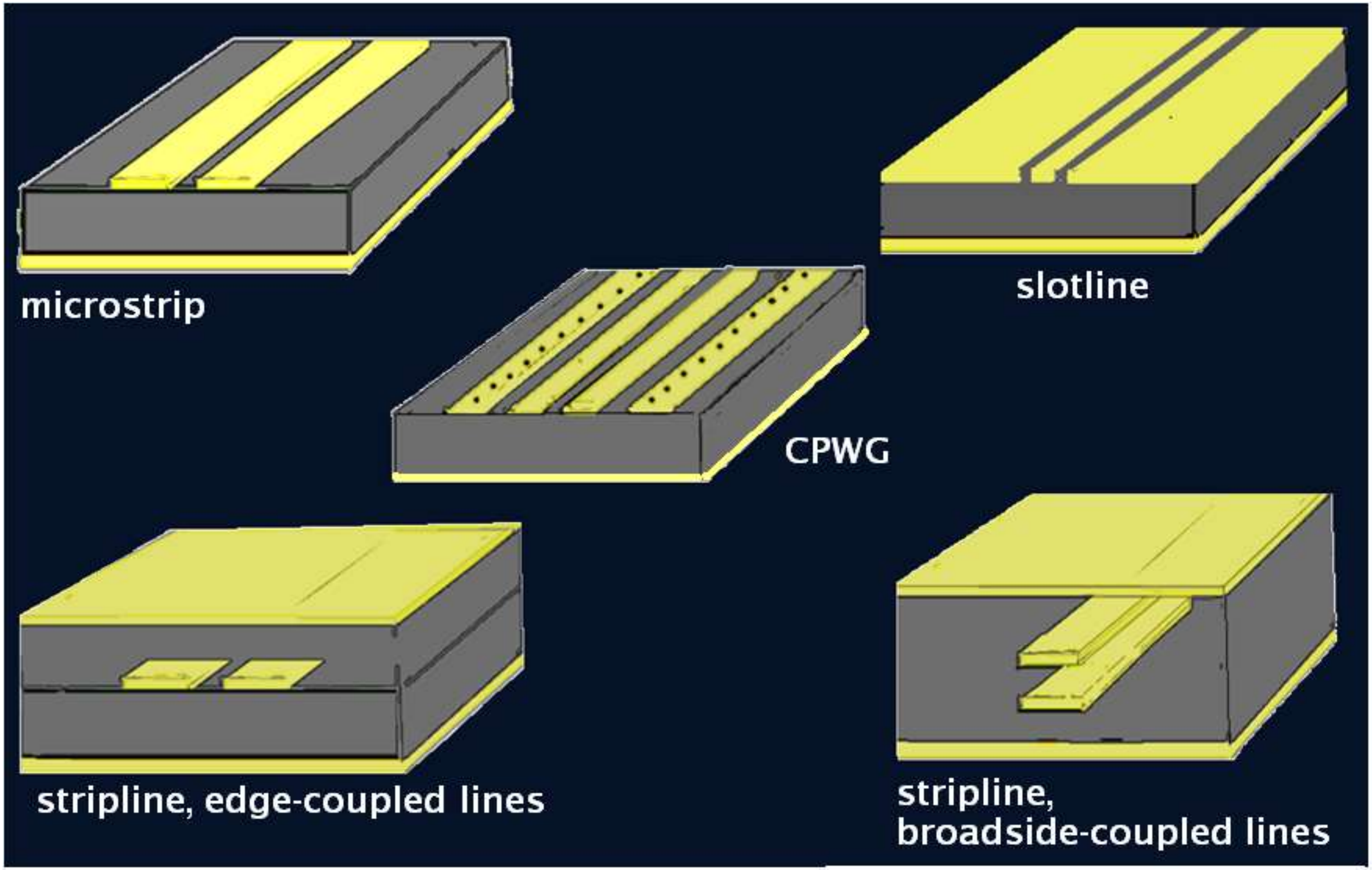


Provides better isolation between nearby RF lines and other signal lines. The lower ground plane acts as a heatsink for circuits with active devices.

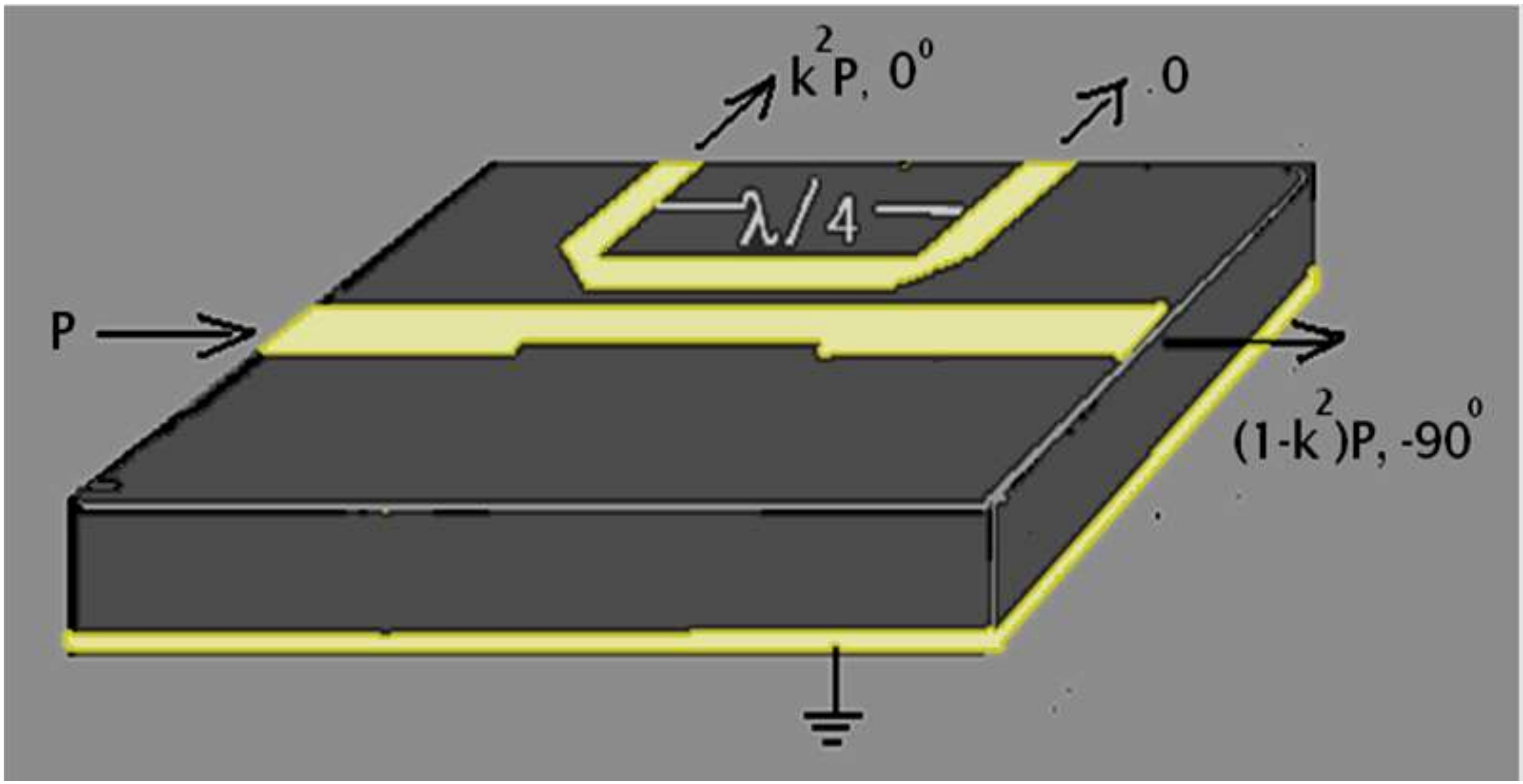
Slotline



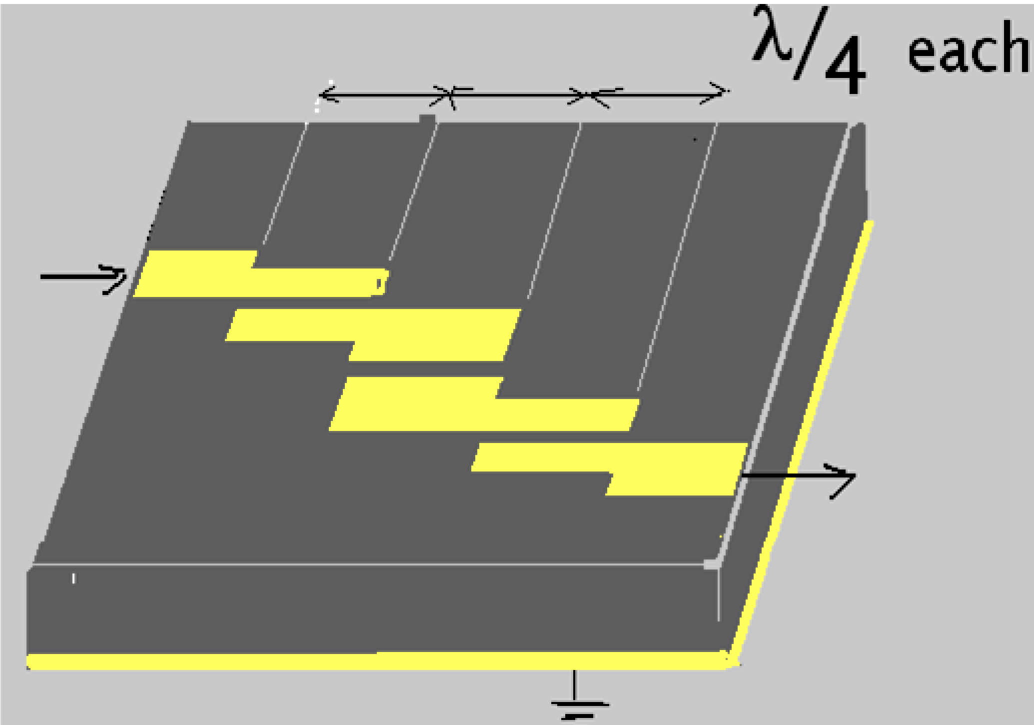
Coupled transmission lines

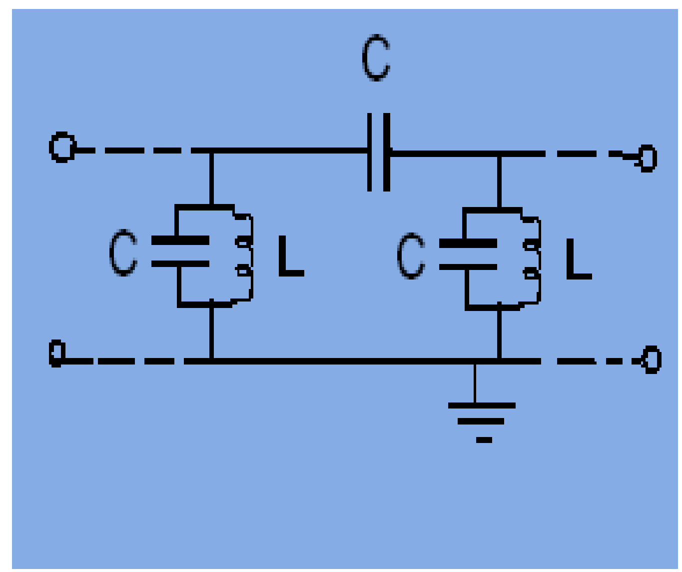


Directional Coupler



Bandpass filter





Rectangular Waveguide

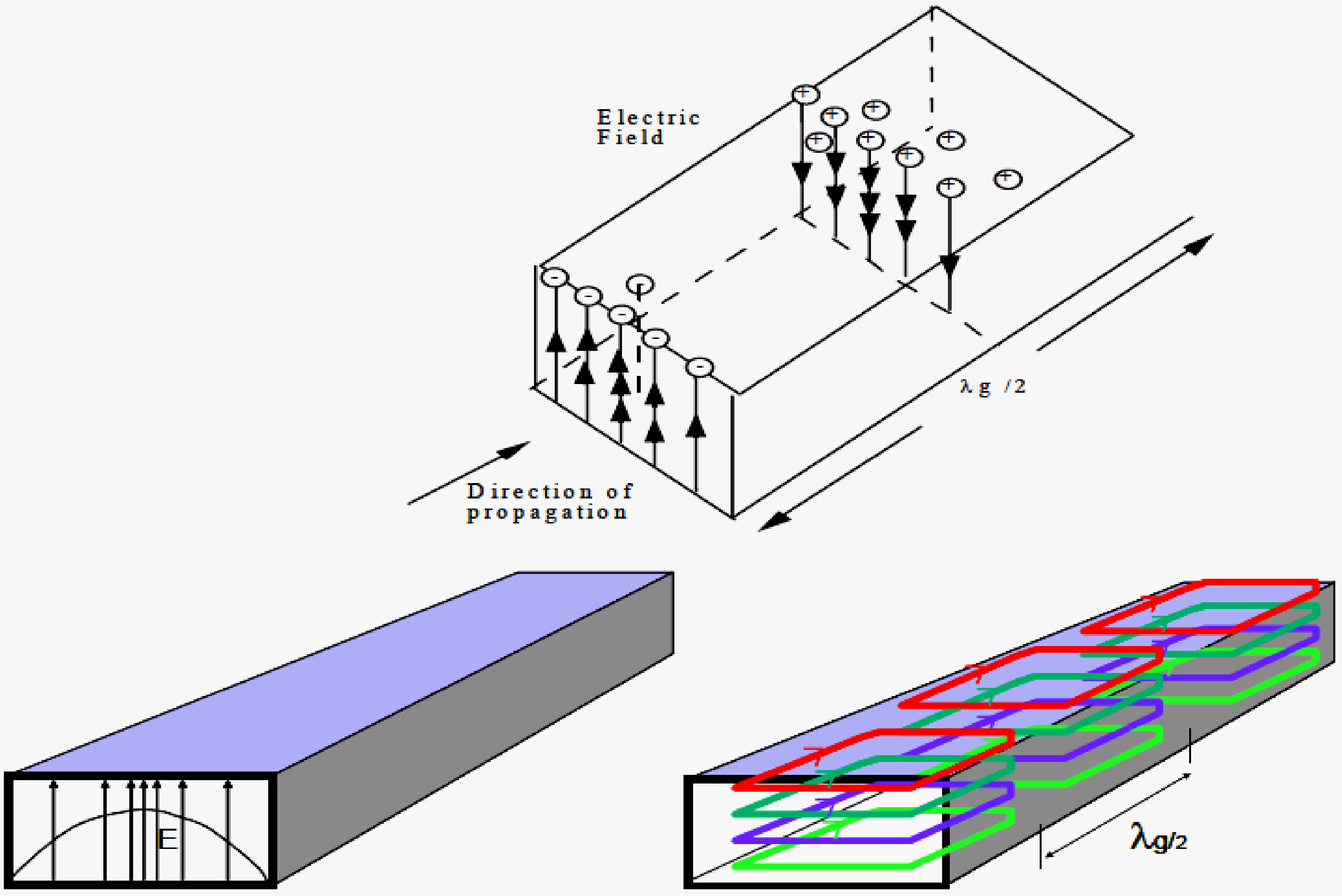
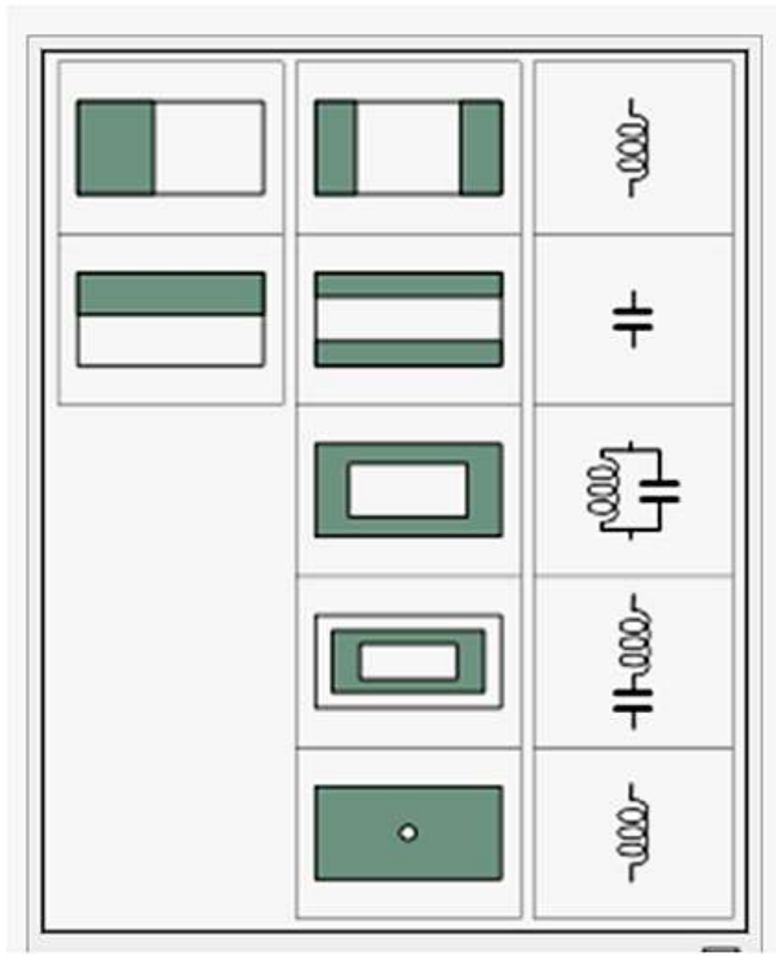
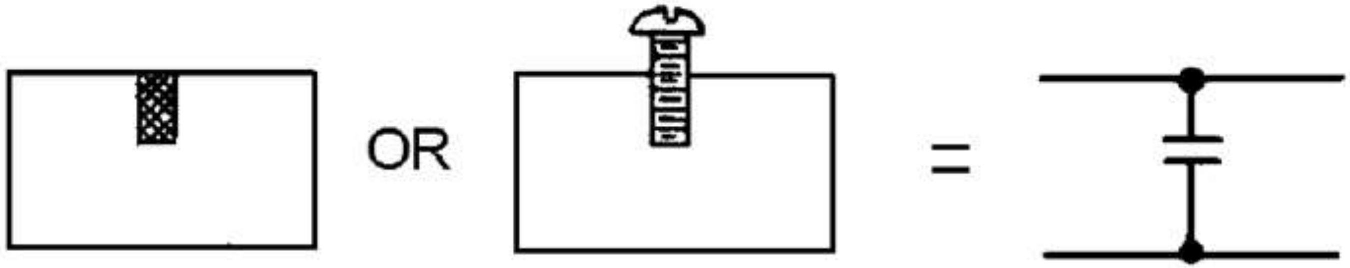


Illustration of the E-field (left) and H-field (right) for the TE10 waveguide mode

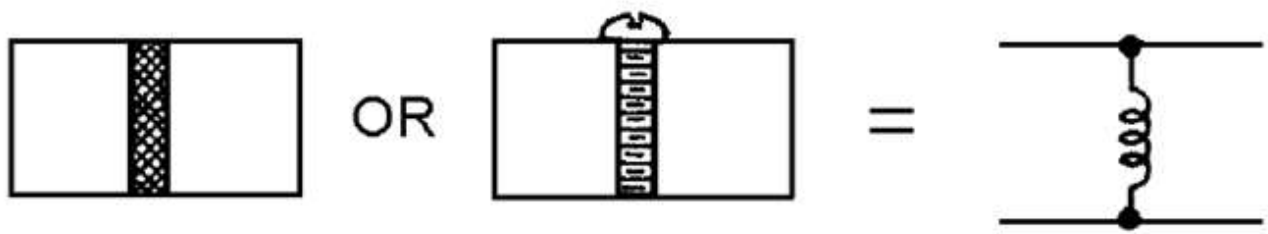
Some waveguide iris geometries and equivalent circuits



Conducting posts and screws

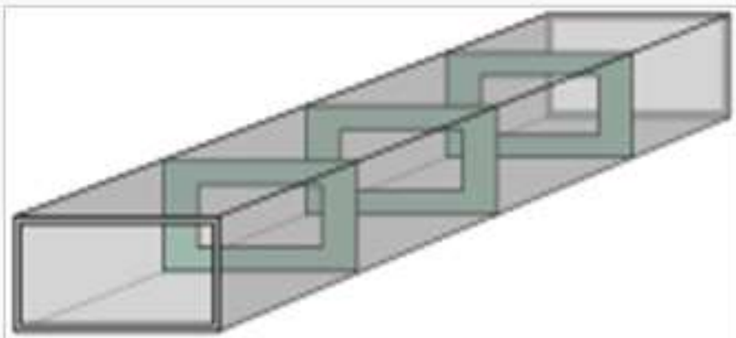


Conducting posts and screws. PENETRATING.

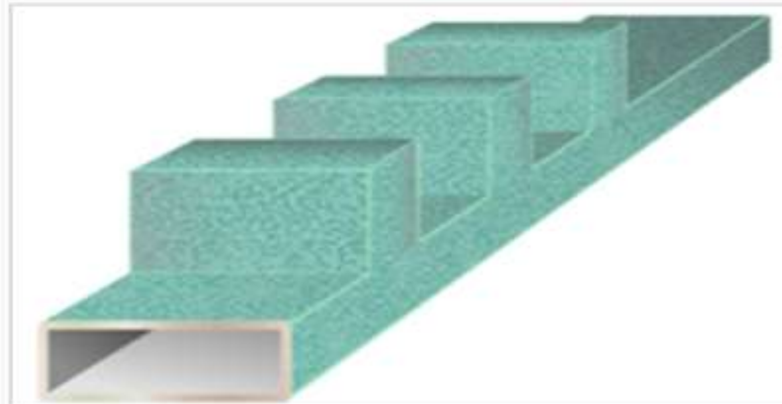


Conducting posts and screws. EXTENDING THROUGH.

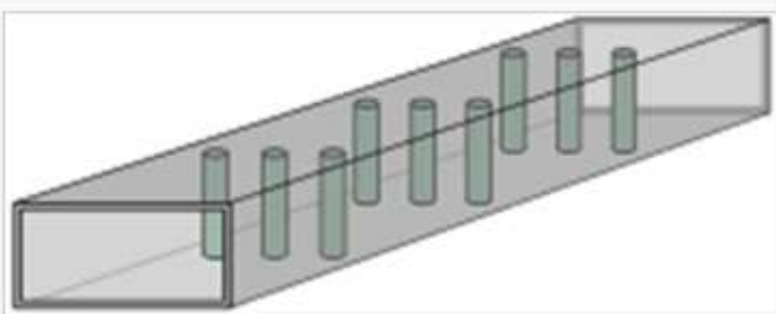
Waveguide filter



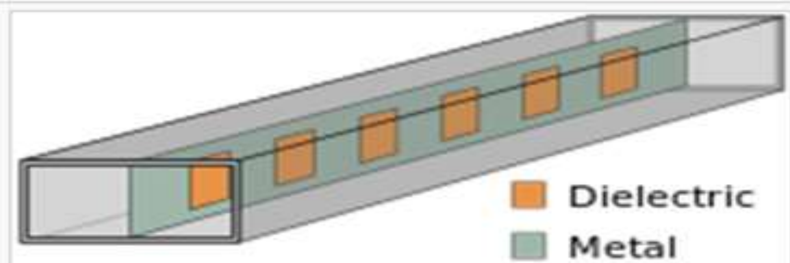
Iris-coupled filter with three irises



Waveguide stub filter
consisting of three stub resonators



Post filter with three rows of posts



Insert filter with six dielectric
resonators in the E-plane.

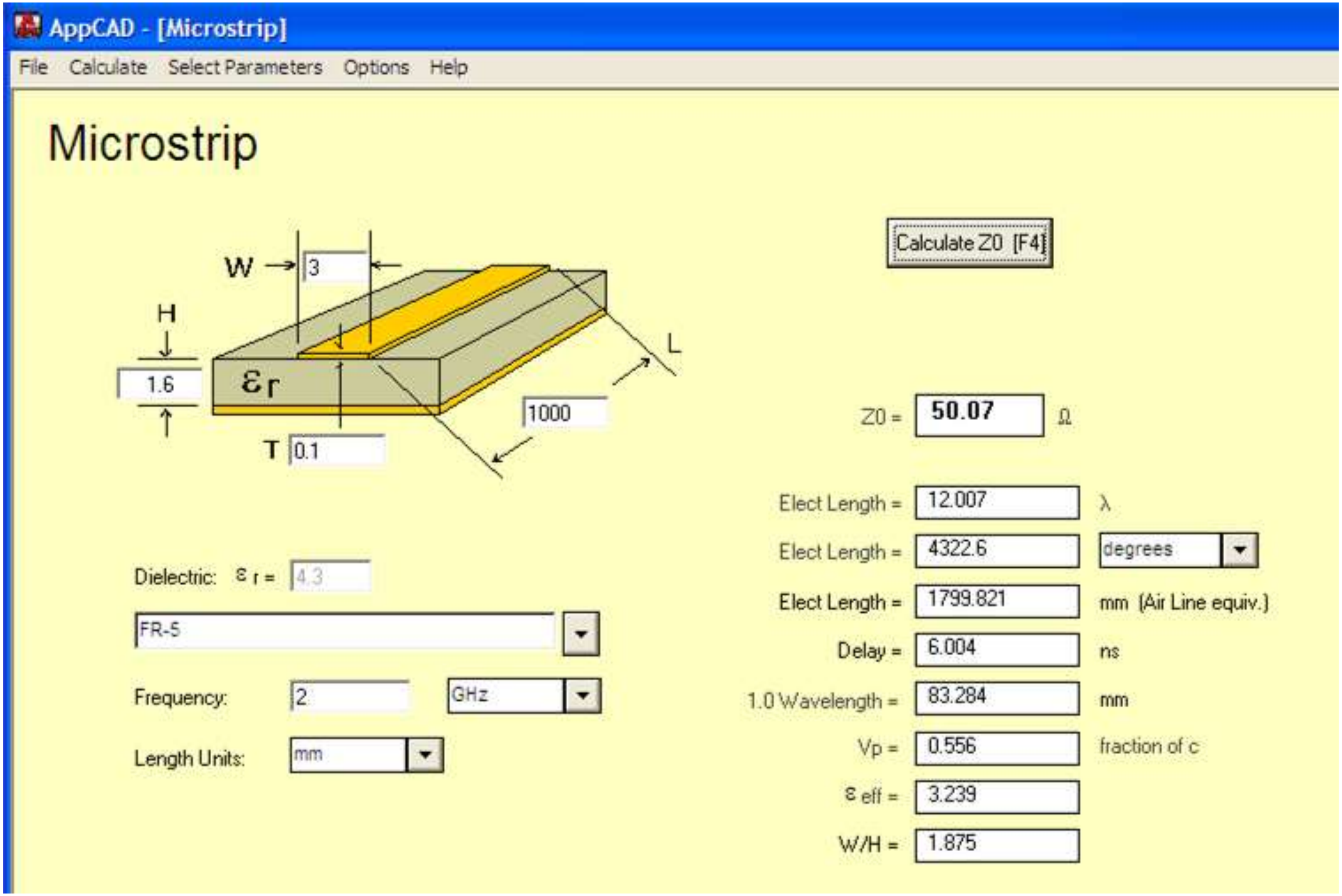
Comparison of transmission lines

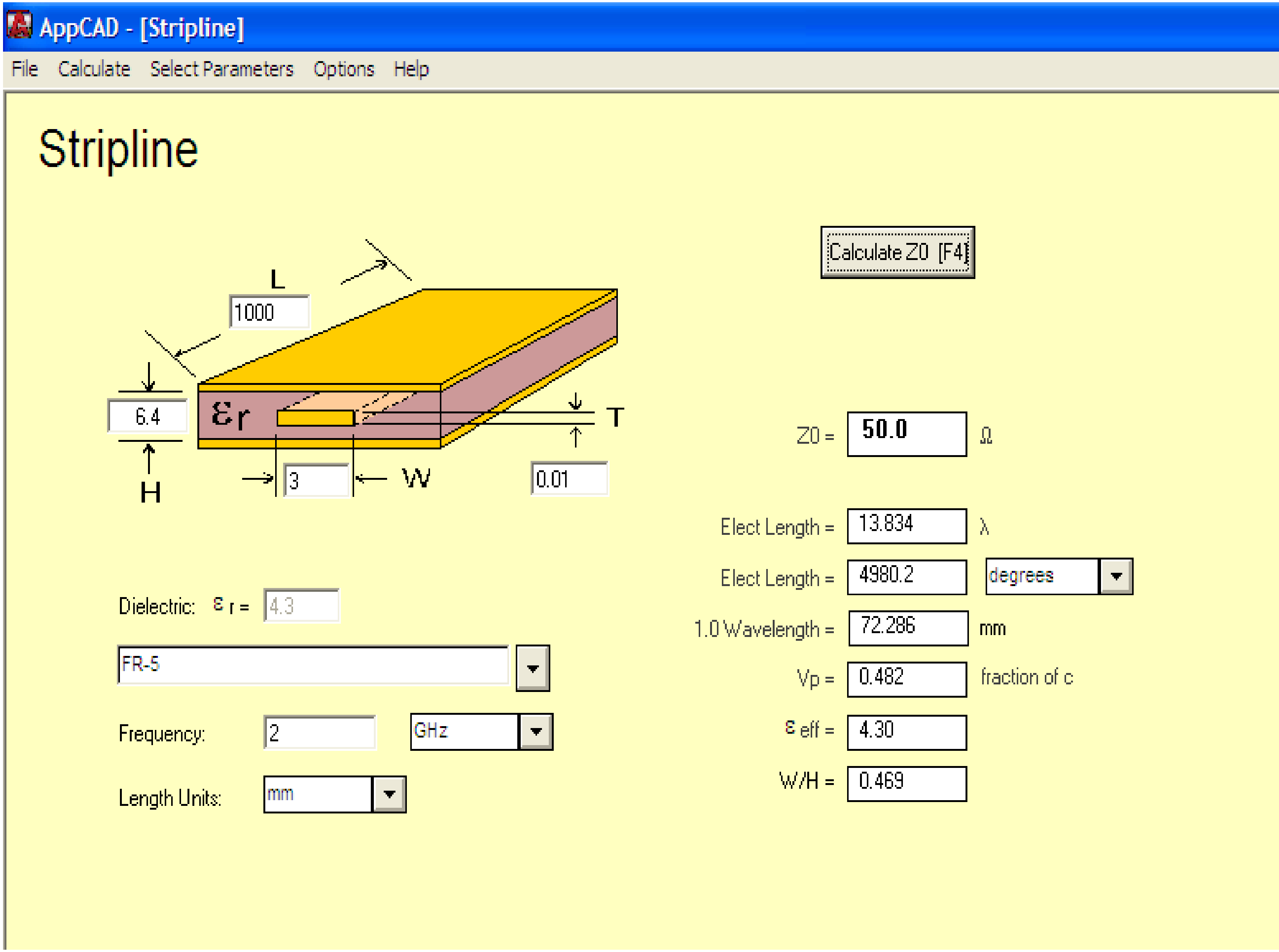
	Waveguide	Coaxial	Stripline	Microstrip
Conductor loss	Very low	low	medium	medium
Q-factor	Very high	high	medium	medium
Maximum power handling	Very high	high	medium	low
Bandwidth	Small	medium	large	large
Isolation between adjacent circuit elements	Very high	Very high	low	low
Weight	Very high	High	low	very low
Miniaturization	Very little	little	high	very high
Possibility of mounting packaged semiconductors	medium	medium	good	very good
Possibility of mounting packaged lumped components	very poor	medium	good	very good
Circuit flexibility	very poor	poor	medium	very good
Series production cost	high	high	medium	low

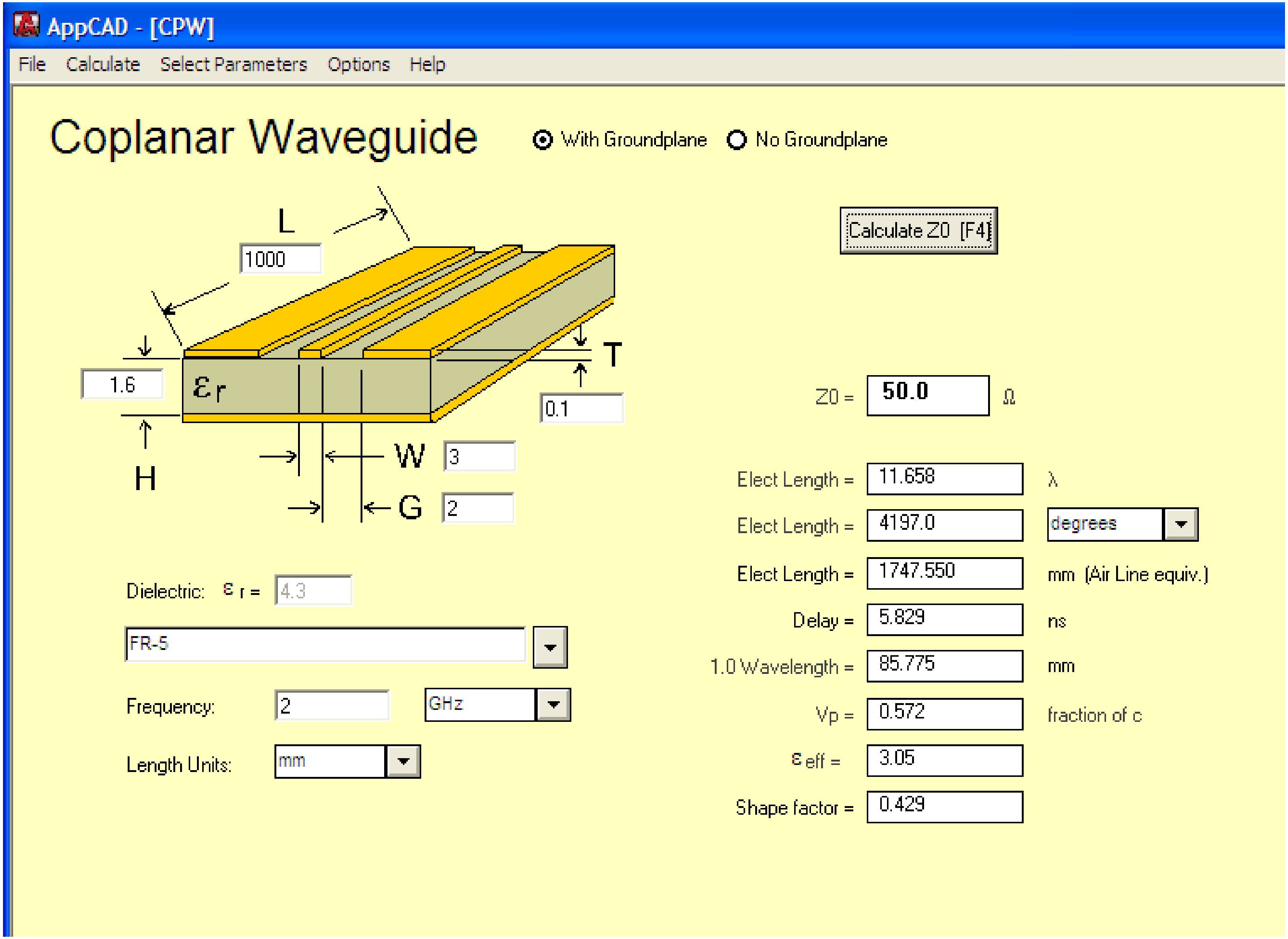
Lumped elements Vs transmission lines

- Lumped elements can be effectively used if the length of the element is very small in comparison to the wavelength.
- At low frequencies, lumped elements have the advantages of smaller size and wider bandwidth as compared to distributive elements
- At high frequencies, however, distributed circuits are used because of lower loss (or higher *Q*), and the size advantage of the lumped element is no longer a significant factor.

AppCAD transmission line calculators




AppCAD - [Stripline]
File
Calculate
Select Parameters
Options
Help
Stripline
Calculate Z0 [F4]
L
1000
6.4
H
εr
3
W
T
0.01
Z0 =
50.0
Ω
Elect Length =
13.834
λ
Elect Length =
4980.2
degrees
1.0 Wavelength =
72.286
mm
Vp =
0.482
fraction of c
ε eff =
4.30
W/H =
0.469
Dielectric: ε r =
4.3
FR-5
Frequency:
2
GHz
Length Units:
mm



AppCAD - [CPW]
File Calculate Select Parameters Options Help
Coplanar Waveguide
With Groundplane
No Groundplane
Calculate Z0 [F4]
L
1000
1.6
εr
T
0.1
H
W
3
G
2
Z0 =
50.0
Ω
Elect Length =
11.658
λ
Elect Length =
4197.0
degrees
Elect Length =
1747.550
mm (Air Line equiv.)
Delay =
5.829
ns
1.0 Wavelength =
85.775
mm
Vp =
0.572
fraction of c
ε eff =
3.05
Shape factor =
0.429
Dielectric: ε r =
4.3
FR-5
Frequency:
2
GHz
Length Units:
mm

S-parameters

S-parameters

$$\begin{bmatrix} \mathbf{b}_1 \\ \mathbf{b}_2 \end{bmatrix} = \begin{bmatrix} \mathbf{S}_{11} & \mathbf{S}_{12} \\ \mathbf{S}_{21} & \mathbf{S}_{22} \end{bmatrix} \begin{bmatrix} \mathbf{a}_1 \\ \mathbf{a}_2 \end{bmatrix}$$

Definition

$$s_{11} = \left.\frac{b_1}{a_1}\right|_{a_2=0}$$ = Input reflection coefficient Γin for case of $Z_L=Z_0$

$$s_{21} = \left.\frac{b_2}{a_1}\right|_{a_2=0}$$ = Forward transmission gain for case of $Z_L=Z_0$

$$s_{12} = \left.\frac{b_1}{a_2}\right|_{a_1=0}$$ = Reverse transmission gain for case of $Z_s=Z_0$

$$s_{22} = \left.\frac{b_2}{a_2}\right|_{a_1=0}$$ = Output reflection coefficient Γout for case of $Z_s=Z_0$

$$(S_{11}) \quad (one-port)$$

$$\begin{pmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{pmatrix} \quad (two-port)$$

$$\begin{pmatrix} S_{11} & S_{12} & S_{13} \\ S_{21} & S_{22} & S_{23} \\ S_{31} & S_{32} & S_{33} \end{pmatrix} \quad (three-port)$$

Etc.

For passive parts, S21 and S12 can't be greater than 0 dB.
For passive reciprocal parts, S12 must equal S21.
Example of **passive non-reciprocal** parts are circulators and isolators.

1-port networks

- Simple lumped elements are 1-ports, but also cavities with a single RF feeding port, terminated transmission lines (coaxial, waveguide, microstrip etc.) or antennas can be considered as 1-ports
- 1-ports are characterized by their reflection coefficient ρ, or in terms of S parameters, by S_{11}.
- Ideal short: $S_{11} = -1$
- Ideal termination: $S_{11} = 0$
- Ideal open: $S_{11} = +1$
- Active termination (reflection amplifier): $|S11| > 1$

2-port networks

- Ideal transmission line of length *l*

$$S = \begin{pmatrix} 0 & e^{-\gamma l} \\ e^{-\gamma l} & 0 \end{pmatrix}$$

For a lossless line we get $|S_{21}| = 1$.

- Ideal phase shifter

$$S = \begin{pmatrix} 0 & e^{-j\varphi_{12}} \\ e^{-j\varphi_{21}} & 0 \end{pmatrix}$$

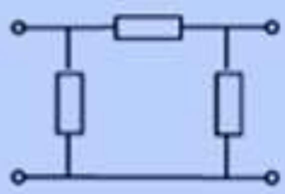
For a reciprocal phase shifter $\varphi_{12} = \varphi_{21}$, while for the gyrator $\varphi_{12} = \varphi_{21} + \pi$

2-port networks

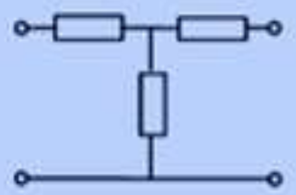
- Ideal attenuator

$$\mathbf{S} = \begin{pmatrix} 0 & e^{-\alpha} \\ e^{-\alpha} & 0 \end{pmatrix}$$

- An attenuator can be realized e.g. with three resistors in a T or Pi circuit



Pi circuit



T circuit

or with resistive material in a waveguide.

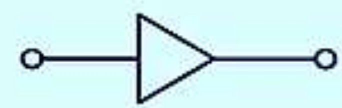
2-port networks

- Ideal isolator

$$\mathbf{S} = \begin{pmatrix} 0 & 0 \\ 1 & 0 \end{pmatrix}$$

- Ideal amplifier

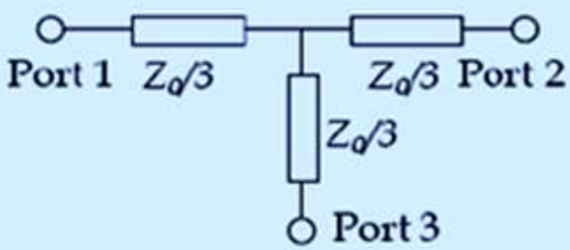
$$\mathbf{S} = \begin{pmatrix} 0 & 0 \\ G & 0 \end{pmatrix}$$



3-port network

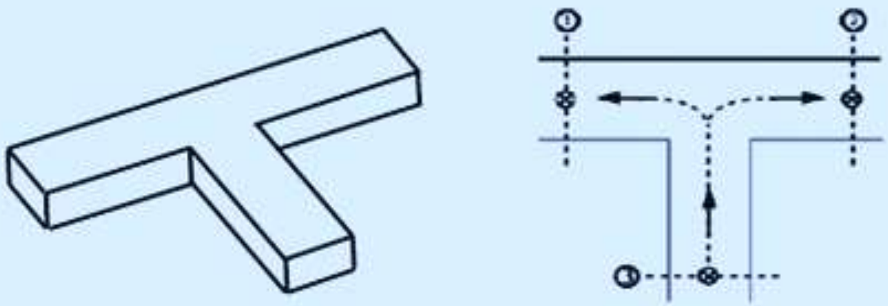
- It can be shown that a 3-port cannot be lossless, reciprocal and matched at all three ports at the same time.

- Resistive power divider: It consists of a resistor network and is reciprocal, matched at all ports but lossy. It can be realized with three resistors in a triangle configuration.



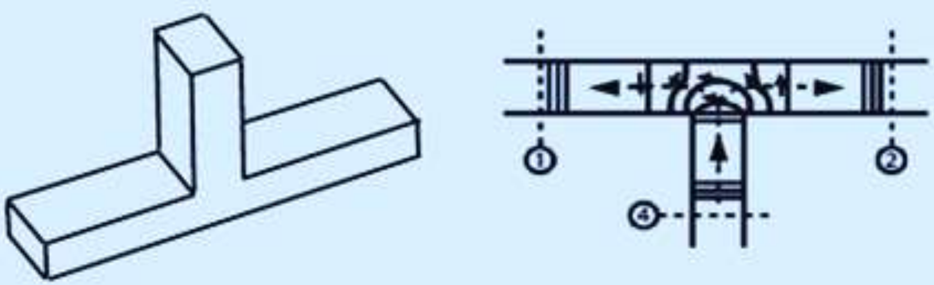
T splitter

- The T splitter is reciprocal and lossless but not matched at all ports. Using the losslessness condition and symmetry considerations one finds for E and H plane splitters



H-plane splitter

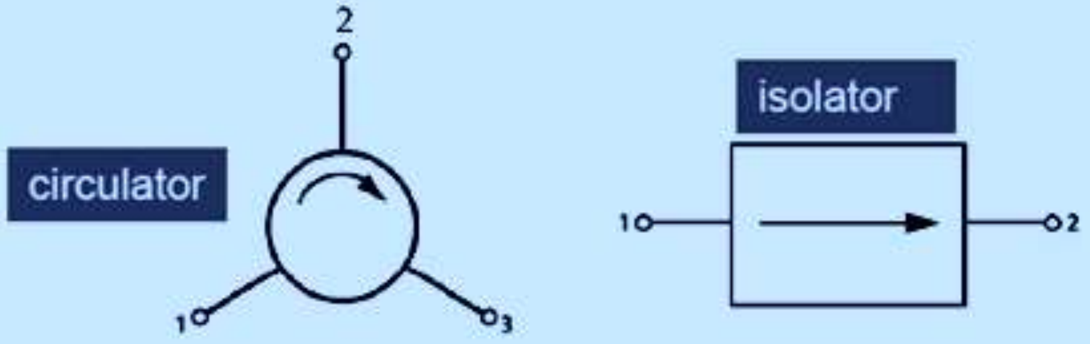
$$S_H = \frac{1}{2}\begin{pmatrix} 1 & -1 & \sqrt{2} \\ -1 & 1 & \sqrt{2} \\ \sqrt{2} & \sqrt{2} & 0 \end{pmatrix}$$



E-plane splitter

$$S_E = \frac{1}{2}\begin{pmatrix} 1 & 1 & \sqrt{2} \\ 1 & 1 & -\sqrt{2} \\ \sqrt{2} & -\sqrt{2} & 0 \end{pmatrix}$$

Circulator and isolator



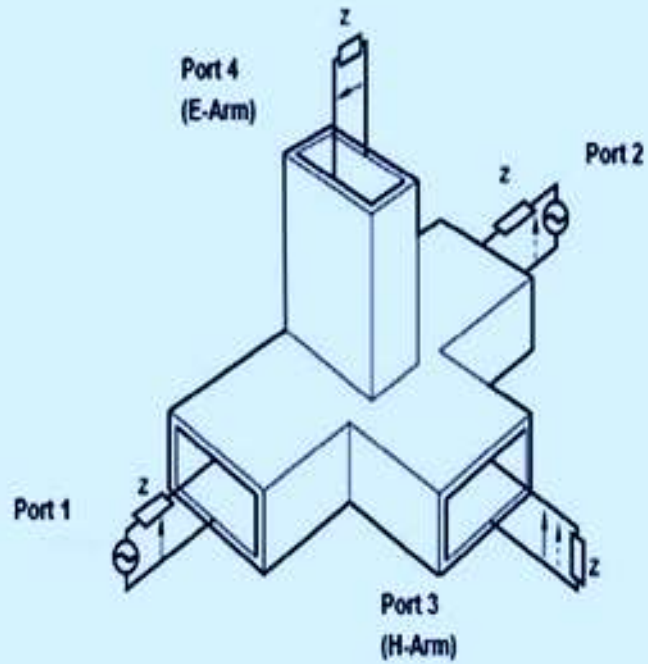
- Its S matrix has a very simple form:

$$\mathbf{S} = \begin{pmatrix} 0 & 0 & 1 \\ 1 & 0 & 0 \\ 0 & 1 & 0 \end{pmatrix}$$

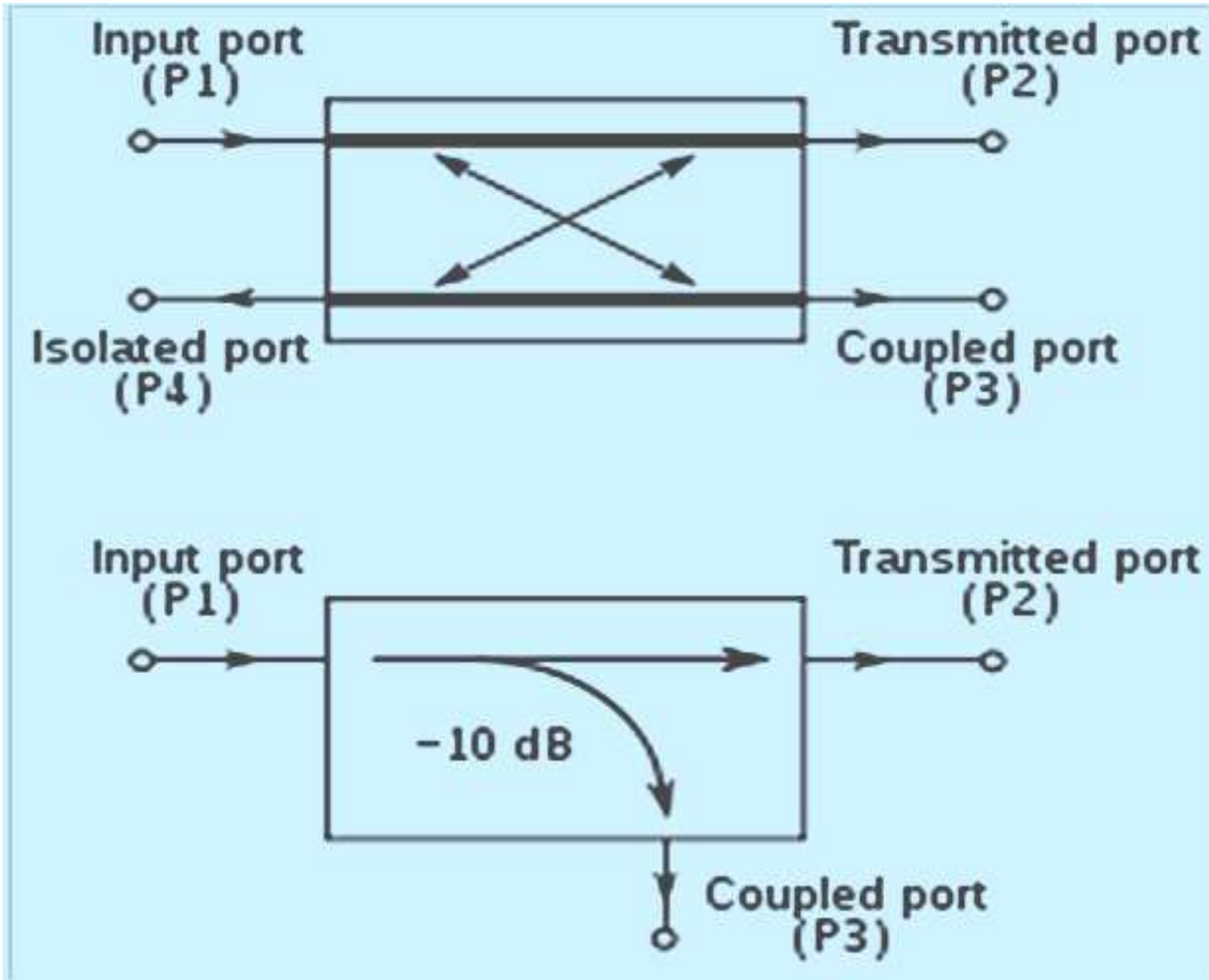
- When port 3 of the circulator is terminated with a matched load we get a two-port called isolator, which lets power pass only from port 1 to port 2

Magic T

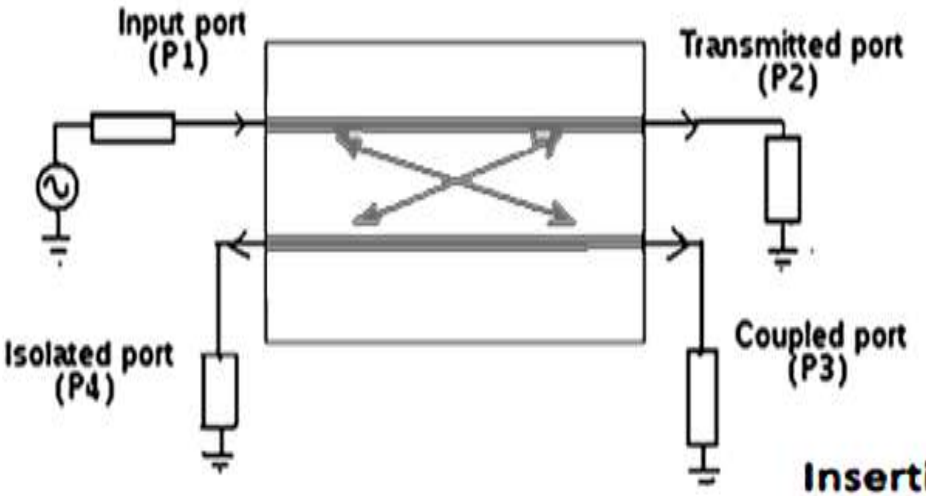
$$S = \frac{1}{\sqrt{2}}\begin{pmatrix} 0 & 0 & 1 & 1 \\ 0 & 0 & 1 & -1 \\ 1 & 1 & 0 & 0 \\ 1 & -1 & 0 & 0 \end{pmatrix}$$



Directional Coupler



Directional Coupler parameters



Insertion loss $L_{i2,1} = -10 \log\left(\frac{P_2}{P_1}\right)\ \mathrm{dB}$

Coupling factor $C_{3,1} = -10 \log\left(\frac{P_3}{P_1}\right)\ \mathrm{dB}$

Isolation $I_{4,1} = -10 \log\left(\frac{P_4}{P_1}\right)\ \mathrm{dB}$

Directivity $D_{3,4} = -10 \log\left(\frac{P_4}{P_3}\right)\ \mathrm{dB}$

S-parameters of directional coupler

$$\mathbf{S} = \begin{bmatrix} 0 & \tau & \kappa & 0 \\ \tau & 0 & 0 & \kappa \\ \kappa & 0 & 0 & \tau \\ 0 & \kappa & \tau & 0 \end{bmatrix}$$

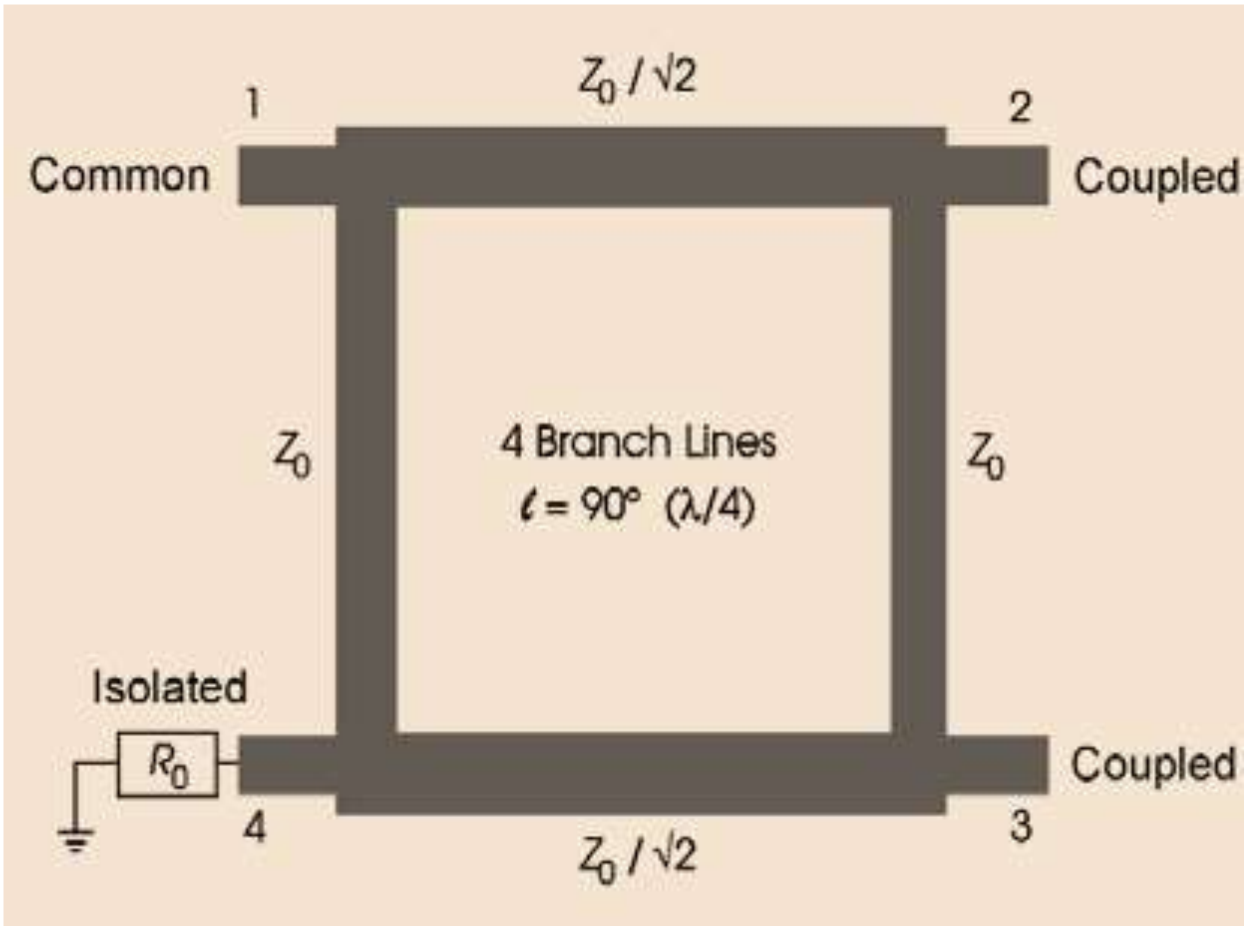
τ is the transmission coefficient and

κ is the coupling coefficient

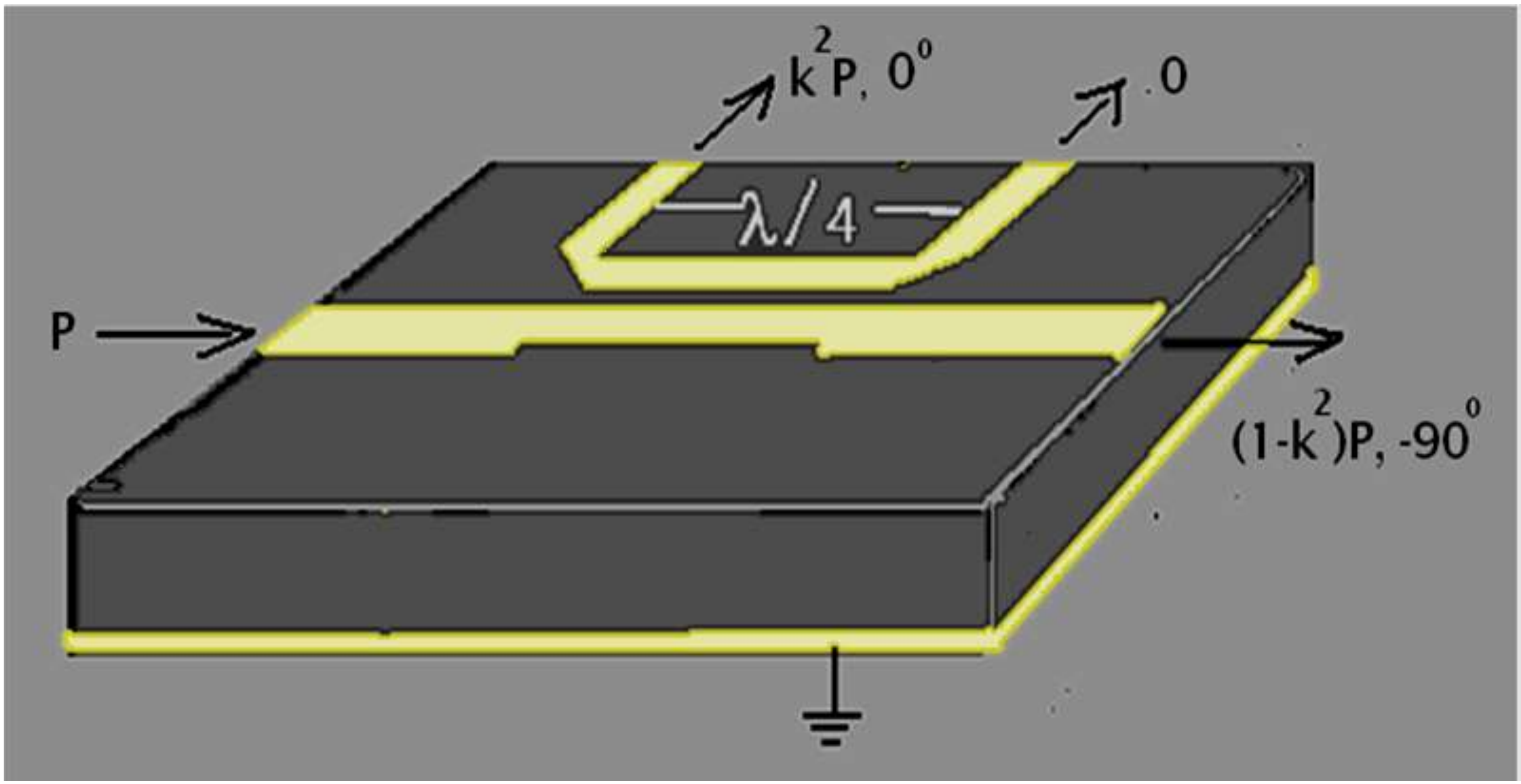
- As an example the S matrix of the 3 dB coupler (π/2-hybrid) can be derived as

$$S_{3dB} = \begin{pmatrix} 0 & 1/\sqrt{2} & \pm j/\sqrt{2} & 0 \\ 1/\sqrt{2} & 0 & 0 & \pm j/\sqrt{2} \\ \pm j/\sqrt{2} & 0 & 0 & 1/\sqrt{2} \\ 0 & \pm j/\sqrt{2} & 1/\sqrt{2} & 0 \end{pmatrix}$$

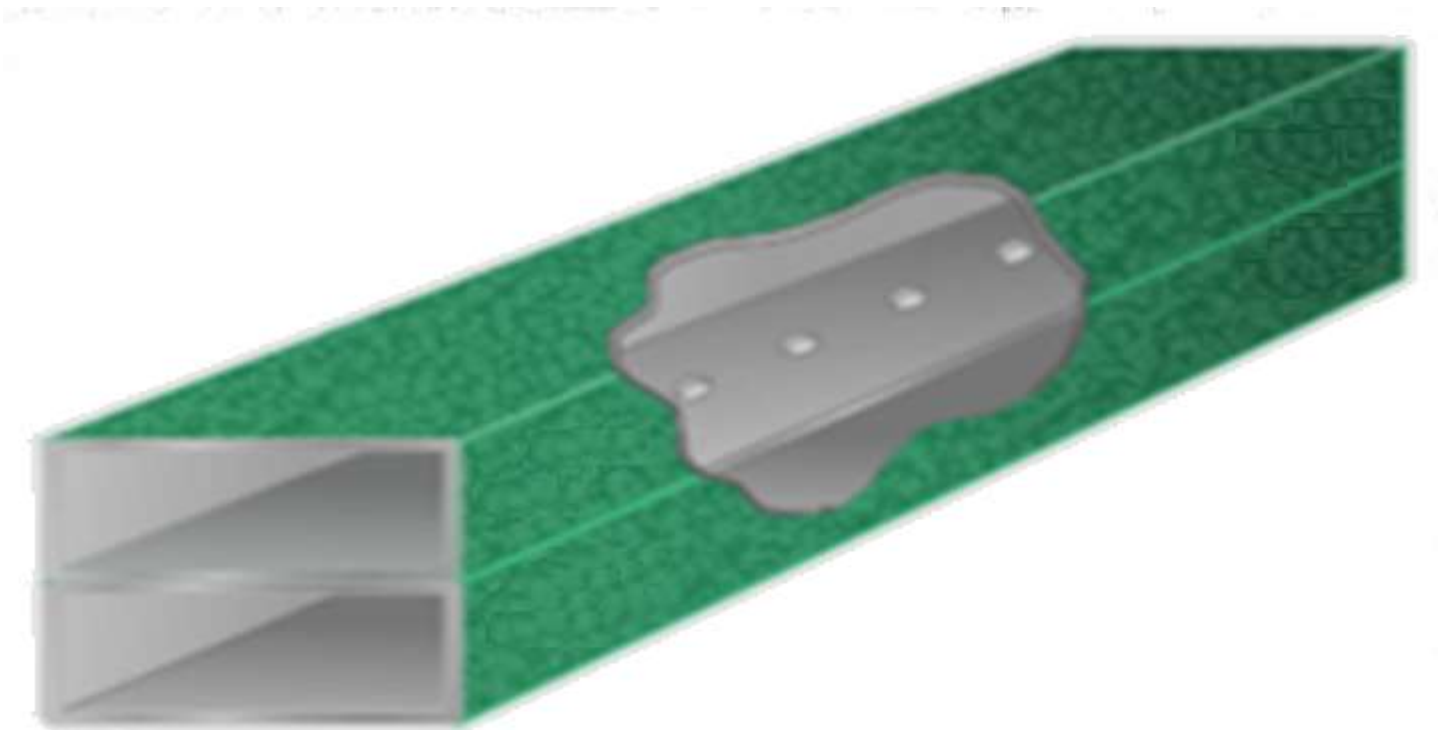
Branchline coupler



Directional Coupler in microstrip

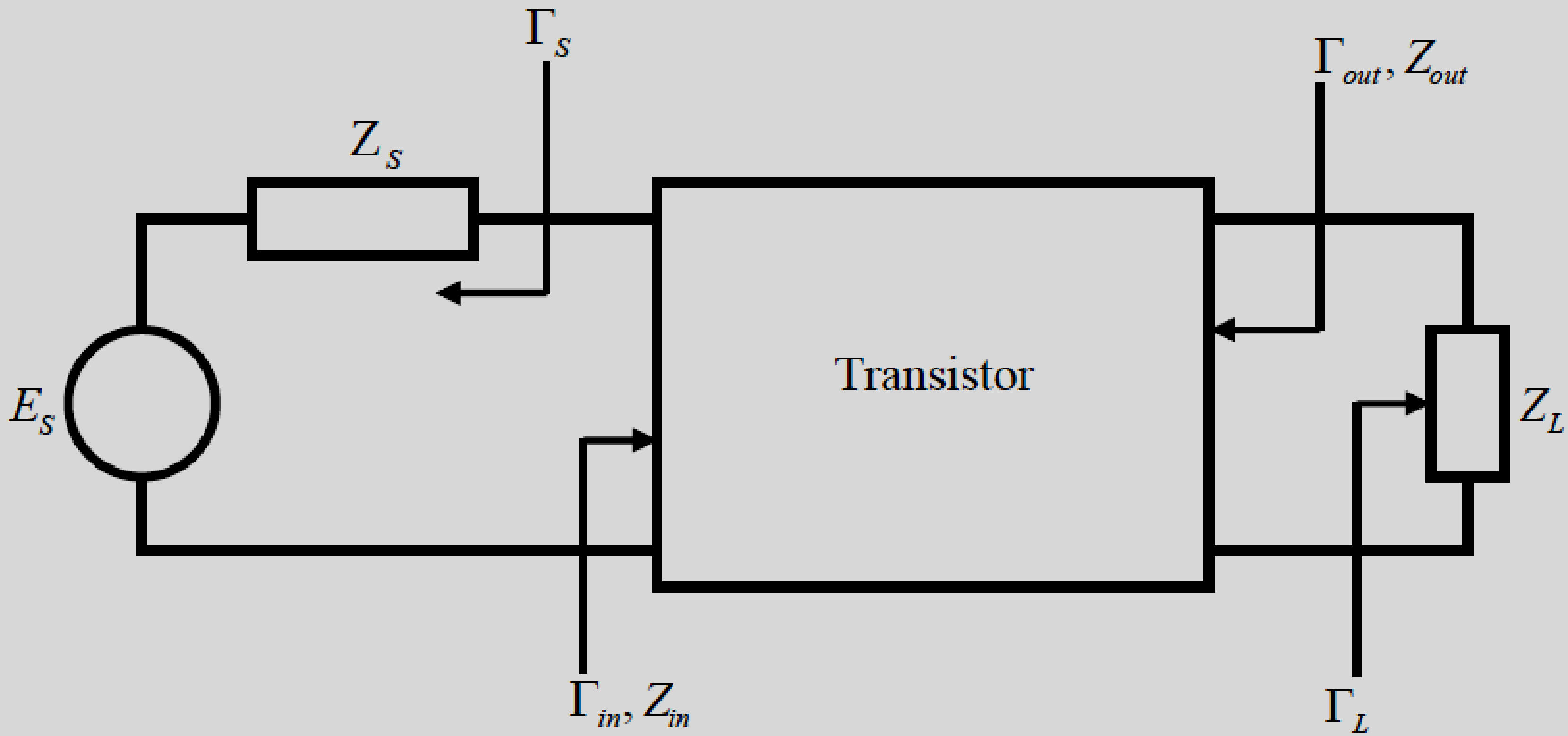


A multi-hole waveguide coupler



S-parameter based amplifier design

Stability



Oscillation is possible if either input or output port has a negative resistance. If a net negative real part exists, that is if $\mathrm{Re}\{Z_S + Z_{in}\}$ or $\mathrm{Re}\{Z_{out} + Z_L\} < 0$, the transient response will grow and oscillation will occur.

Conditional and unconditional stability

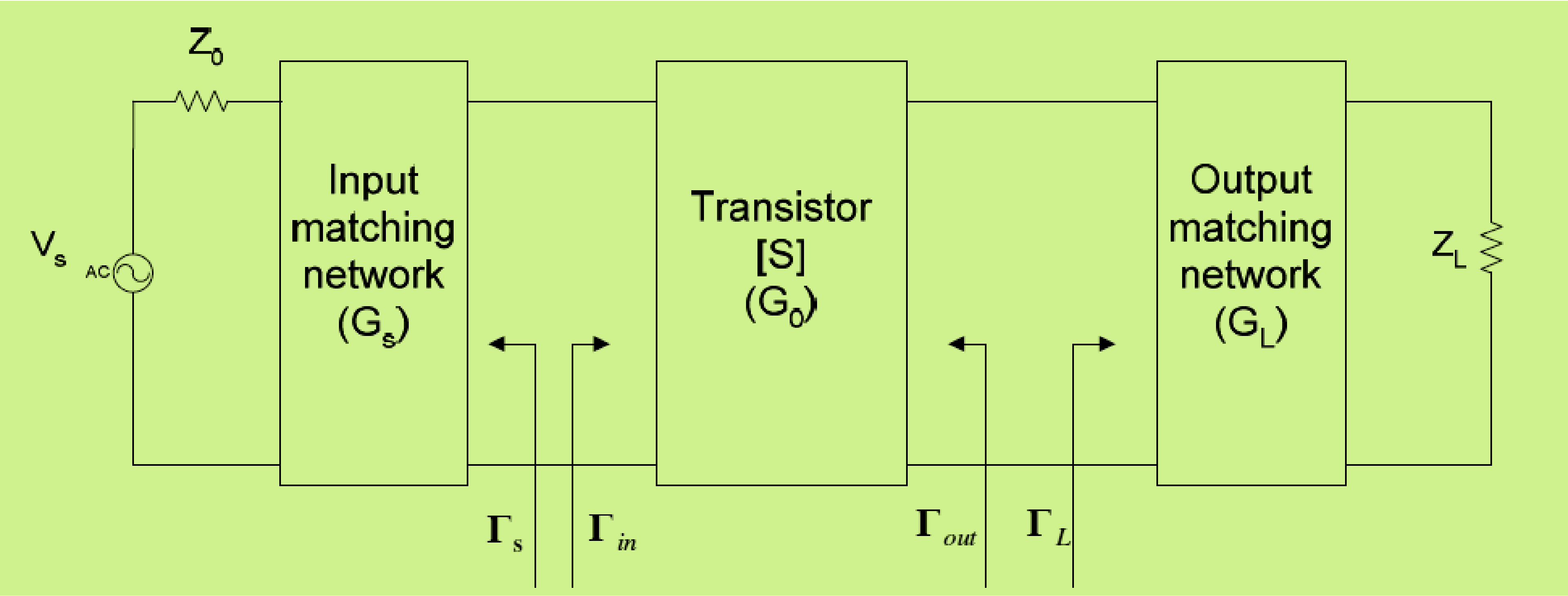
•A network is **conditionally stable if** Re{Zin} and Re{Zout} are >0 for **some positive real Zs and Zl** at a specific frequency.

•A network is **unconditionally stable if** Re{Zin} and Re{Zout} are >0 for **all positive real Zs and Zl** at a specific frequency.

•The above conditions will have to be investigated at many frequencies to ensure broadband stability.

•Temperature changes, drifting of S-parameters of the transistor, piece to piece variations, board-to-board variations of the matching network impedances also need to be considered.

Amplifier two-port network



$$\Gamma_{IN} = \left(\frac{S_{21}\Gamma_L S_{12}}{(1 - S_{22}\Gamma_L)} + S_{11} \right)$$

$$\Gamma_{OUT} = \left(\frac{S_{21}\Gamma_s S_{12}}{(1 - S_{11}\Gamma_s)} + S_{22} \right)$$

Load stability circles

Γ_L or "load plane"

Load stability circles locate the boundary (values of Γ_L)

between $|\Gamma_{in}| < 1$ (stable) and $|\Gamma_{in}| > 1$ (unstable).

Setting $|\Gamma_{in}| = \left| S_{11} + \dfrac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} \right| = 1$ and solving for Γ_L:

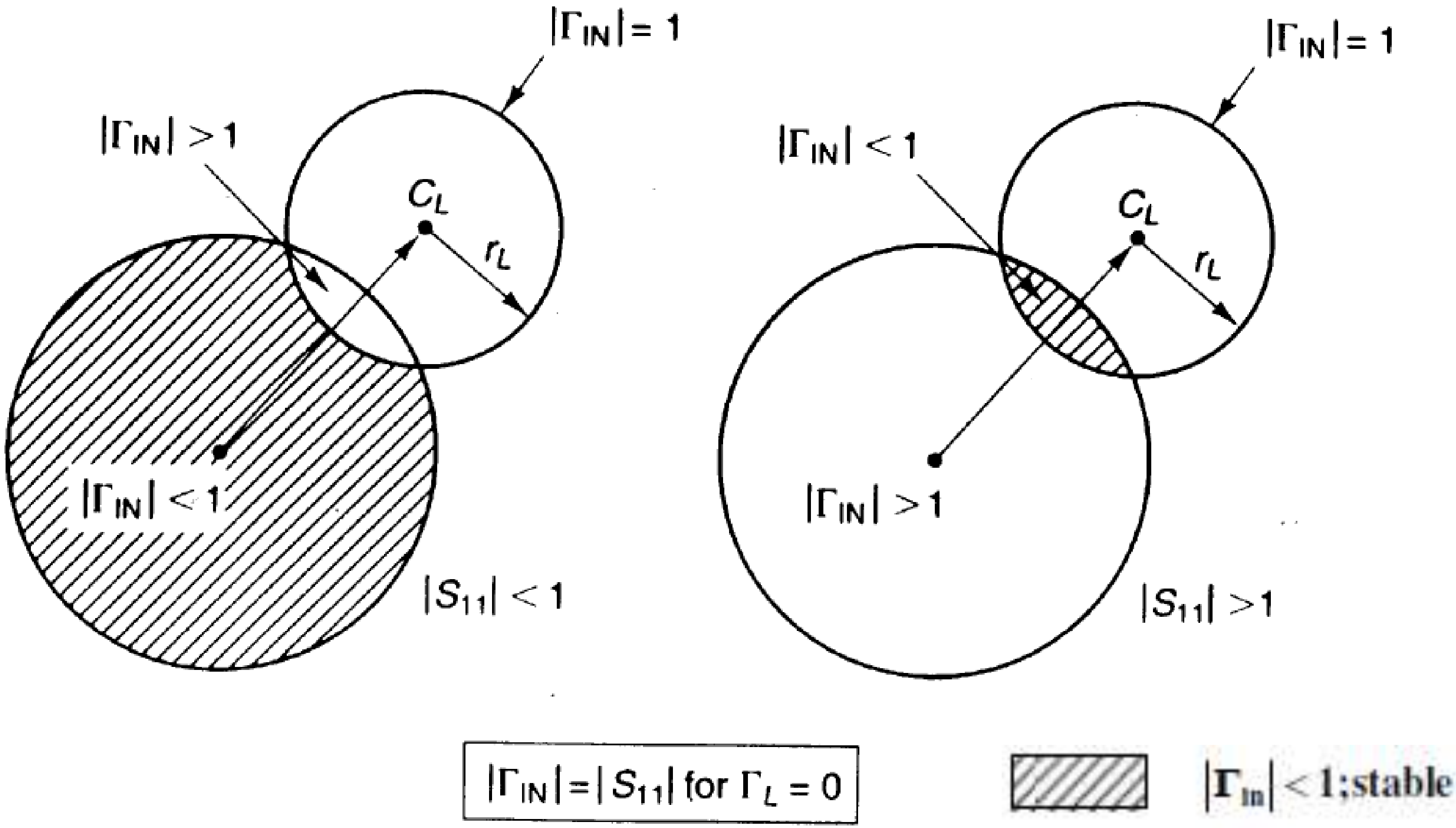
Solution lies on a circle of

center $c_L = \left| \dfrac{\left(S_{22} - \Delta S_{11}^*\right)^*}{|S_{22}|^2 - |\Delta|^2} \right|$

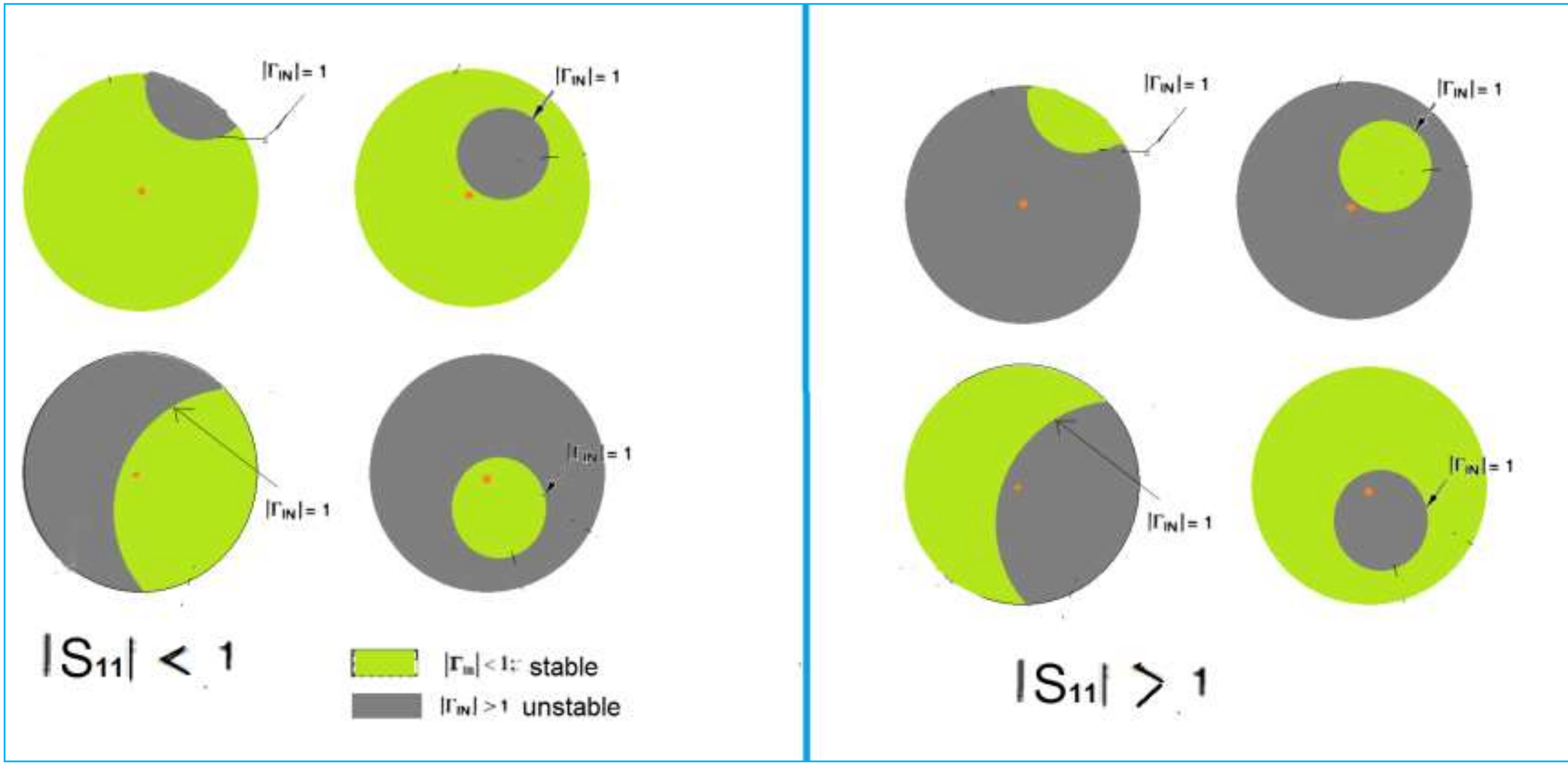
radius $r_L = \left| \dfrac{S_{12}S_{21}}{|S_{22}|^2 - |\Delta|^2} \right|$

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

Load stability circles



Stable and unstable region



Source stability circles

Γ_s *or " source plane"*

Source stability circles locate the boundary (values of Γ_s)

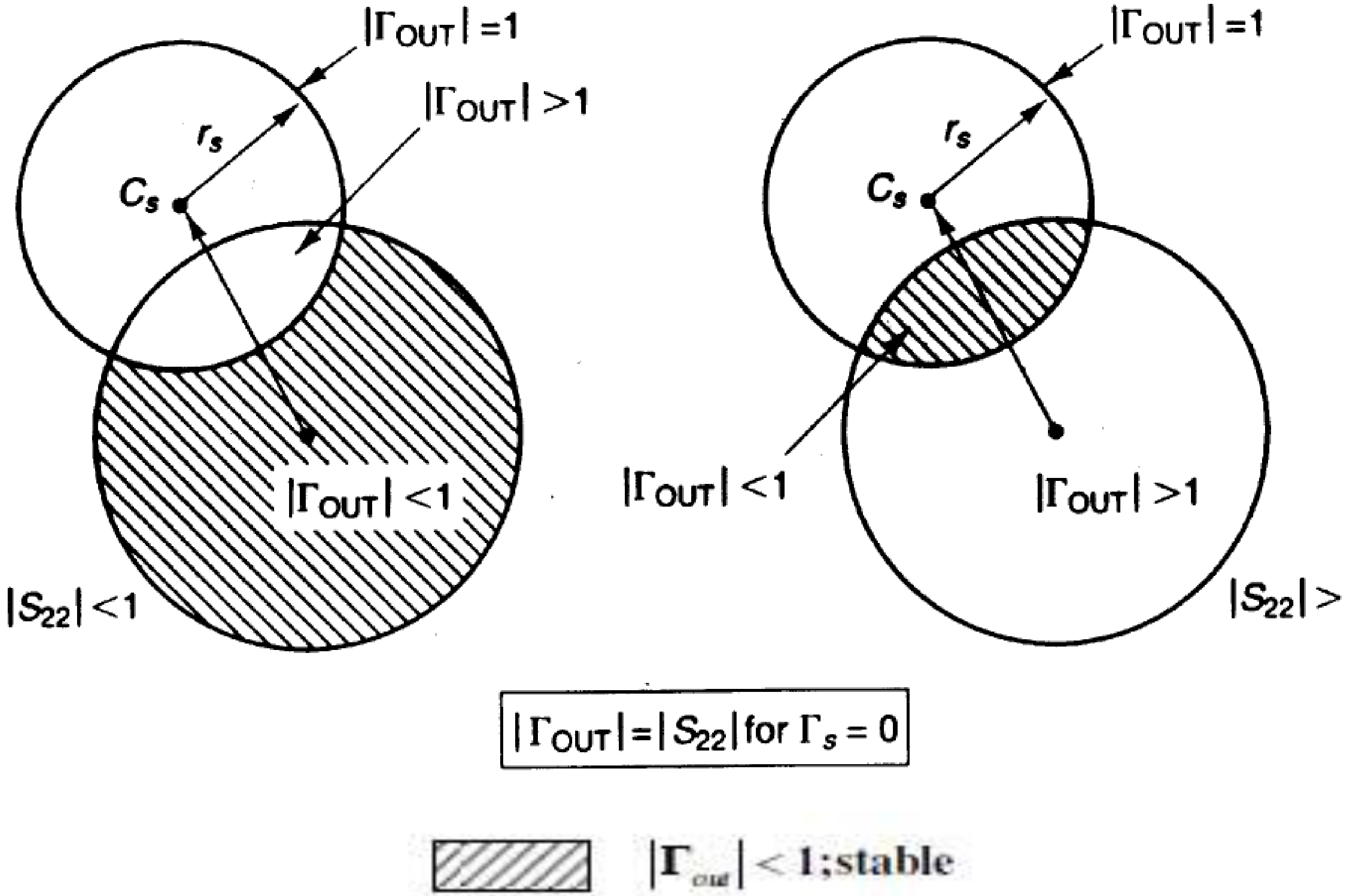
between $|\Gamma_{out}| < 1$ and $|\Gamma_{out}| > 1$
(stable) (unstable)

Setting $|\Gamma_{out}| = \left| S_{22} + \dfrac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S} \right| = 1$ and solving for Γ_S:

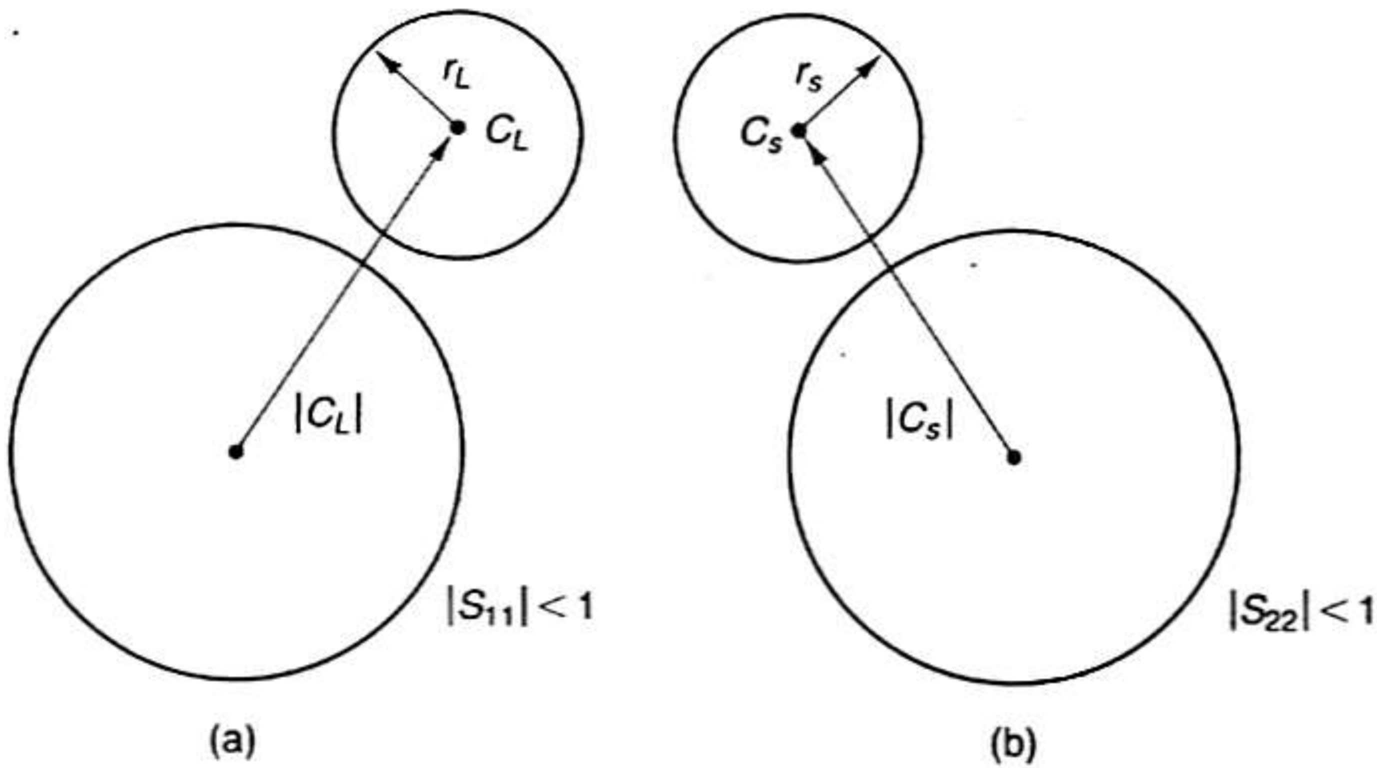
The source stability circles on the Γ_s plane are defined by:

$$r_S = \left| \frac{S_{12}S_{21}}{|S_{11}|^2 - |\Delta|^2} \right| \qquad c_S = \left| \frac{(S_{11} - \Delta S_{22}^*)^*}{|S_{11}|^2 - |\Delta|^2} \right|$$

Source stability circles



Load and source stability circles for unconditional stability



Conditions for unconditional stability: (a) Γ_L plane; (b) Γ_s plane.

Stability circle outside of Smith chart.
No value of $|\Gamma_L| < 1$ can cause instability.

Criteria for unconditional stability

$$\text{Stability Factor} \quad k = \frac{1 + \overbrace{\left|S_{11}S_{22} - S_{12}S_{21}\right|}^{\Delta}{}^{2} - \left|S_{11}\right|^{2} - \left|S_{22}\right|^{2}}{2\left|S_{12}\right|\left|S_{21}\right|} > 1$$

<u>**and**</u>

$$|\Delta| = \left|S_{11}S_{22} - S_{12}S_{21}\right| = \det S < 1$$

will guarantee <u>unconditional stability</u>.

Conjugate matching of input and output (maximum gain)

$$\Gamma_{IN} = \left(\frac{S_{21}\Gamma_L S_{12}}{(1 - S_{22}\Gamma_L)} + S_{11} \right)$$

$$\Gamma_{OUT} = \left(\frac{S_{21}\Gamma_s S_{12}}{(1 - S_{11}\Gamma_s)} + S_{22} \right)$$

From input conjugate matching

$$\Gamma_s^* = \Gamma_{in}$$

From output conjugate matching

$$\Gamma_L^* = \Gamma_{out}$$

Solutions for Γ_S and Γ_L

$$\text{solution for } \Gamma_S = \frac{B_1 \pm \sqrt{B_1^2 - 4|C_1|^2}}{2C_1}$$

$$\text{solution for } \Gamma_L = \frac{B_2 \pm \sqrt{B_2^2 - 4|C_2|^2}}{2C_2}$$

where

$$B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2$$

$$B_2 = 1 + |S_{22}|^2 - |S_{11}|^2 - |\Delta|^2$$

$$C_1 = S_{11} - \Delta S_{22}^*$$

$$C_2 = S_{22} - \Delta S_{11}^*$$

Gain circles

Gain circles show where Γ_S or Γ_L must be to achieve certain gain from the device

$$G_{TU} = G_S G_0 G_L$$

G_0 remains constant $= |S_{21}|^2$

G_S, G_L depend on Γ_S, Γ_L respectively

Drawing constant gain circles

Determine gain steps of interest and calculate normalized gain factor $g_i = \dfrac{G_i}{G_{i,\max}}$

where $0 \le g_i \le 1$. $\quad i = S$ or L.

example: Suppose $G_{S,\max} = 3.3dB$ and you want to draw gain circle for 2 dB.

$$3.3dB \Rightarrow 2.14 \qquad g_S = \frac{1.58}{2.14} = 0.743$$
$$2\,dB \Rightarrow 1.58$$

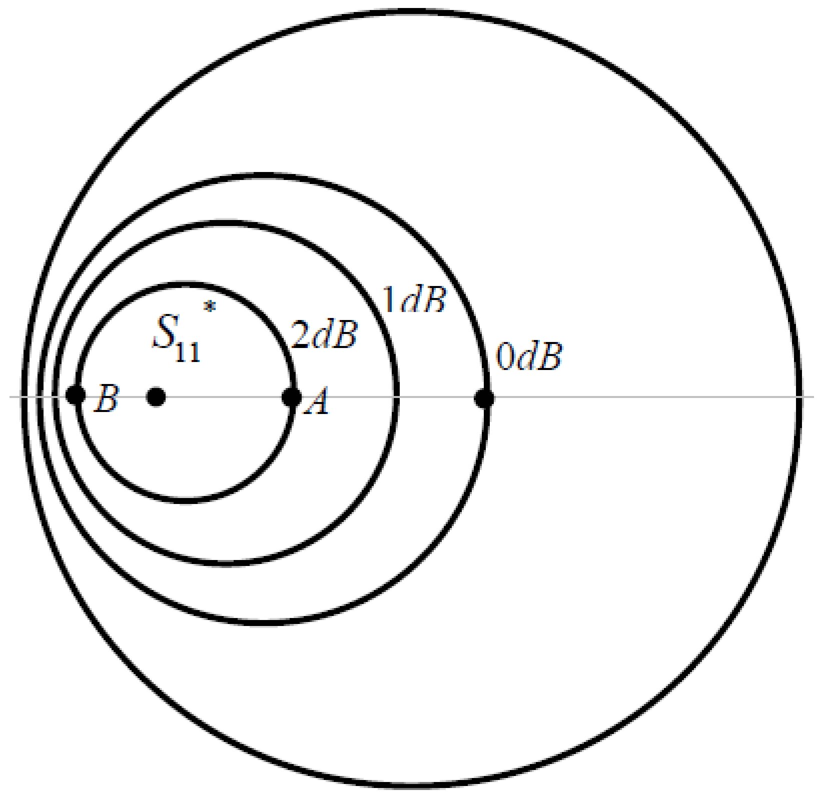
Calculate $C_g = \dfrac{g_S S_{11}^*}{1 - |S_{11}|^2 (1 - g_S)}$

Calculate $r_{gs} = \dfrac{\sqrt{1 - g_S}\left(1 - |S_{11}|^2\right)}{1 - |S_{11}|^2 (1 - g_S)}$

Low noise ampllifier

- Noise figure depends primarily on the input match.
- The input is often mismatched to obtain the best noise figure at the expense of gain.
- The output is normally matched for maximum gain under the mismatched input conditions.

Choosing Γ_s



Where should you choose Γ_S for say $2dB$ gain?

Noise circles

To calculate noise circles:

Define:

$$N_i = \frac{F_i - F_{\min}}{4r_n}\left|1+\Gamma_{opt}\right|^2 = noise\ figure\ parameter$$

Rearrange the equation so that

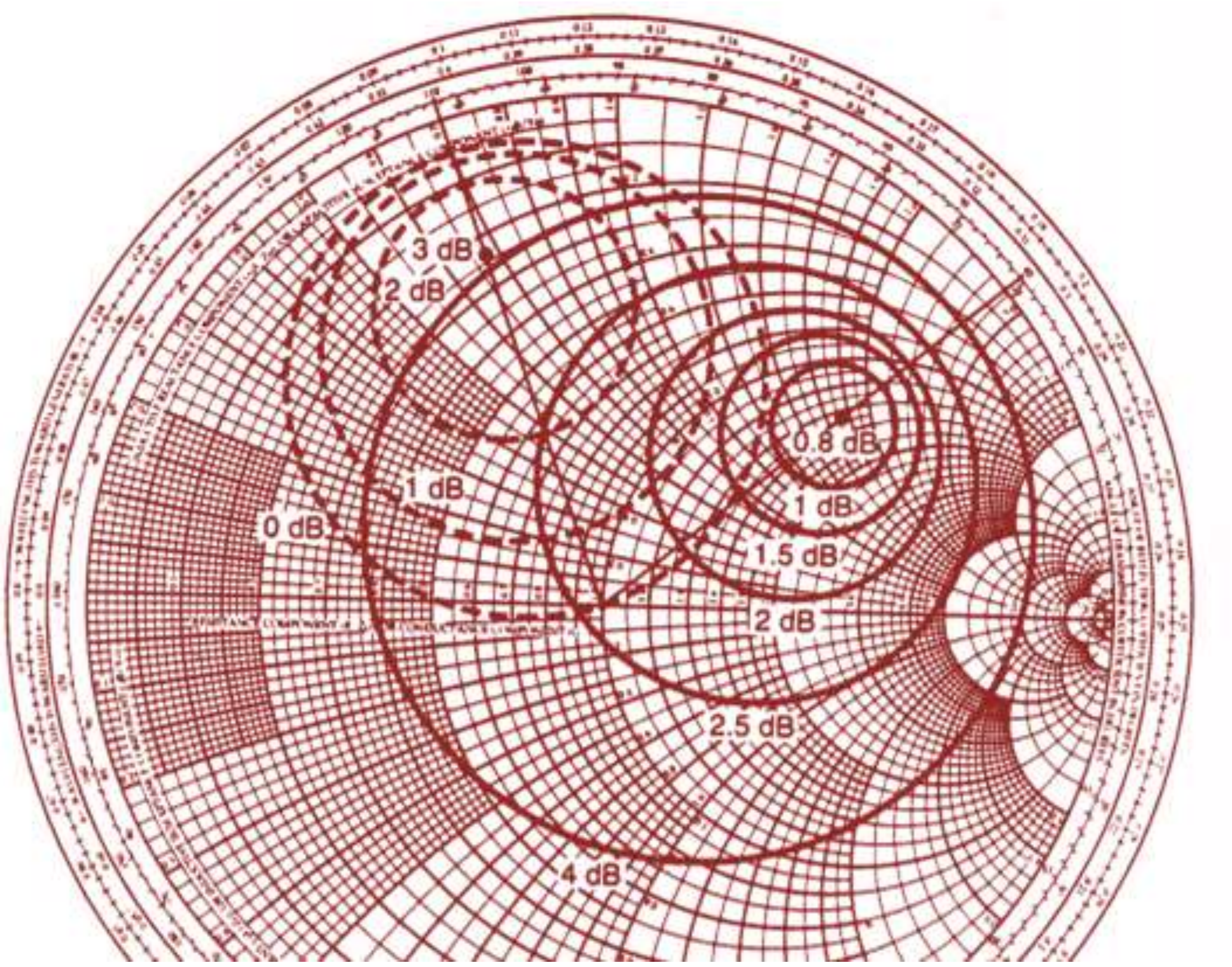
$$\left|\Gamma_S - C_{Fi}\right| = r_{Fi}$$

where:

$$C_{F_i} = \frac{\Gamma_{opt}}{1+N_i} \quad \text{(center)}$$

$$r_{Fi} = \frac{1}{N_i+1}\sqrt{N_i^2(1-|\Gamma_{opt}|^2)} \quad \text{(radius)}$$

Noise figure circles and constant Gs circles



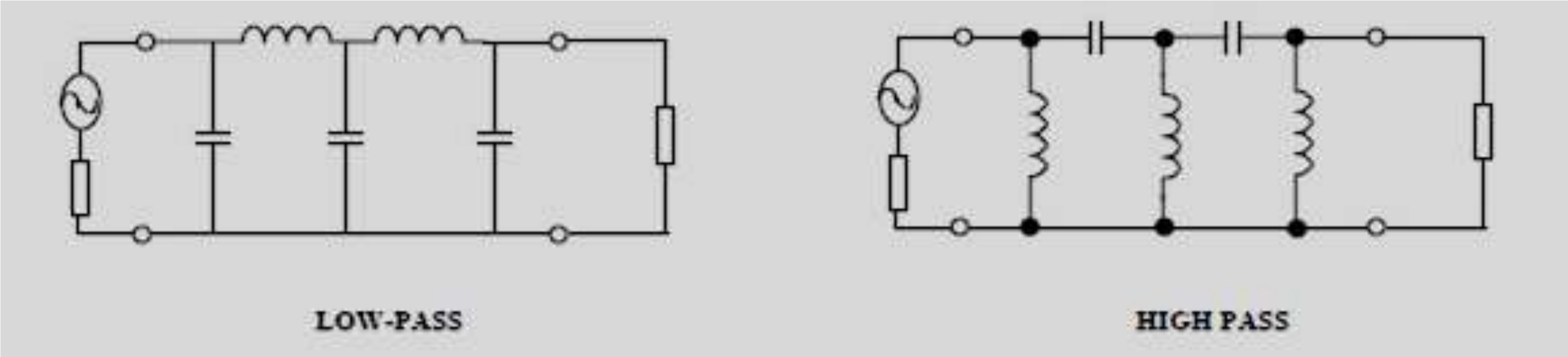
Common causes for LNA instability

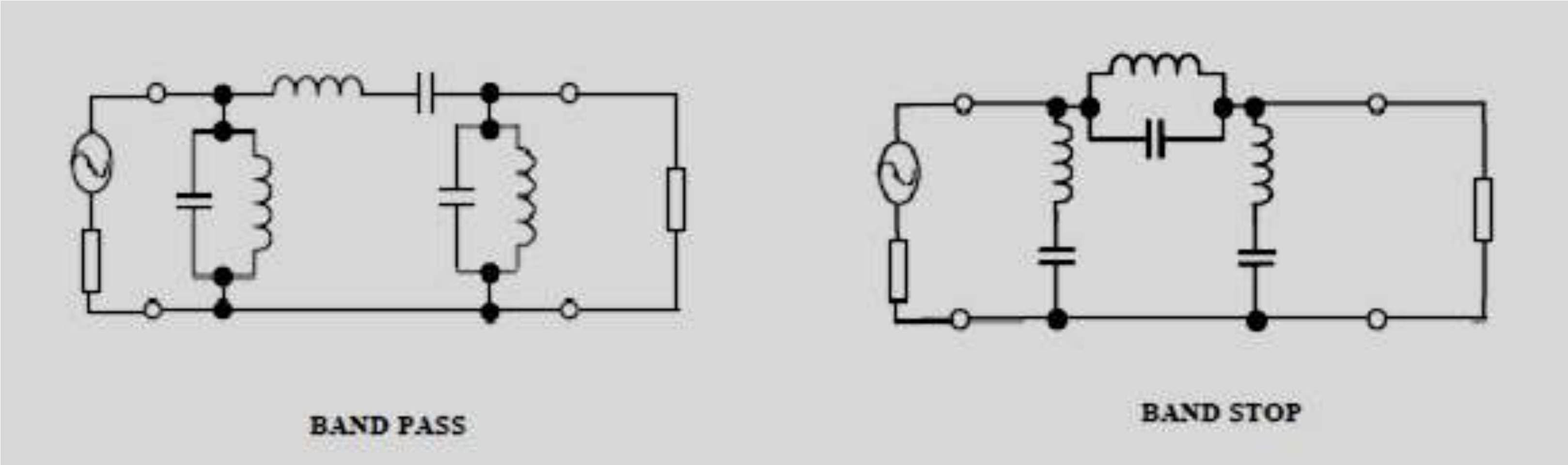
- Insufficient RF decoupling between supply lines of the amplifier bias
- Parasitic inductance in GND connections
- Excess in-band and/or out-of-band gain
- Electro-Magnetic coupling and feedback

Methods of circuit stabilization

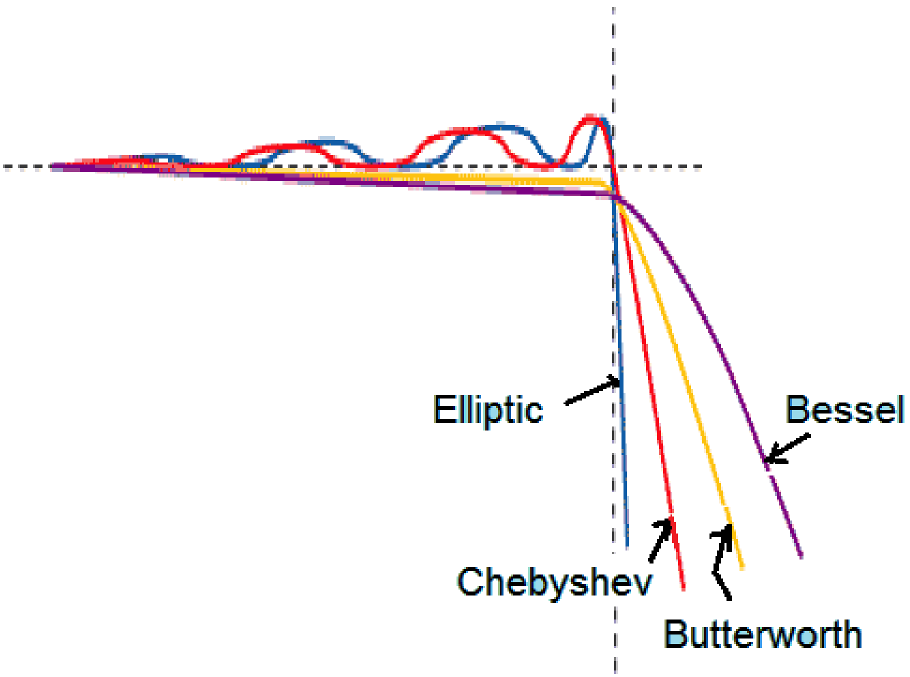
- **Resistive loading of the input** improves the stability of the circuit but also degrades the noise of the LNA.
- **Output resistive loading** should be carefully used because it effects are lower gain and lower P1dB point (thus IP3 point).
- **Collector to base resistor-inductor-capacitor (RLC) feedback** lowers the gain at lower frequencies and hence can improve the stability of the circuit. ·
- **Filter matching** at the output of the transistor will decrease the gain at a specific frequency. This method is frequently used for eliminating gain at high frequencies. Short circuit quarter wave lines or simple capacitors with the same resonant frequency as the frequency of oscillation (or excessive gain) can be used.
- **Emitter feedback inductor** can make the circuit more stable at higher frequencies. But if the source inductance is increased, the K-factor at higher frequencies eventually falls bellow 1. This limits the amount of source inductance that can safely be used.

RF filter design





Filter types



Butterworth or maximally flat Filter provides the maximum in band flatness, better group delay performance

Bessel filter provides the best step response

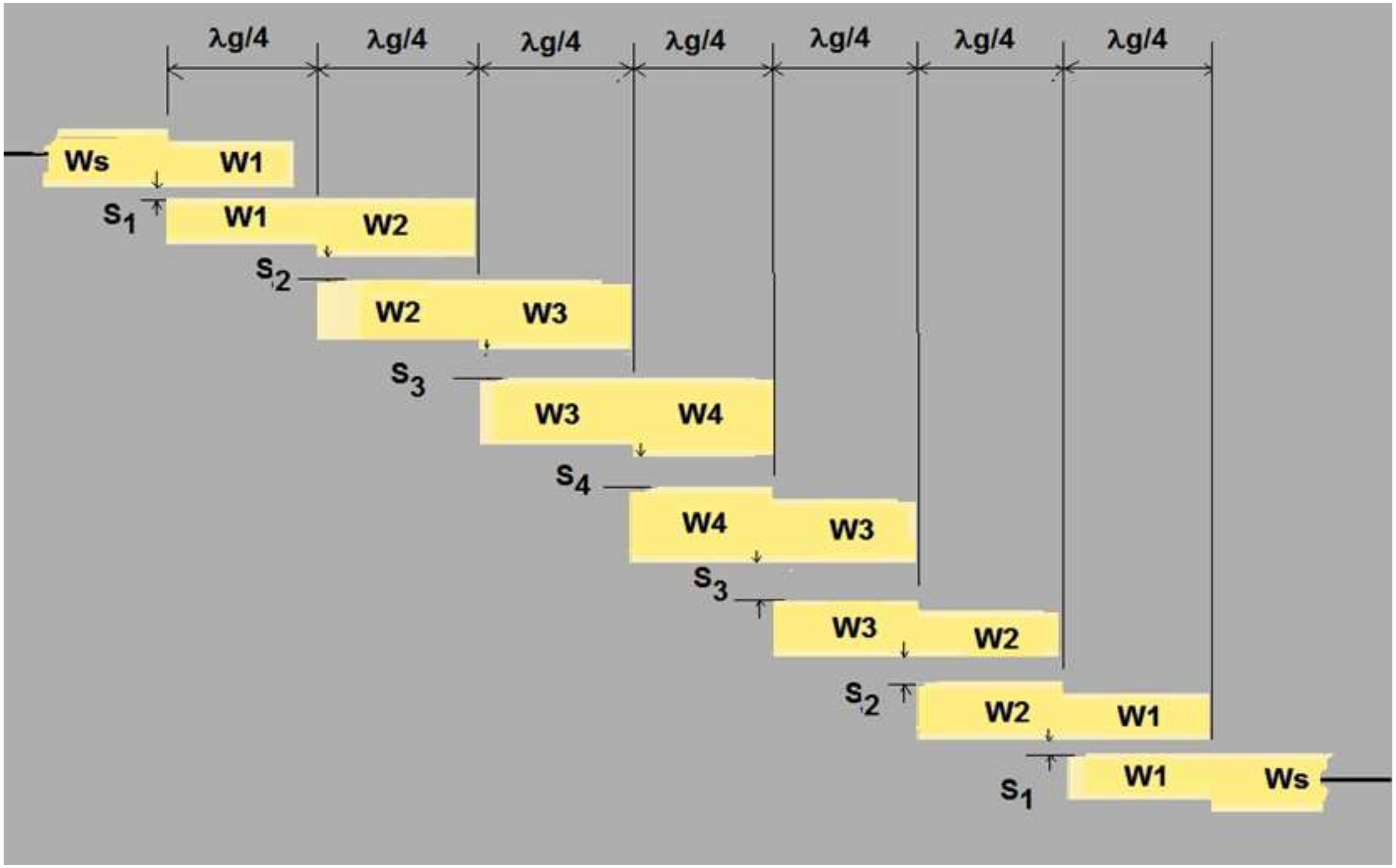
Chebyshev or equal ripple filter provides fast roll off after the cut off frequency is reached

Elliptic filter (a.k.a Cauer filter) has significant levels of in band

Key filter design parameters

- **Pass-band:** the region in which the signal passes through relatively un-attenuated.
- **Stop-band:** the band where the filter provides the required rejection.
- **Cut-off frequency:** the point at which the response has fallen by 3 dB.
- **Ripple band:** The variation within the pass-band
- **Transition band or** "skirt"**:** the region between pass-band and stop-band.

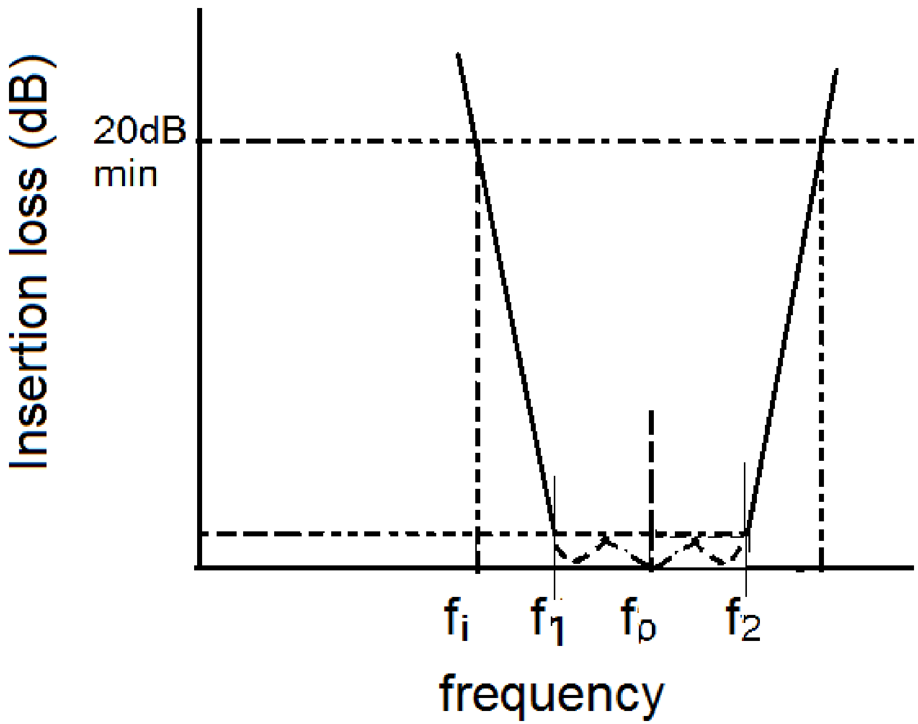
Parallel-coupled (or edge-coupled) BPF design



Parallel-coupled BPF design

- Maximum coupling is obtained between parallel microstrips when the length of coupled region in $\lambda g/4$.
- For resonance each resonator element has to be $\lambda g/2$ in length where λg = midband and average microstrip wavelength

Step 1- Fractional bandwidth and Transformation ratio



Fractional bandwidth

$$\delta = \frac{f_2 - f_1}{f_0}$$

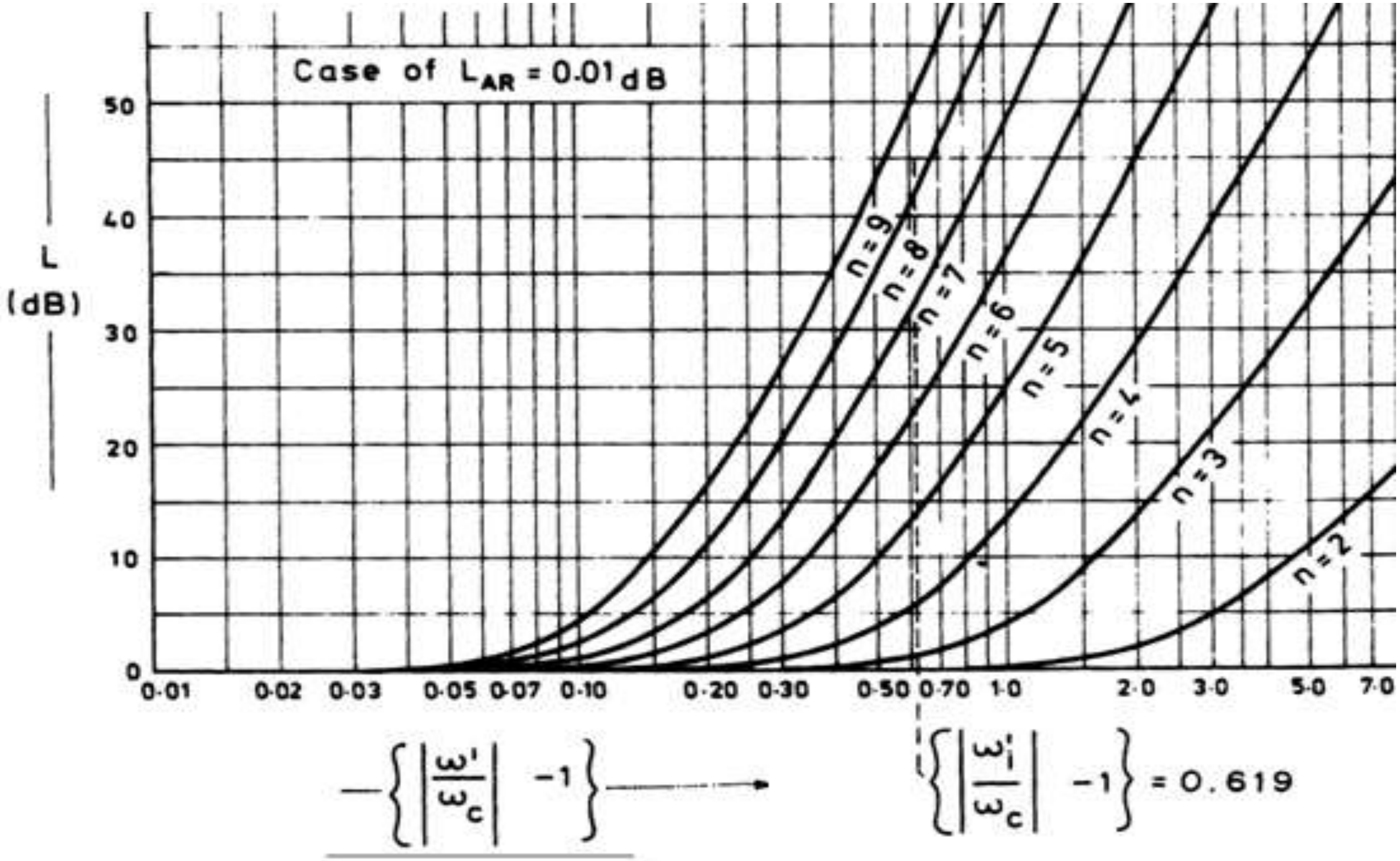
Transformation ratio

$$\frac{\omega_i'}{\omega_c} = \frac{2}{\delta}\left[\frac{f_i - f_0}{f_0}\right]$$

For f_i = 9.65 GHz, f_1= 9.98 GHz, f_0= 10.5 GHz and f_2= 11.03 GHz, δ = 0.1 and ω_i'/ω_c = -1.619

Step 2 -Prototype element values

Insertion loss charateristics for a passband ripple of 0.01dB



A sixth-order Chebyshev design gives 23dB insertion loss .

Normalised Chebyshev element values

Normalized Chebyshev element values, 0.01 dB ripple

Order	C1	L2	C3	L4	C5	L6	C7	L8	C9
2	0.4489	0.4078	0.9085						
3	0.6292	0.9703	0.6292						
4	0.7129	1.2004	1.3213	0.6476	0.9085				
5	0.7563	1.3049	1.5773	1.3049	0.7563				
6	.07814	1.3600	1.6897	1.5350	1.4970	0.7098	0.9085		
7	0.7970	1.3924	1.7481	1.6331	1.7481	1.3924	.07970		
8	0.8073	1.4131	1.7824	1.6833	1.8529	1.6193	1.5555	0.7334	0.9085
9	0.8145	1.4271	1.8044	1.7125	1.9058	1.7125	1.8044	1.4271	0.8145
	L1	C2	L3	C4	L5	C6	L7	C8	L9

Step 3 – Calculating inverter admittances

For the *first* coupling structure:

$$\frac{J_{01}}{Y_0} = \sqrt{\frac{\pi\delta}{2g_0g_1}}$$

For the *intermediate* coupling structures:

$$\left.\frac{J_{j,j+1}}{Y_0}\right|_{j=1 \text{ to } (n-1)} = \frac{\pi\delta}{2\omega_c'\sqrt{g_jg_{j+1}}}$$

For the *final* coupling structure:

$$\frac{J_{n,n+1}}{Y_0} = \sqrt{\frac{\pi\delta}{2g_ng_{n+1}}}$$

where ωc' is the prototype cutoff frequency (equals 1)

Coupled line impedances

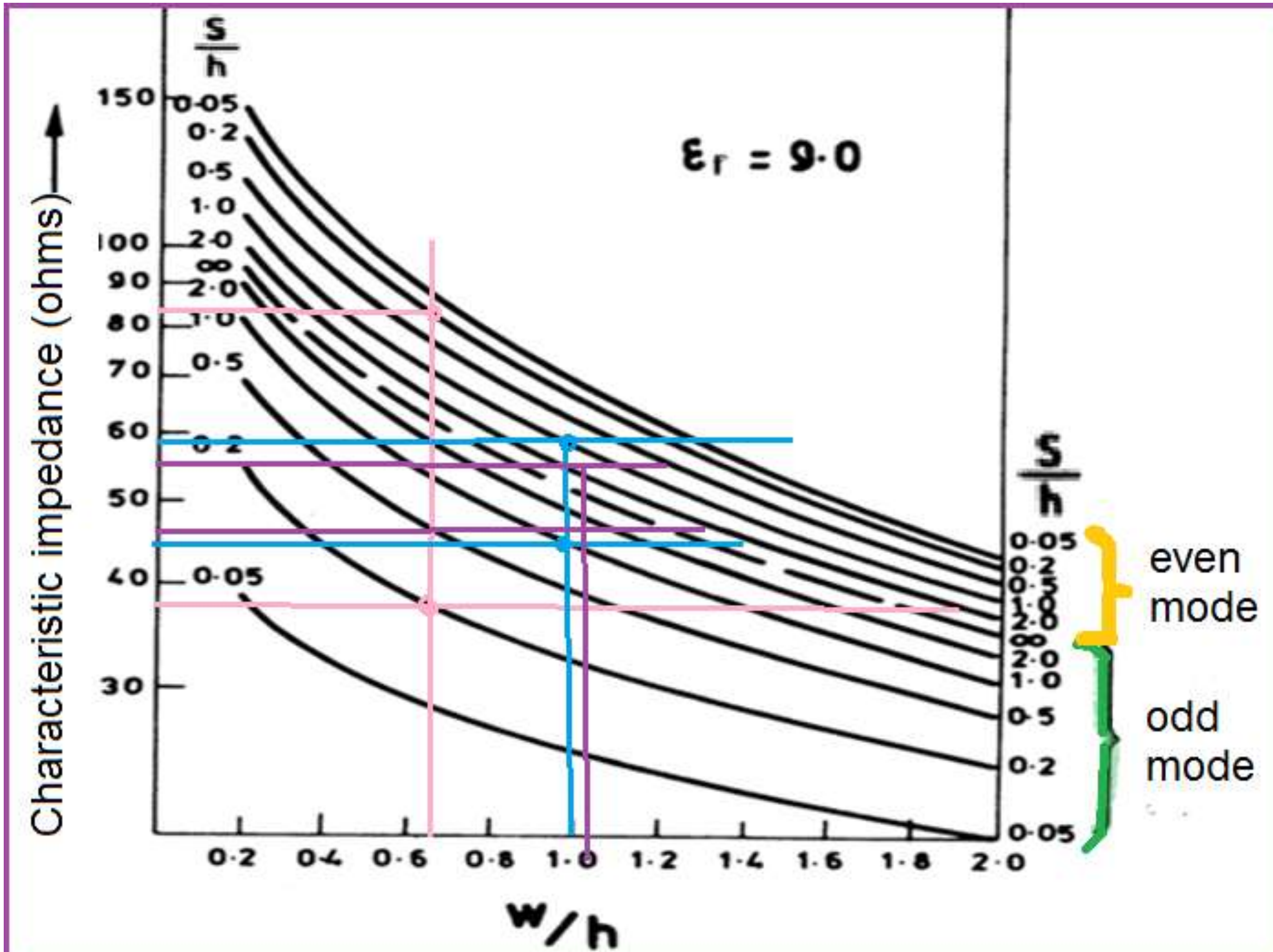
$$(Z_{0e})_{j,j+1} = Z_0(1 + aZ_0 + a^2Z_0^2)$$

$$(Z_{0o})_{j,j+1} = Z_0(1 - aZ_0 + a^2Z_0^2)$$

$$a = J_{j,j+1}$$

j	$J_{i,j+1}/Y_0$	$(Z_{0e})_{j,j+1}(\Omega)$	$(Z_{0o})_{j,j+1}(\Omega)$
0	0.449	82.5	37.6
1	0.1529	58.8	43.5
2	0.1038	55.7	45.3
3	0.0976	55.4	45.6

Step 4 – Obtaining the microstripline dimensions



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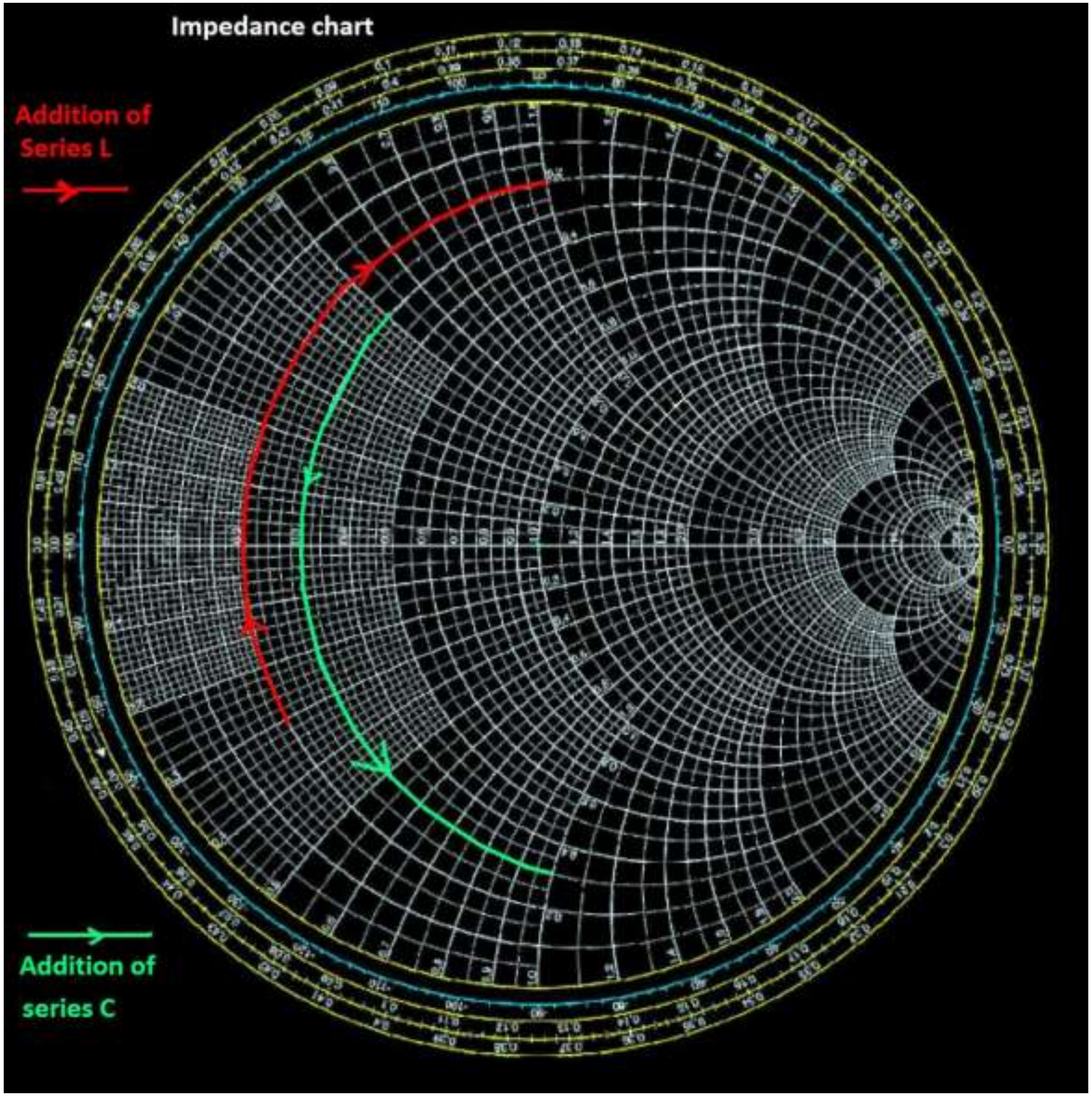
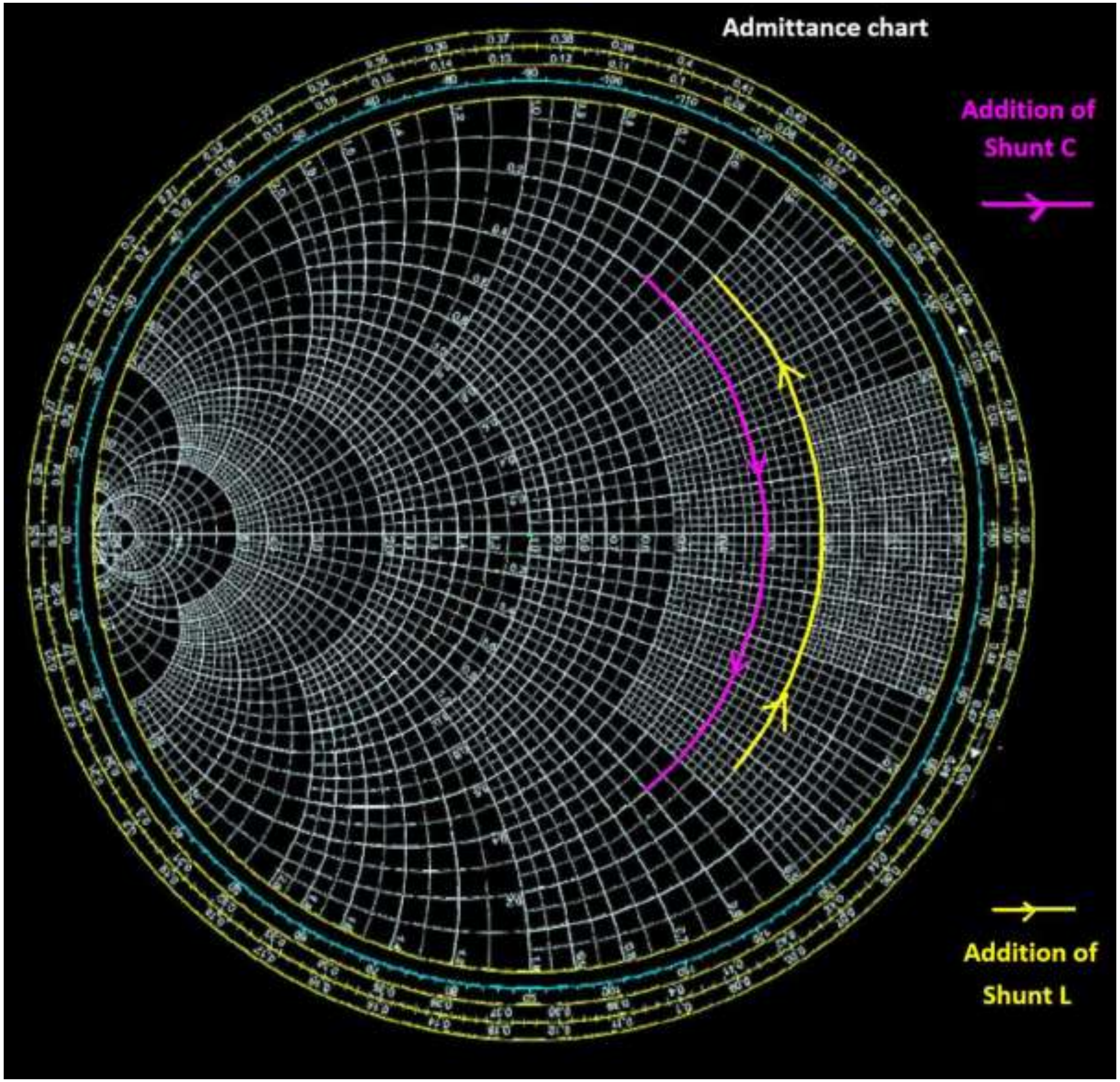
1. Foundations for microstrip circuit design , Second edition by Terry Edwards, John Wiley & Sons Ltd., 1992
2. Handbook of microwave and optical components Volume 1 edited by Kai Chang, John Wiley & Sons Ltd., 1989
3. Handbook of microwave integrated circuits, Reinmut K. Hoffman, Artech House,1987
4. http://www.microwaves101.com/
5. Transmission Line and Lumped Element Quadrature Couplers, Gary Breed, High Frequency Electronics, November 2009

Smith Chart and impedance matching

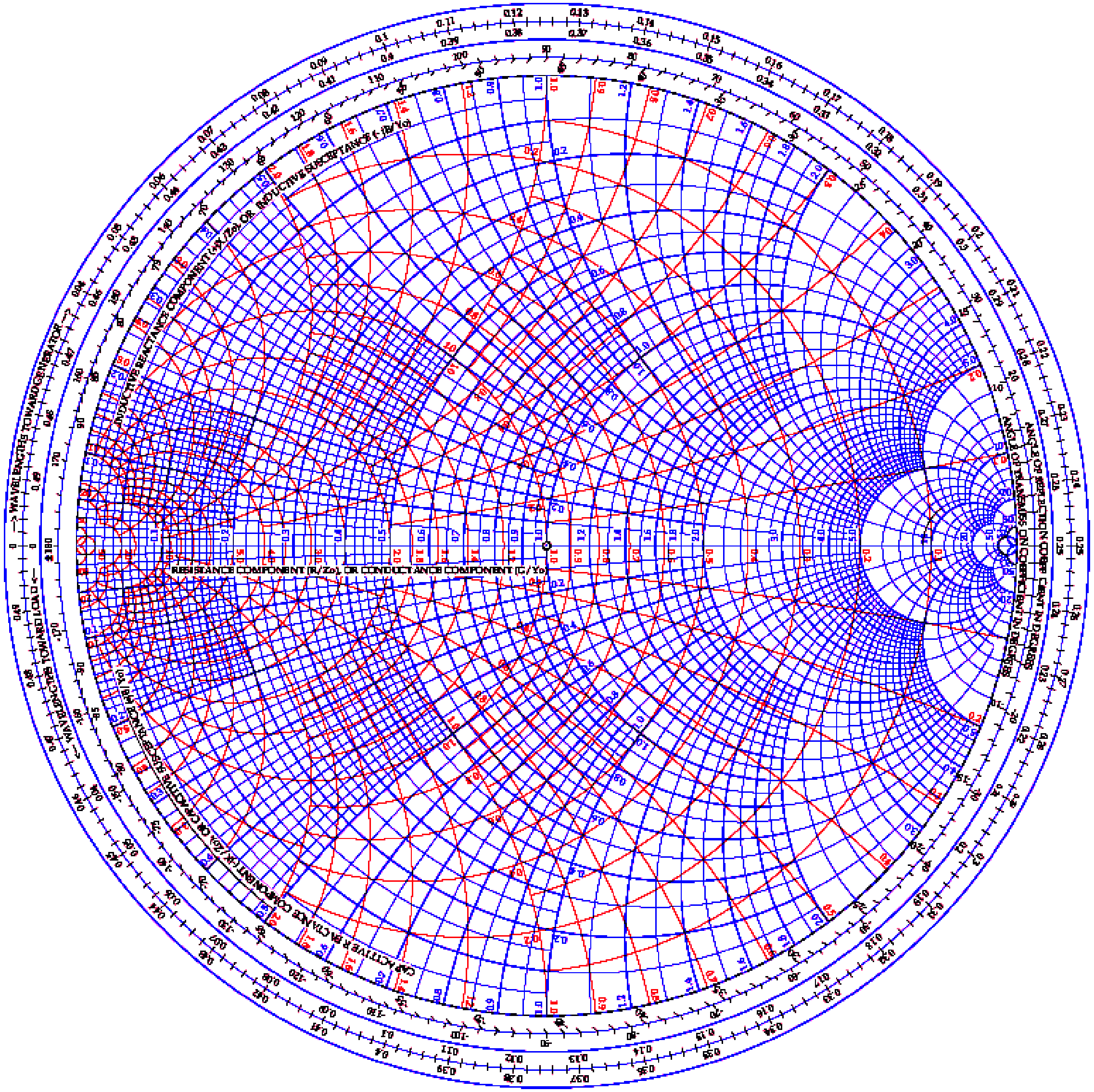
Adding series reactance



Adding shunt susceptance



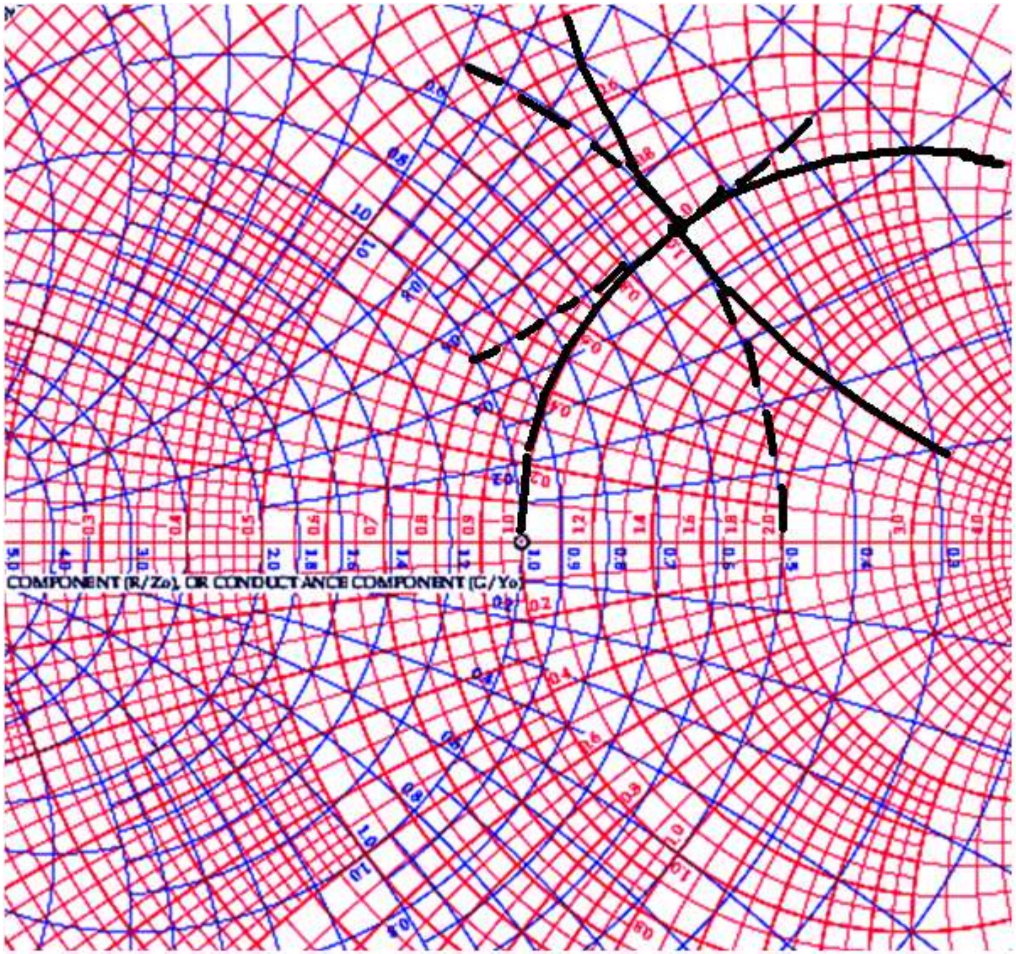
Impedances and admittances superimposed



How to use Smith Chart

- Impedance matching on the Smith Chart
- Two-element
- Three-element
- Multi-element

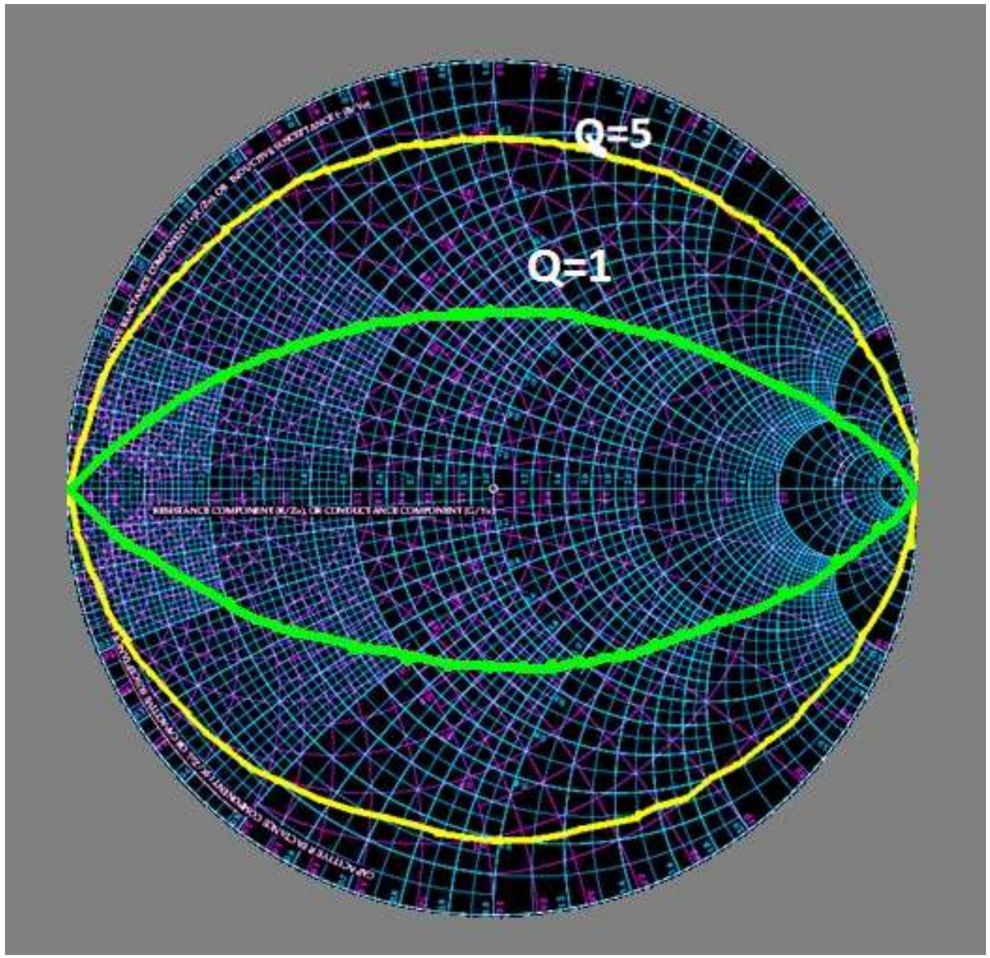
Conversion of impedance to admittance



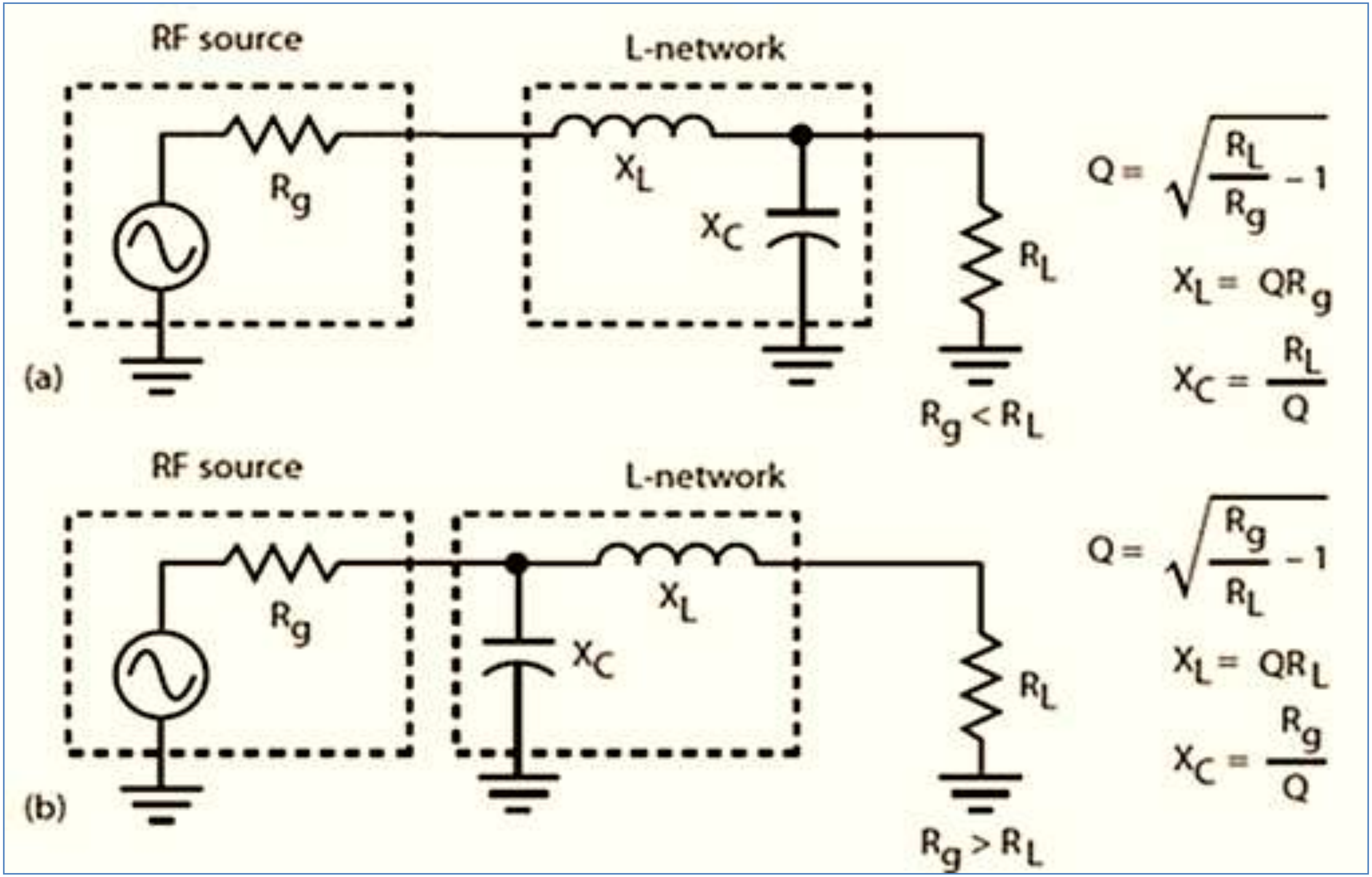
$Z=1+j1$

$Y=0.5-j0.5$

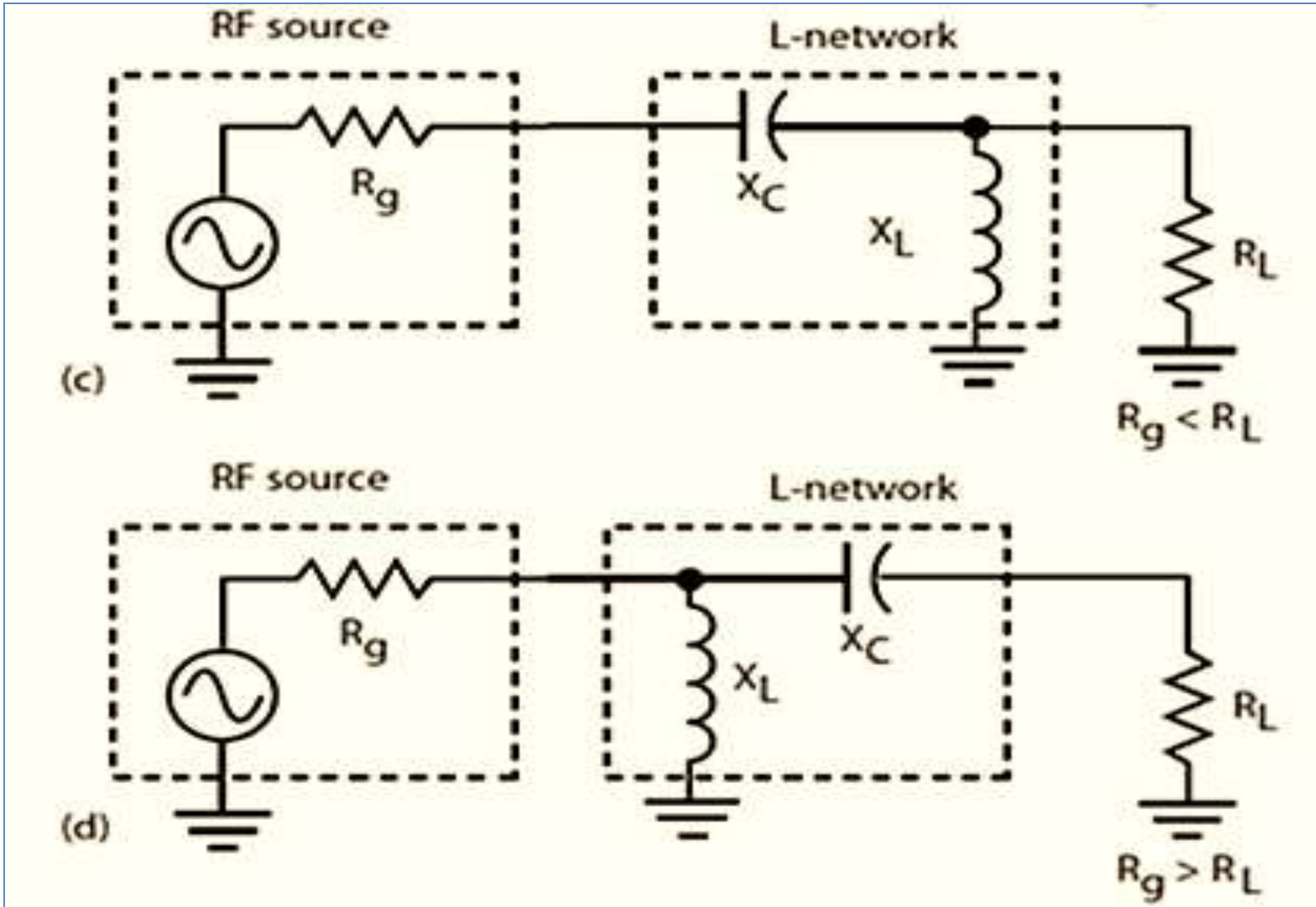
Circuit Q representation



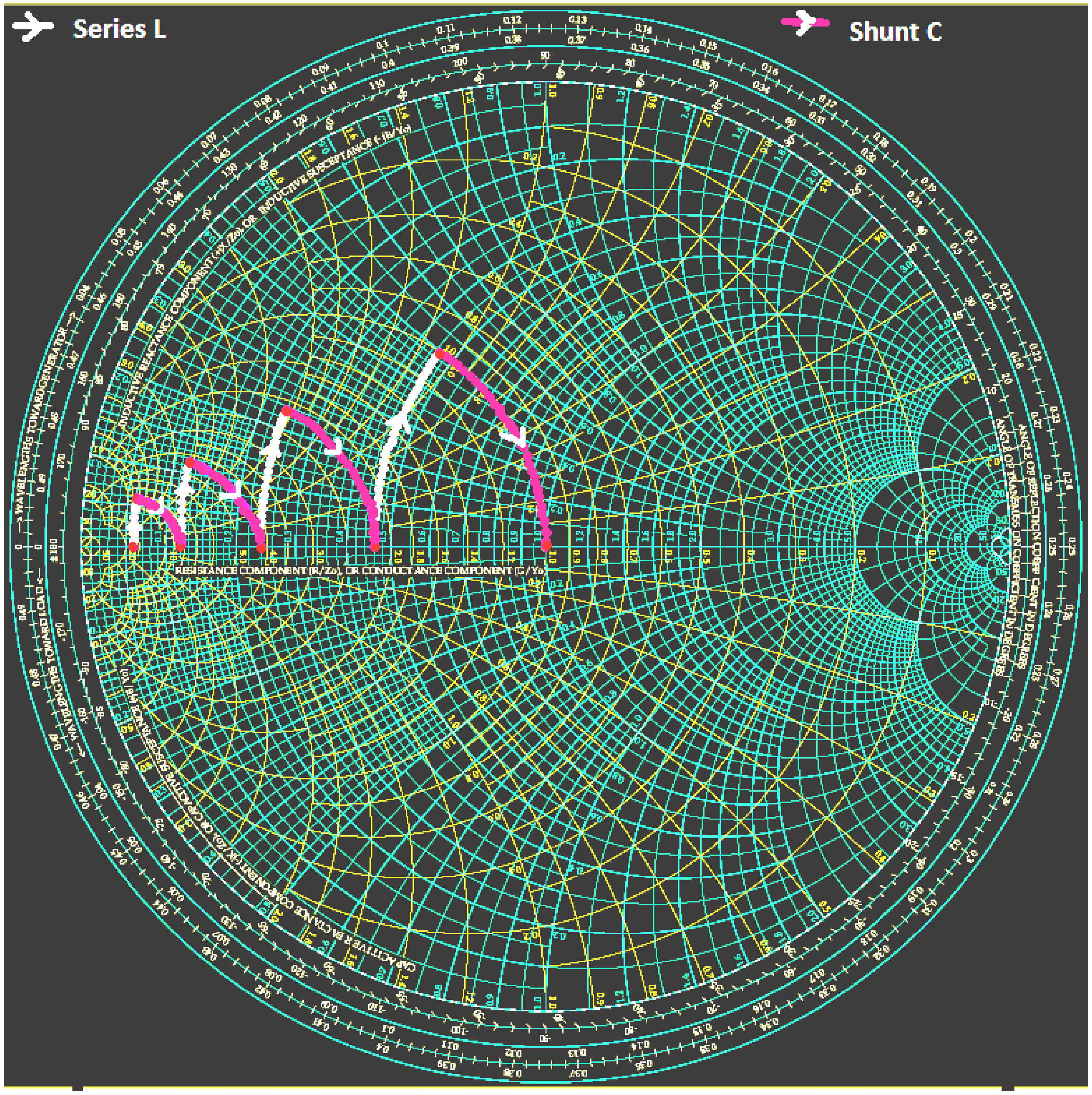
L network - Low pass



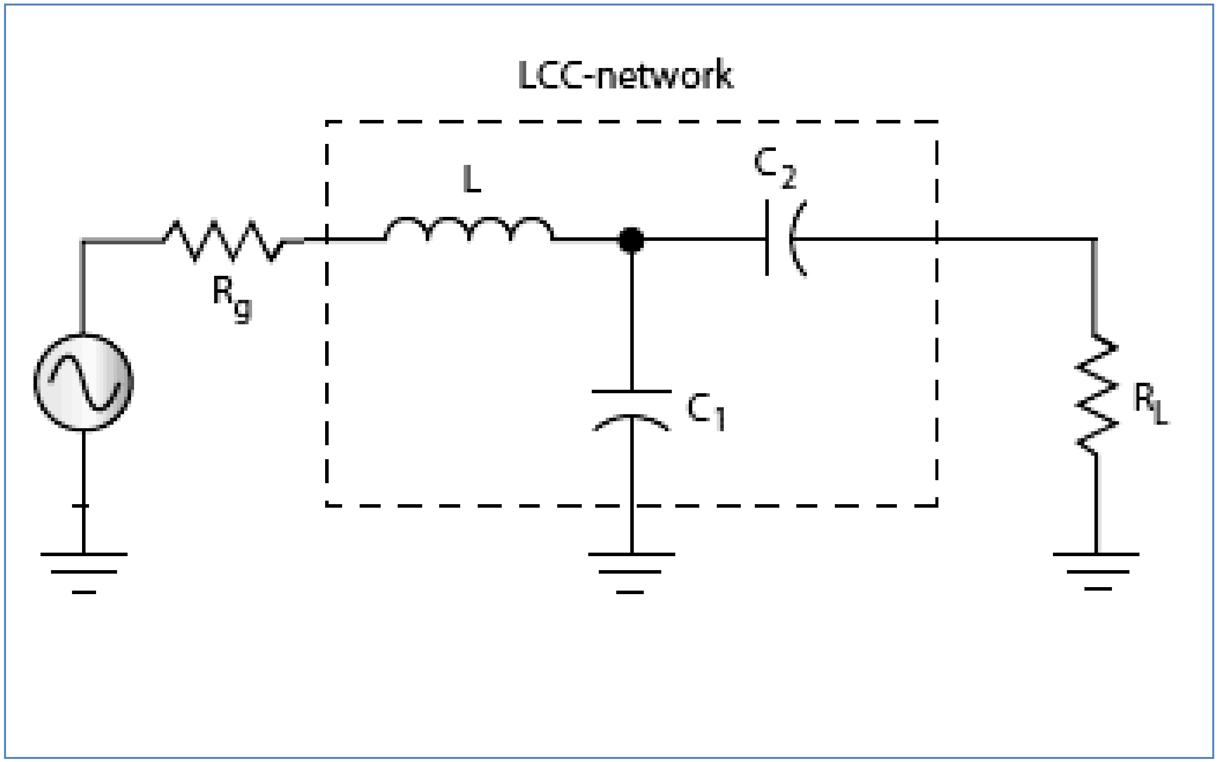
L Network-High pass



Multiple L-sections low pass matching



Popular T-network



T network design

- 1. Select the desired bandwidth and calculate Q.
- 2. Calculate $X_L = QR_g$
- 3. Calculate $X_{C2} = R_L\sqrt{[R_g (Q^2 + 1)/R_L – 1]}$
- 4. Calculate $X_{C1} = R_g (Q^2 + 1)/Q[QR_L/(QR_L – XC^2)]$
- 5. Calculate the inductance $L= X_L/2\pi f$
- 6. Calculate the capacitances $C = 1/2\pi fX_C$

T matching network example

- Assume a source or generator resistance of 10 Ω and a load resistance of 50 Ω. Let Q be 10 and the operating frequency be 315 MHz.

Computing L, C1 and C2

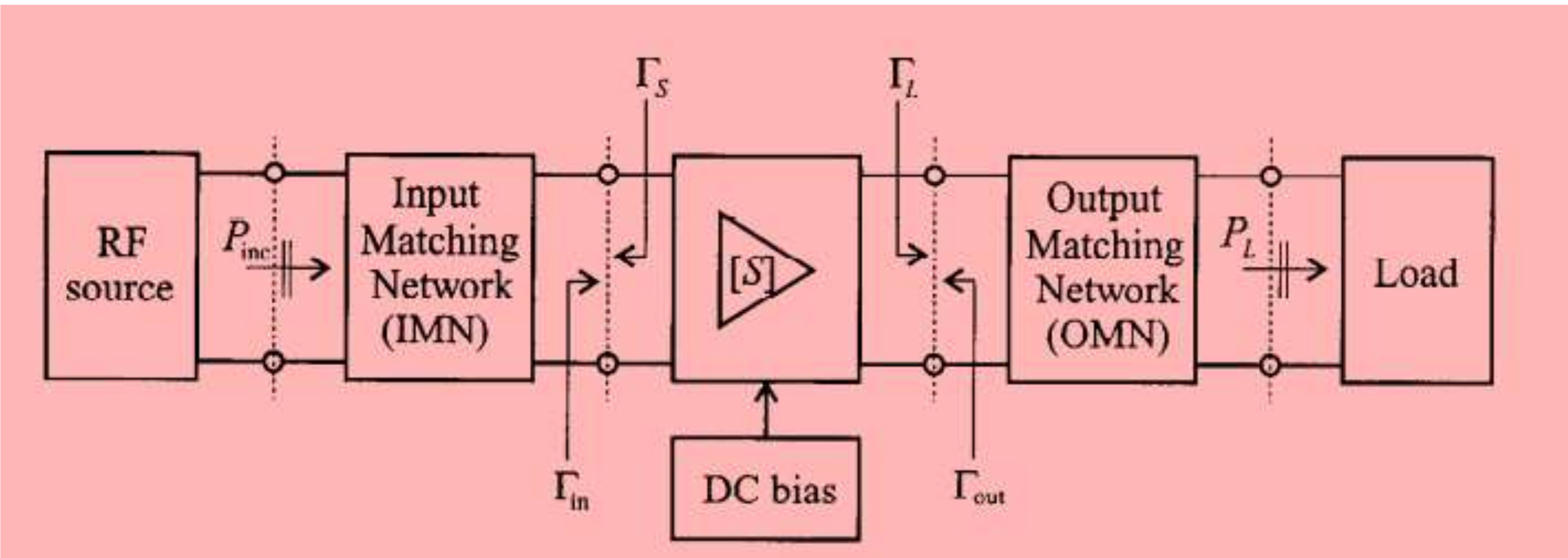
- $X_L = QR_g = 10(10) = 100\ \Omega$
- $L = X_L/2\pi f = 100/2(3.14)(315 \times 106) = 50$ nH
- $X_{C2} = R_L\sqrt{[R_g\ (Q^2 + 1)/R_L - 1]} = 50\sqrt{\{[10(101)/50] - 1\}} = 219\ \Omega$
- $X_{C1} = R_g\ (Q^2 + 1)/Q\ [QR_L/(QR_L - X_{C2})] = 10(101)/10[500/(500 - 219)] = 179\ \Omega$
- $C_2 = 1/2\pi fX_C = 1/2(3.14)(315 \times 10^6)(219) = 2.31$ pF
- $C_1 = 1/2\pi fX_C = 1/2(3.14)(315 \times 10^6)(179) = 2.82$ pF

‘π’ matching network

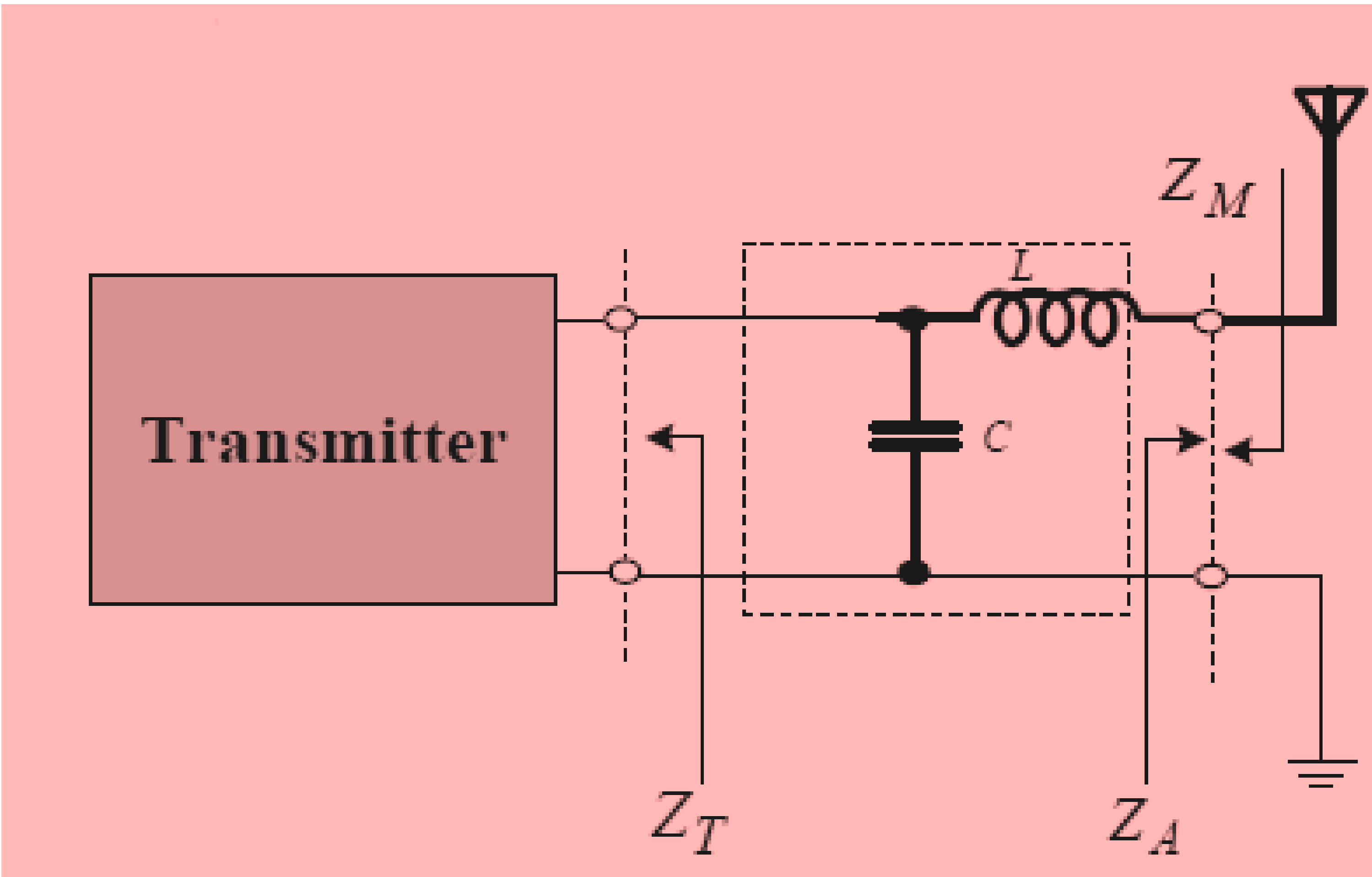
- The π-network matching circuit is used mostly in high- to low-impedance transformations

Tunable impedance matching networks

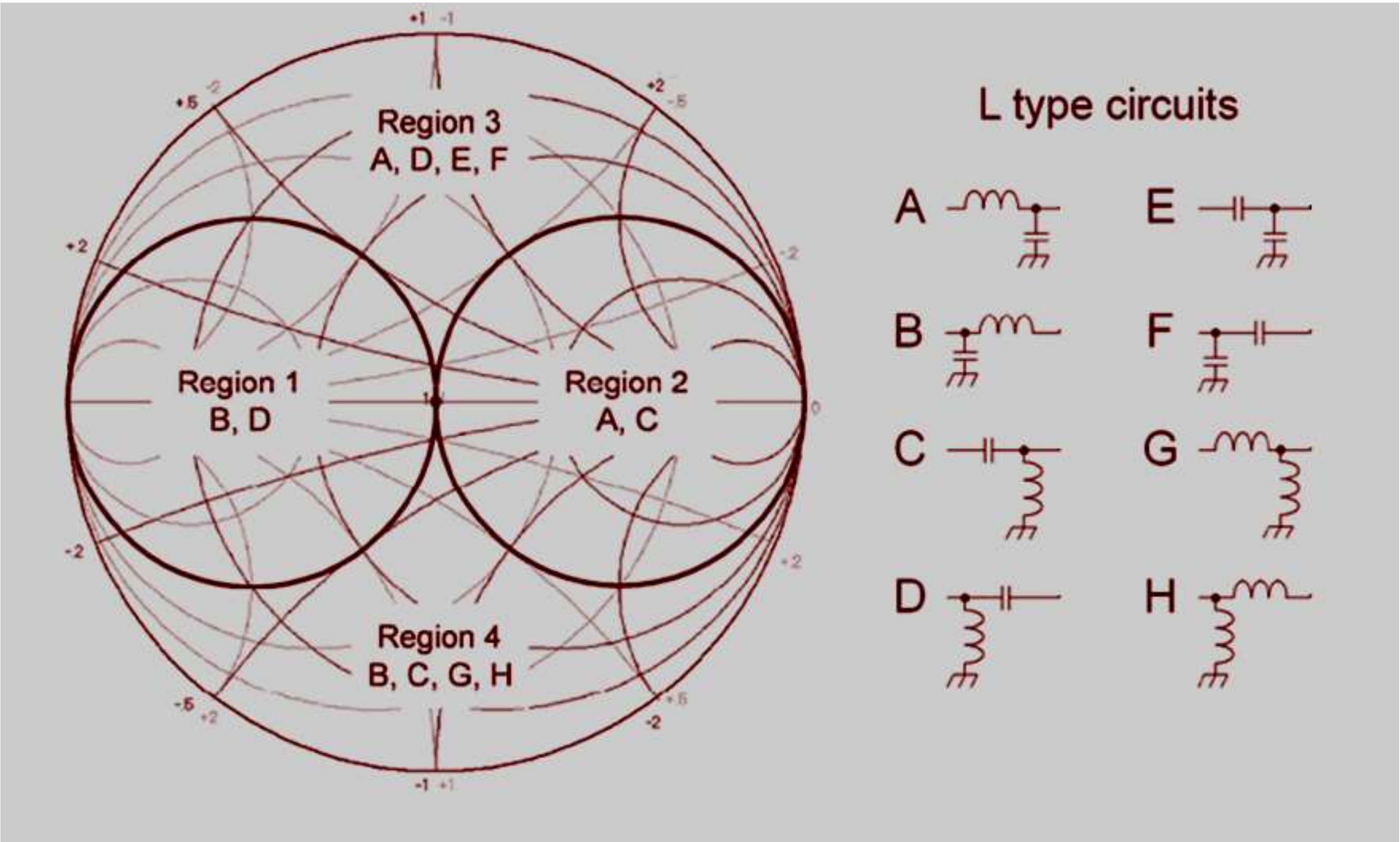
Impedance matching networks for amplifiers



Impedance matching network for antenna tuning

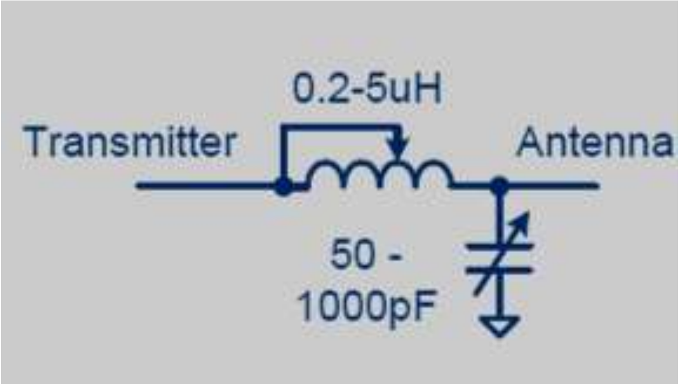


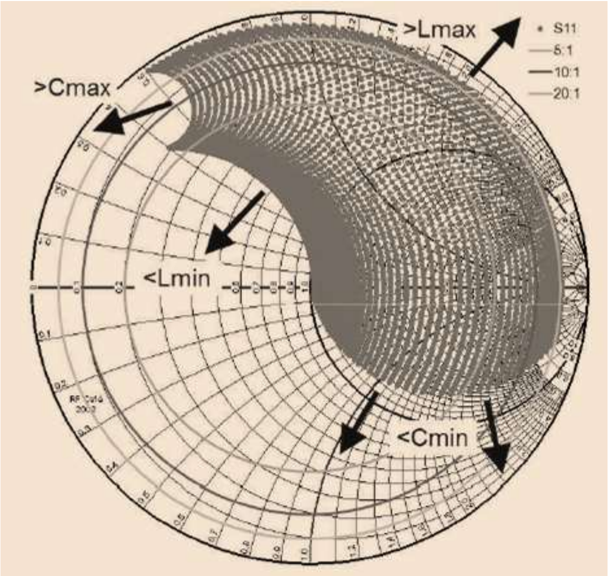
Load impedances that can be matched to 50 ohms source



Low Pass "L" Network

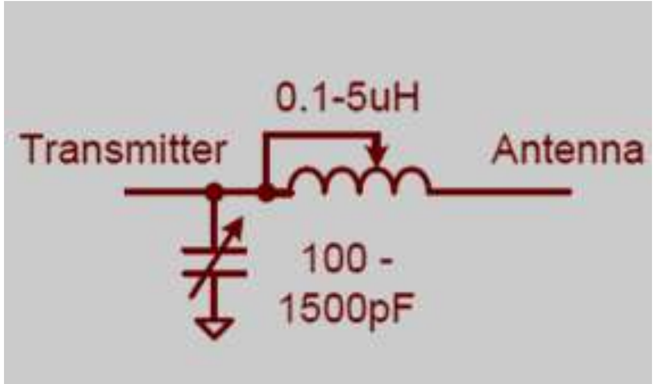
Type "**A**"

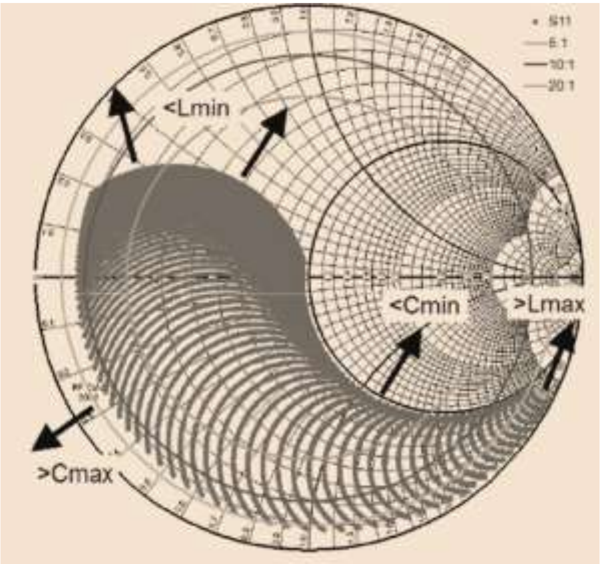




7 MHz

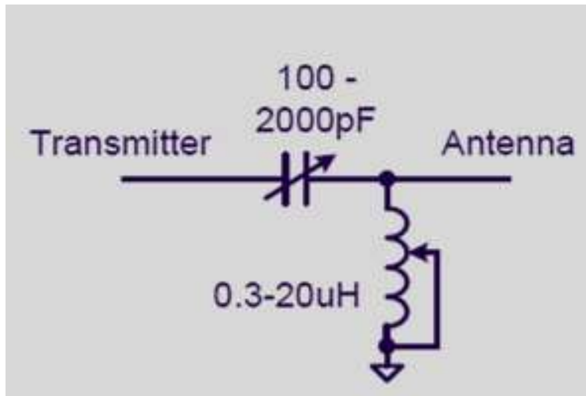
Type "**B**"

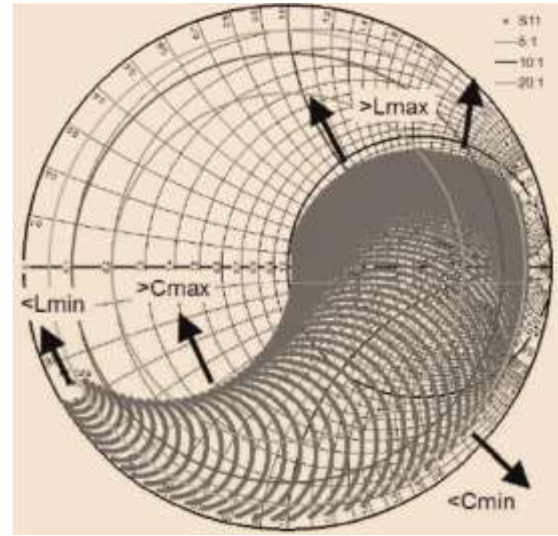




High Pass “L” network

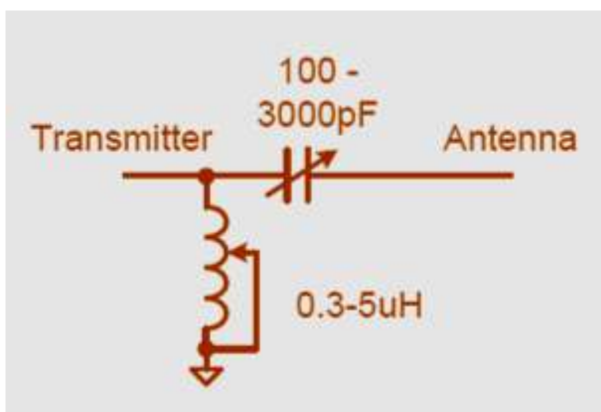
Type “C”

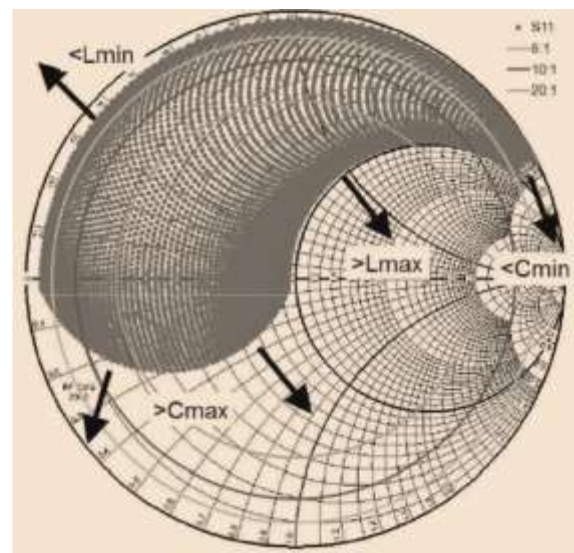




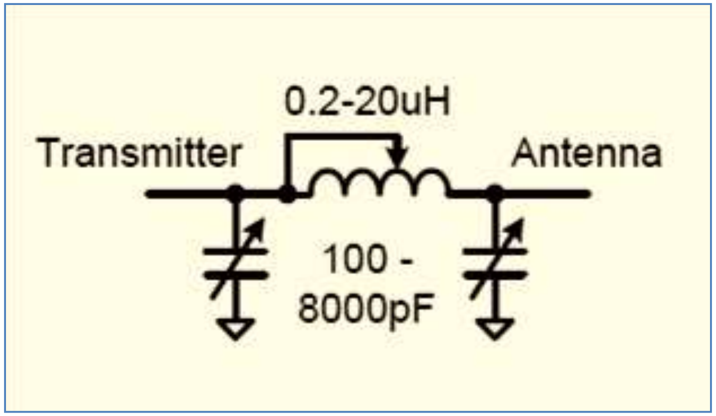
7 MHz

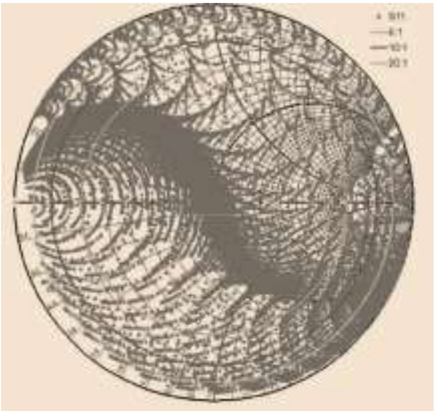
Type “D”



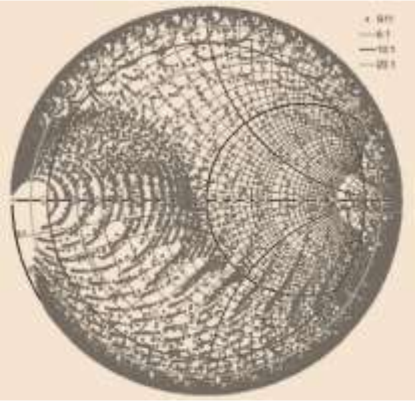


“Pi” Network

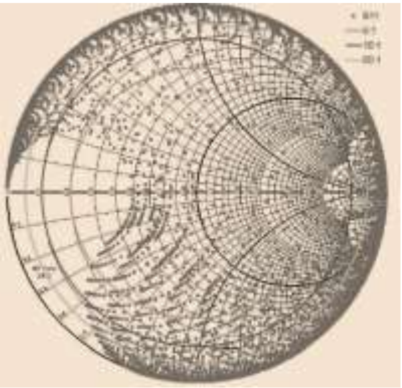




1.8MHz

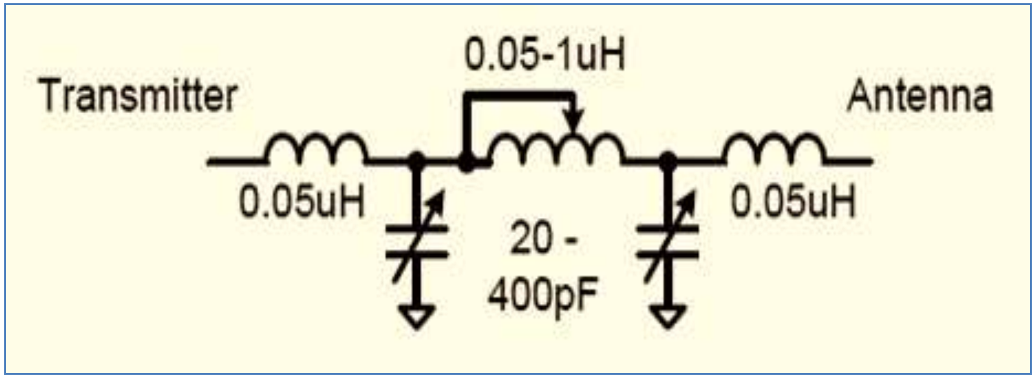


14MHz

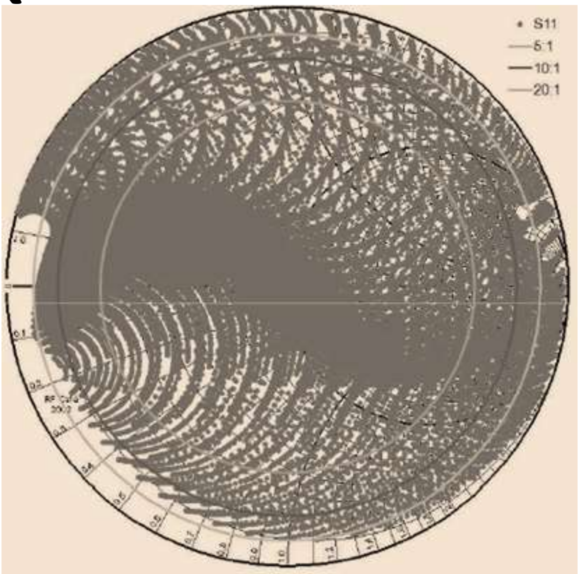


28 MHz

Including stray C and L to "Pi" Network



28 MHz

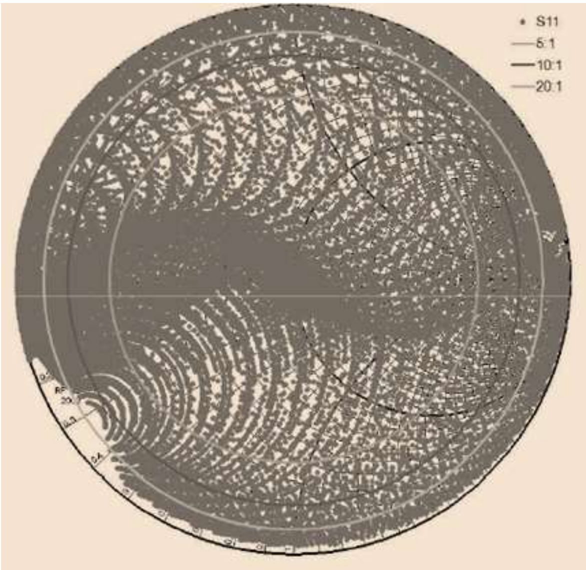


Very good matching range

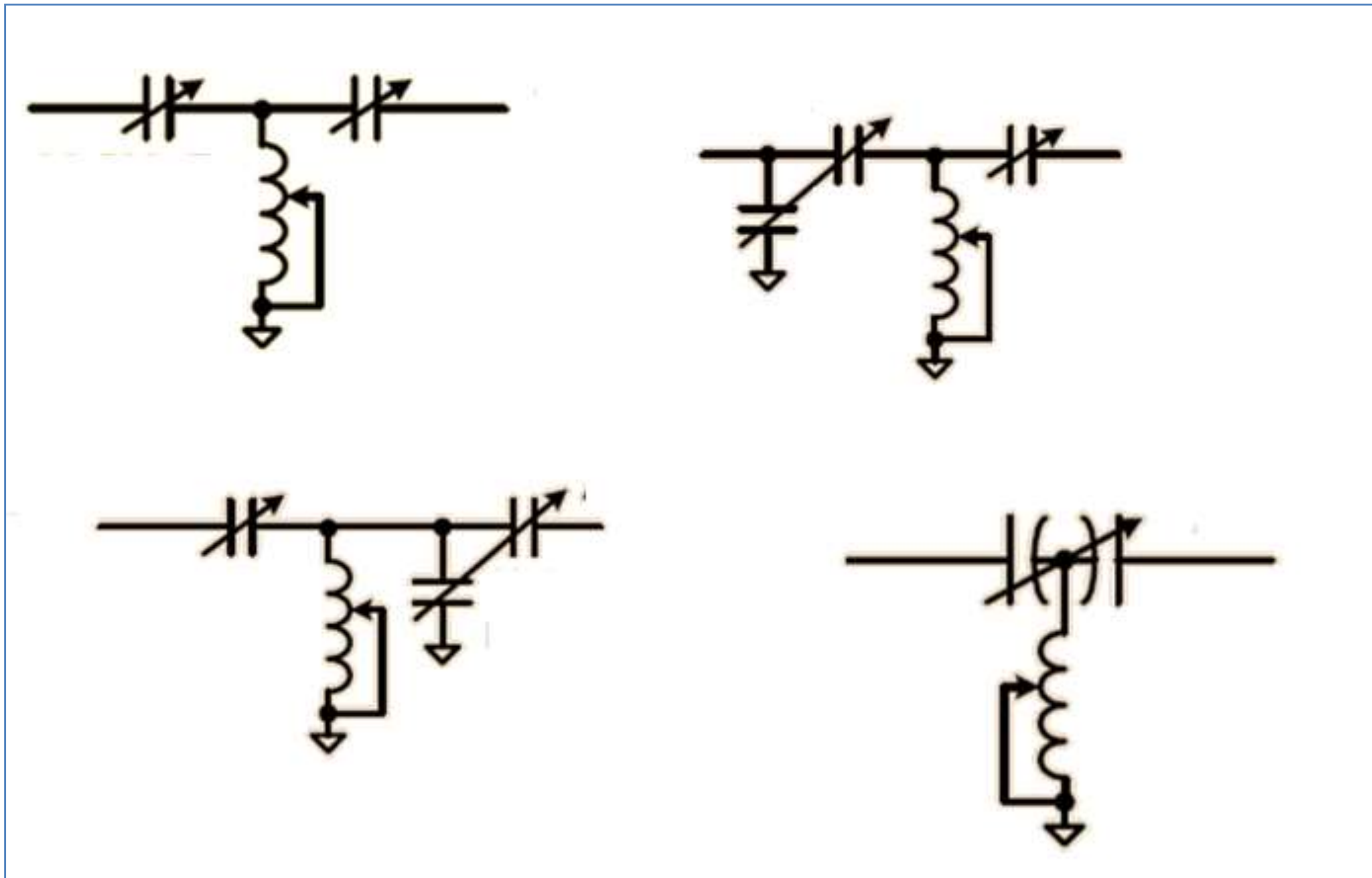
Modified for <Cmin which includes stray C to Gnd

Includes stray L on input and output

50 MHz



High-pass network topologies



Technologies used in developing variable capacitance

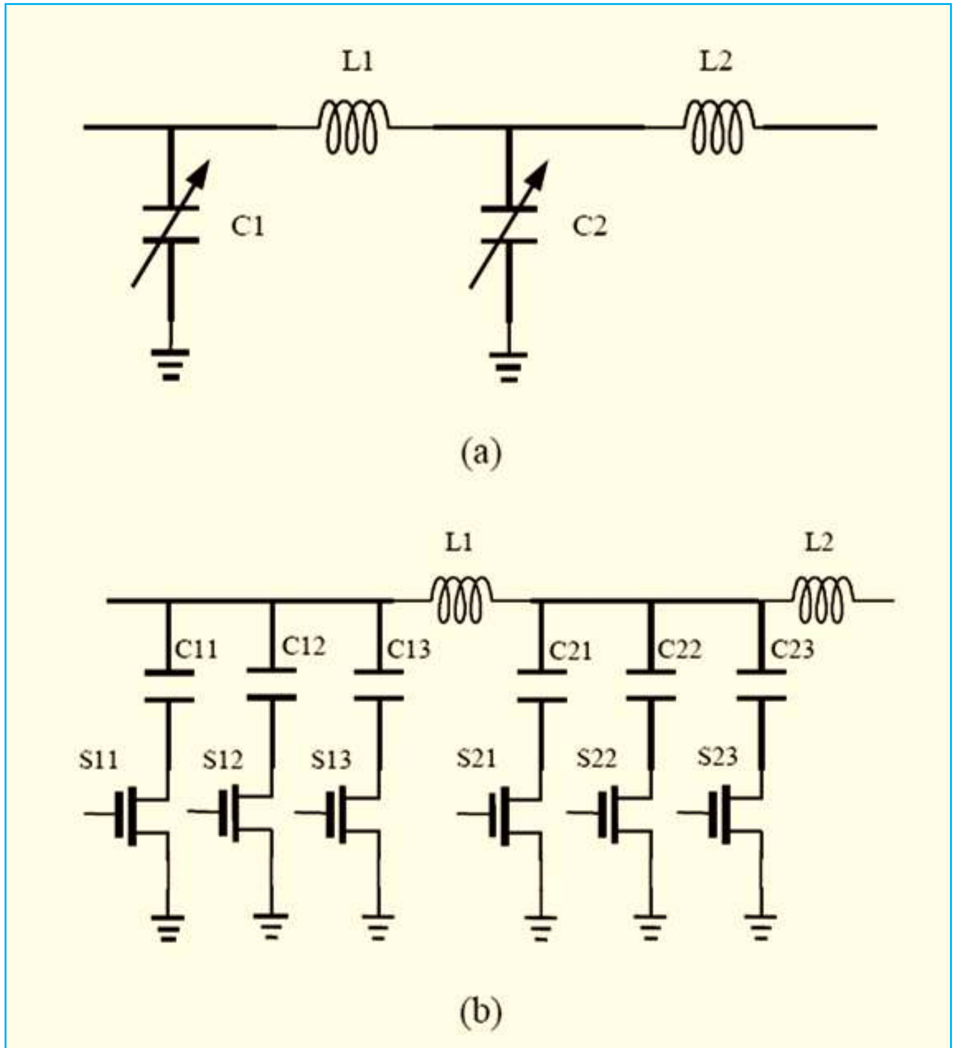
MEMS (Micro-electro-mechanical systems)

One of the two metallic plates is fixed while the position of the other plate is adjusted

BST (Barium Strontium Titanate) technology uses thin-film ferroelectric materials whose dielectric constant can be changed by applying High Voltage DC-bias

DTC (Digitally tunable capacitors) use variable capacitance in circuits controlled by digital control signals

Schematic of the tunable matching network



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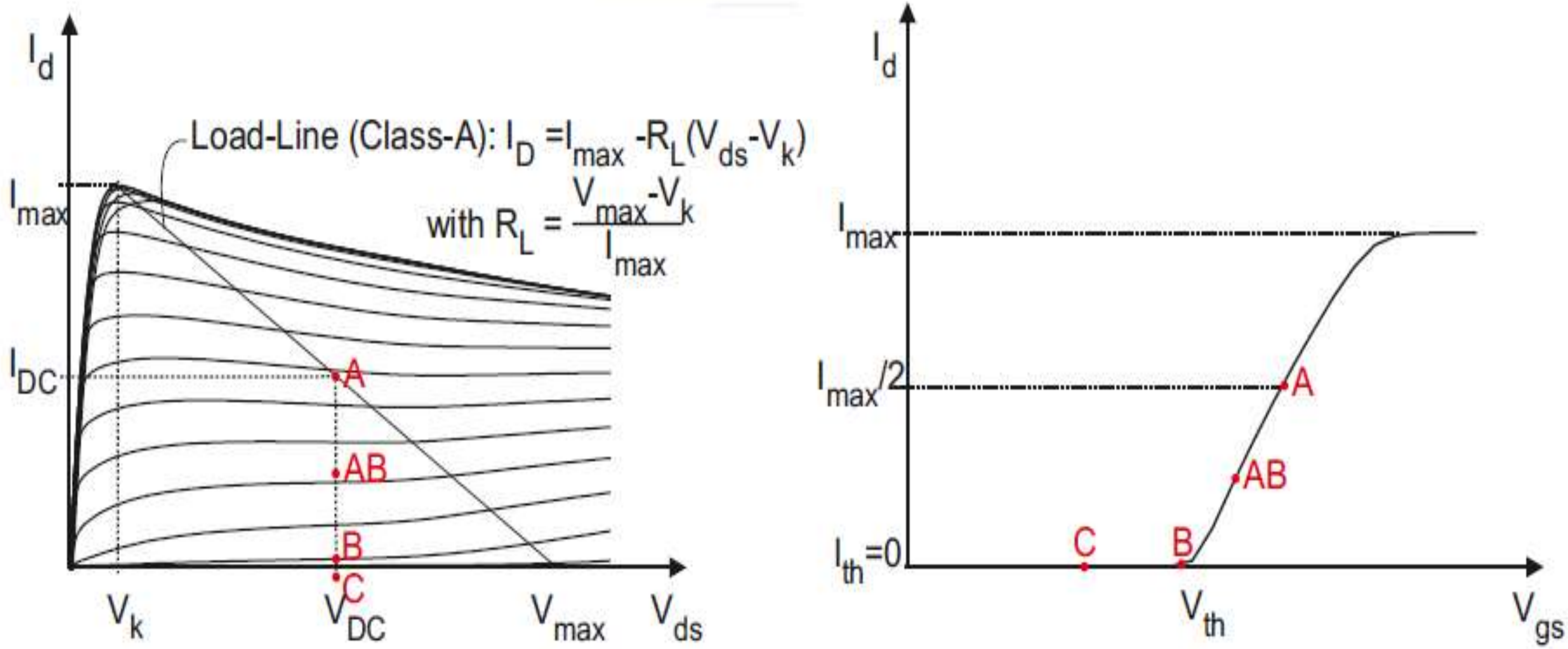
Power Amplifiers

Introduction to power amplifiers

- Low , medium, high power
- Narrowband and broadband
- CW and pulsed

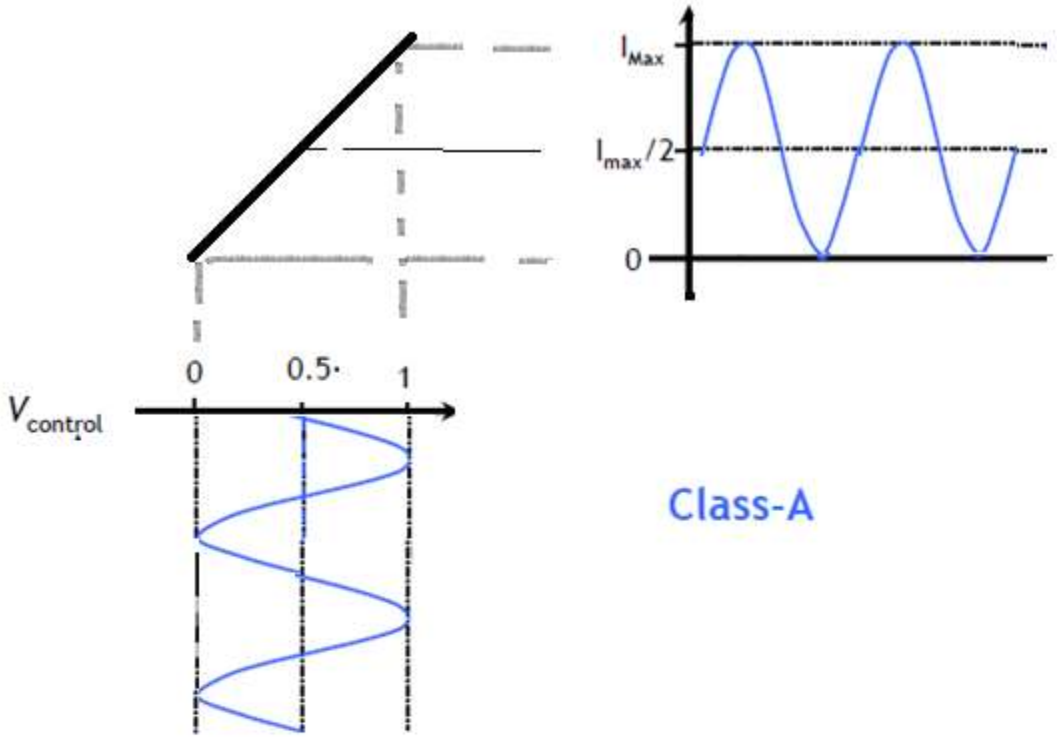
Class A, AB, B and C



Class A, AB, B and C are defined by the length of their conduction state over some portion of the waveform, such that the transistor state lies somewhere between "fully on" and "fully off".

Class A

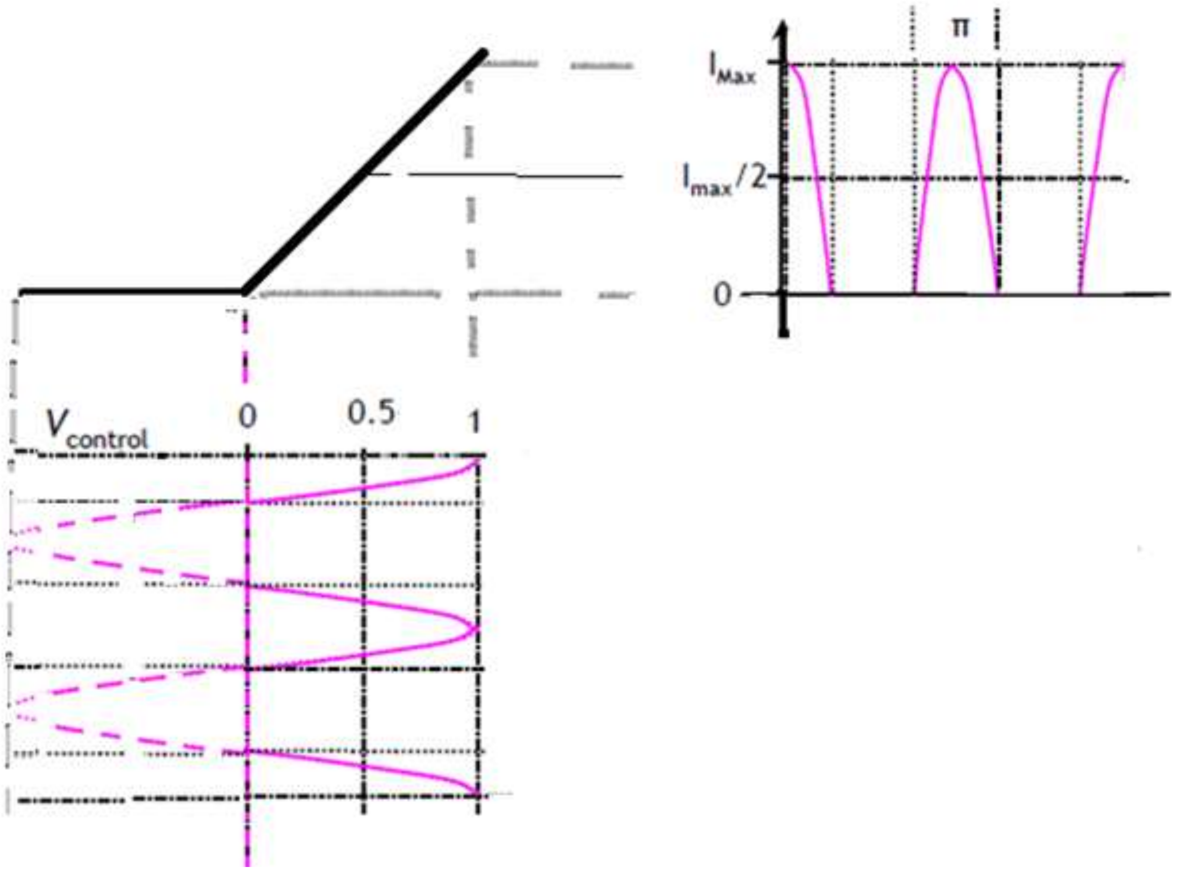
Conducts current over the full 360 degrees of a cycle.



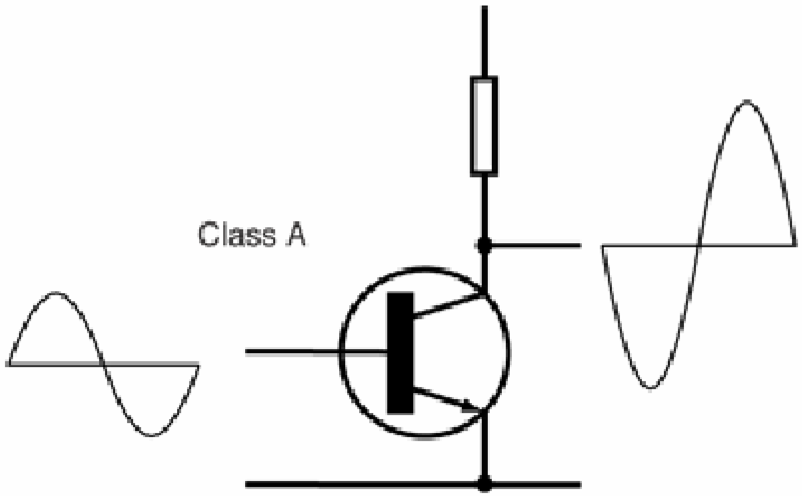
Class-A

Class B

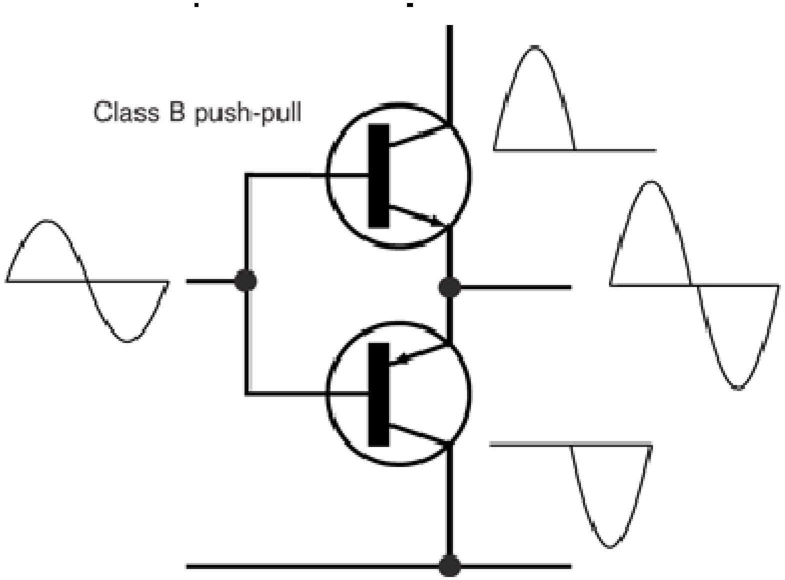
The conduction angle for the transistor is 180 degrees.



Class A vs. Class B

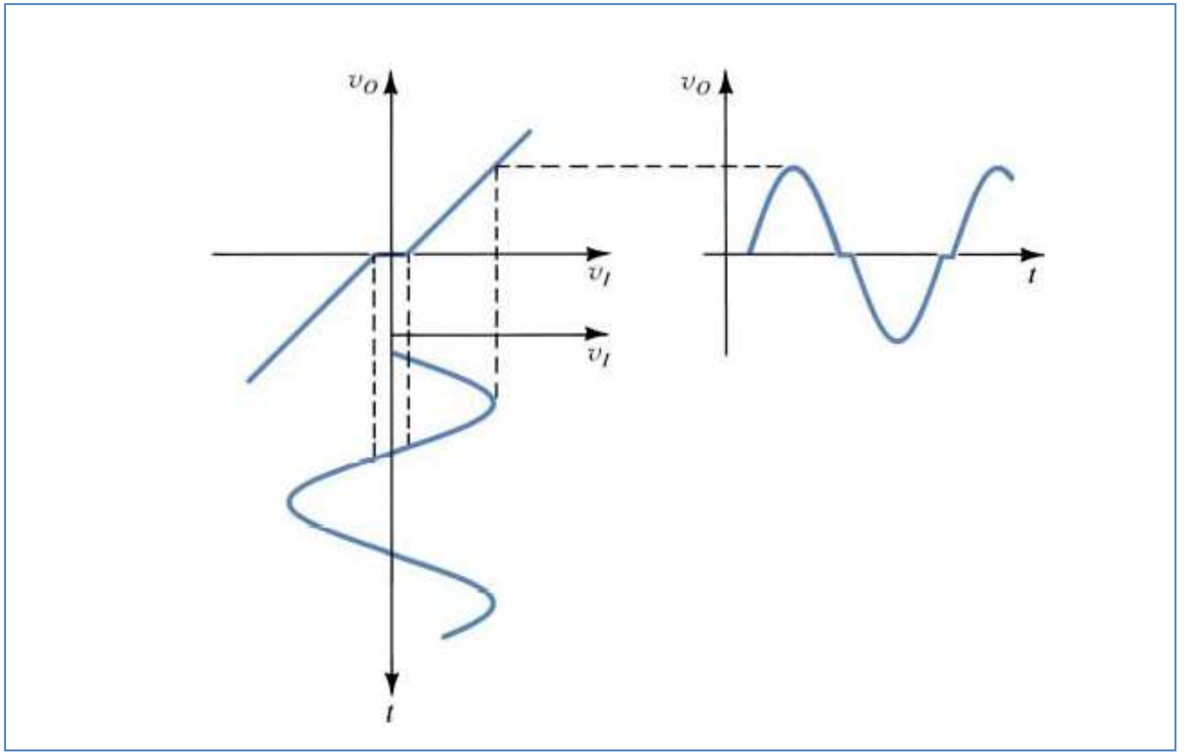


Class A Amplifier



Class B Push-Pull Amplifier

Crossover Distortion



Since the transistors are biased at Idq=0, the operating range includes the nonlinear cutoff region. This results in some output signal distortion known as **crossover distortion**

Class A

- **Advantages:**
- Simplicity
- Easy matching to standard impedances
- Linearity
- Wideband
- Low noise
- High gain

- **Disadvantages:**
- Limited Efficiency
- Low power

Class B

Advantages

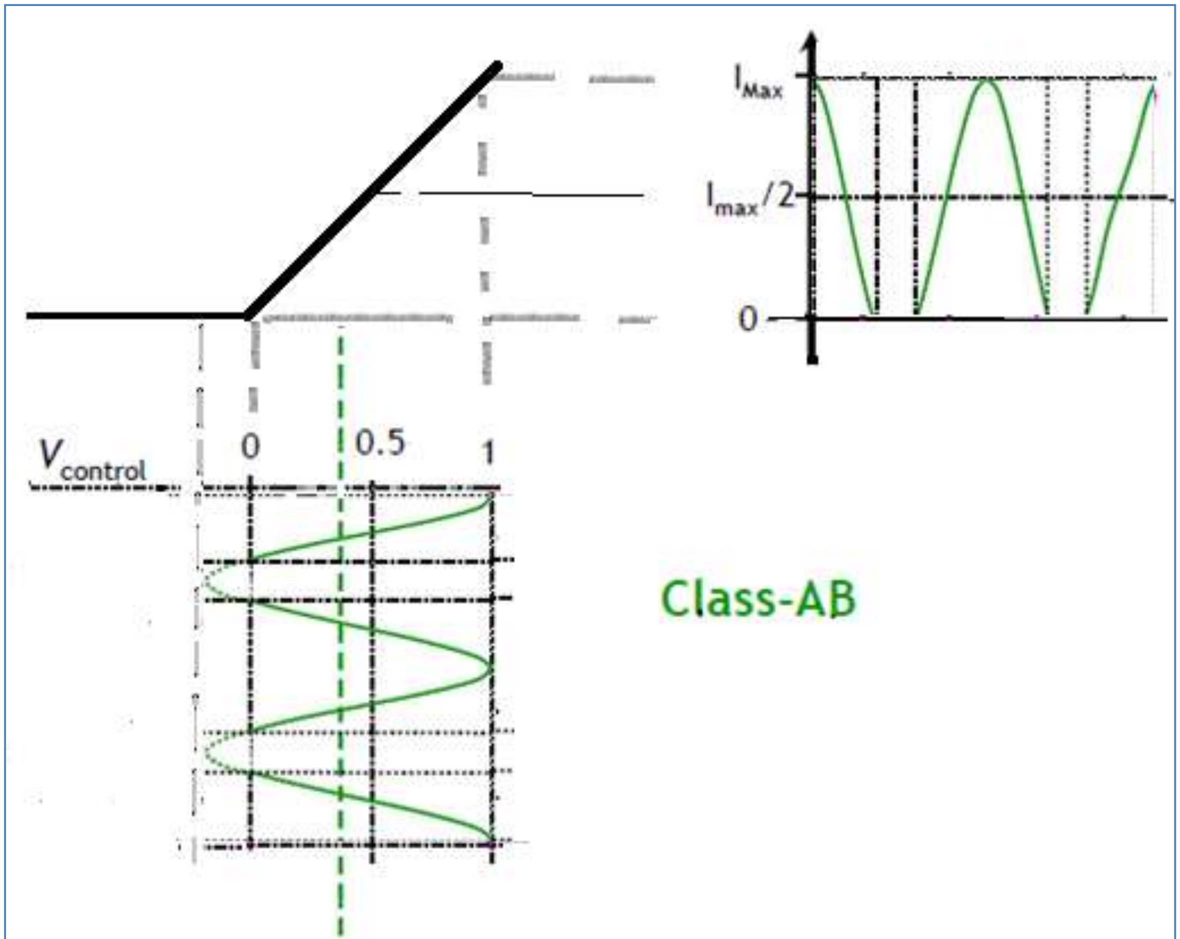
- Very low standing bias current
- Negligible power consumption without signal
- Can be used for much more powerful outputs than class A
- More efficient than Class A

Disadvantages

- Creates Crossover distortion
- Supply current changes with signal, stabilized supply may be needed
- More distortion than Class A

Class AB

The conduction angle is slightly >180 degrees.



Class AB amplifier

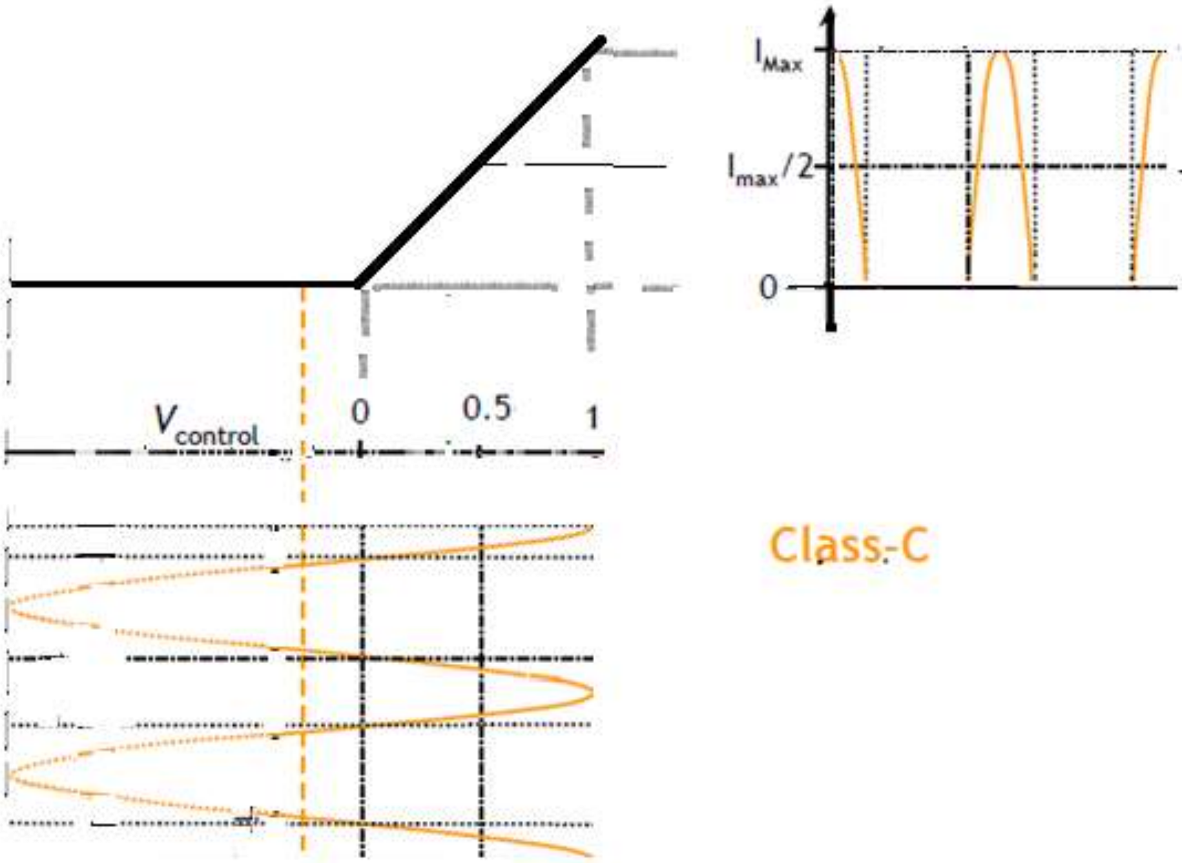
Advantages

- The Q-point remains in the linear region of the characteristic curves, which avoids the nonlinear distortion of the cutoff region and hence more linear than Class B amplifier

- Consumes some DC power for zero RF input however more efficient than Class A amplifier

Class C

The conduction angle is < 180 degrees



The output contains the fundamental plus higher frequency harmonics. The output may then be passed through an LC (inductor-capacitor) circuit tuned to the frequency of the input sinusoid.

Class C

Advantages:

- Simple topology
- Highly efficient
- Suitable for constant-carrier high power amplifiers
- External biasing is usually not needed, because is possible to force the transistor to provide its own bias, using an RF choke from base to ground.

Class C

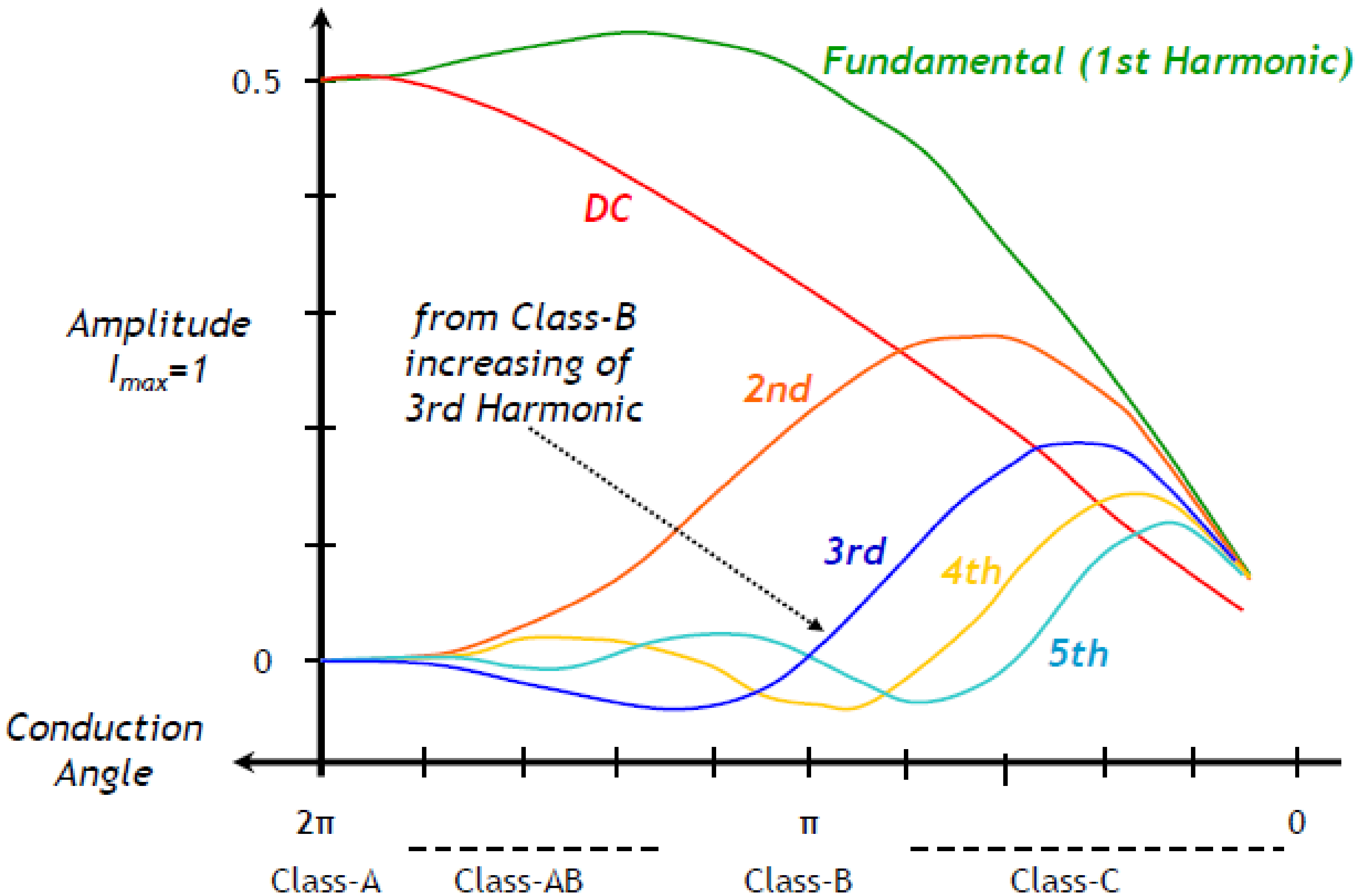
Disadvantages:

- Lower gain than Class-A or AB
- Strong drive signal needed
- Abundant harmonic generation
- Band-limited because usually resonant network needed on output.
- The large negative swing of the input voltage is the worst condition for reverse breakdown in any kind of transistor, and even small amounts of leakage current flowing at this point of the cycle have an important effect on the efficiency.
- In order to survive Class-C operation, the transistor should have a collector voltage breakdown that is at least three times the active device's DC voltage supply.

Class A, AB, B and C

Biasing	Efficiency max.	Possible bandwidth	Power capability	Gain
Class-A	50 %	high	low	high
Class-AB	50 - 78.5 %	high	medium/high	high
Class-B	78.5 %	moderate	high	medium/high
Class-C	78.5 - 100 %	low	high	low

Harmonics



Fundamental (1st Harmonic)
0.5
DC
Amplitude
I_{max}=1
from Class-B
increasing of
3rd Harmonic
2nd
3rd
4th
5th
0
Conduction
Angle
2π
π
0
Class-A
Class-AB
Class-B
Class-C

Class-D, E and F – switching amplifiers

- The class D, E, F and inverse F switching amplifiers limit dissipation by virtue of carefully selecting the times when the transistor switches.

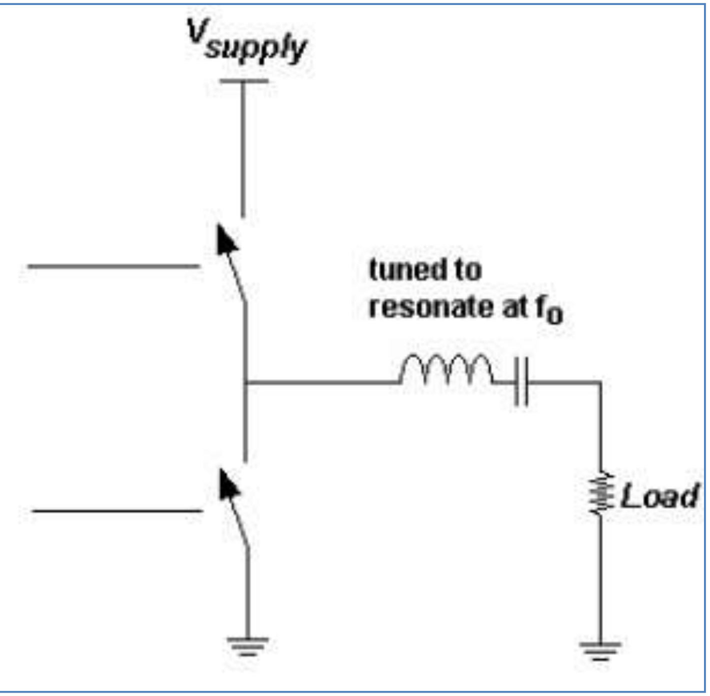
Class D

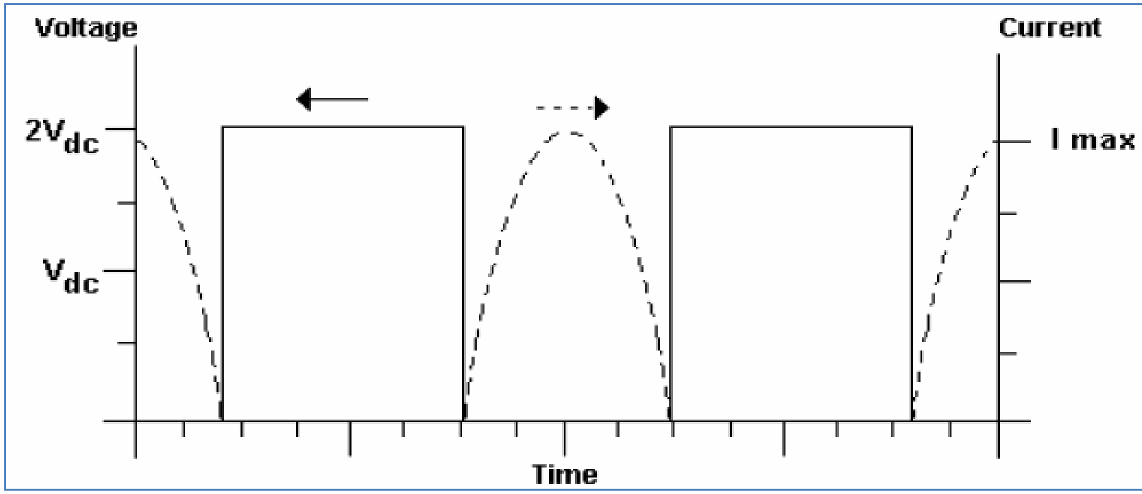
- The voltage mode Class D amplifier is defined as a switching circuit that results in the generation of a half-sinusoidal current waveform and a square voltage waveform.

- **Class-D PAs use two or more transistors as switches.**

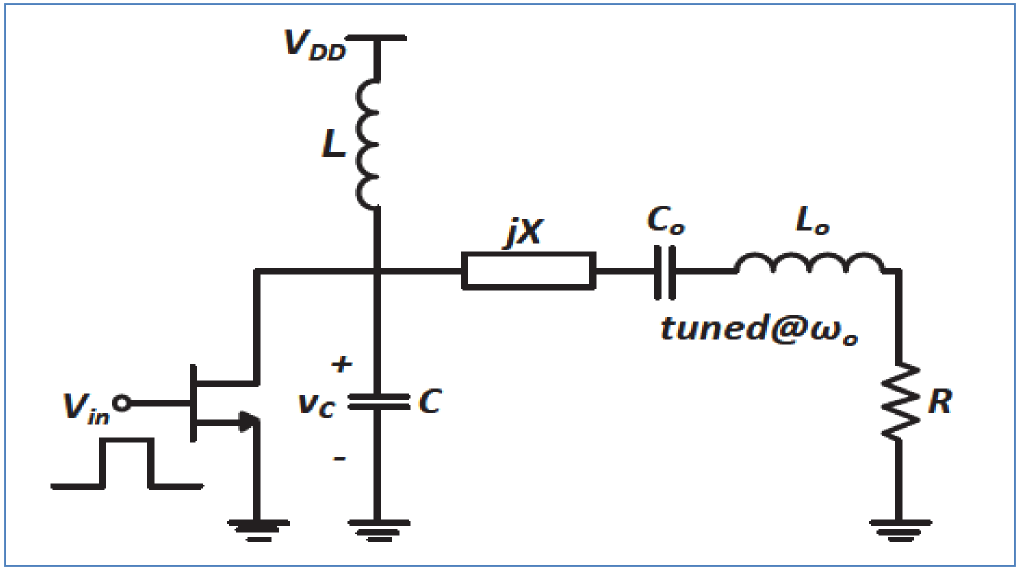
- A series-tuned output filter passes only the fundamental-frequency component to the load

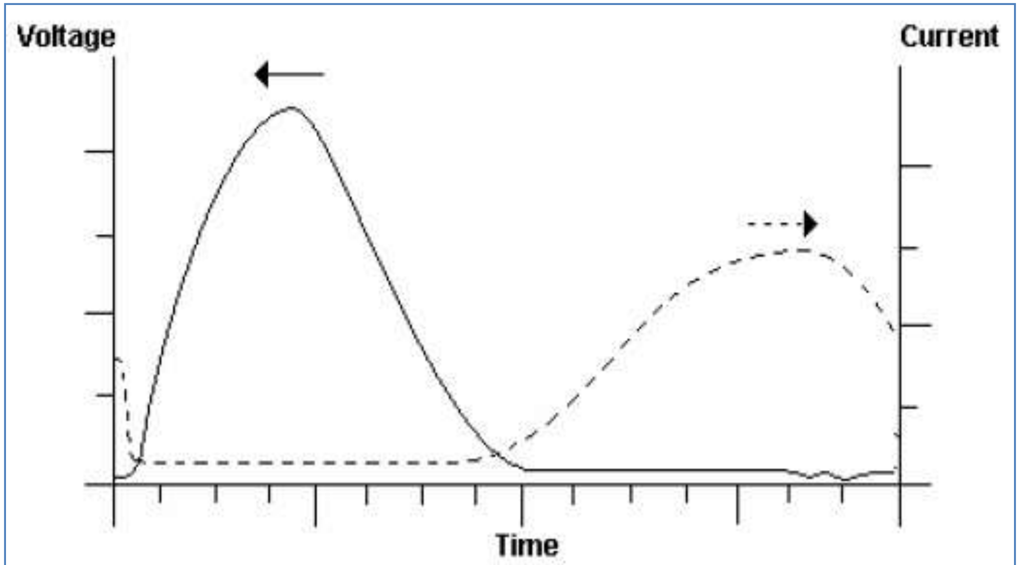
Class D voltage and current waveforms





Class E

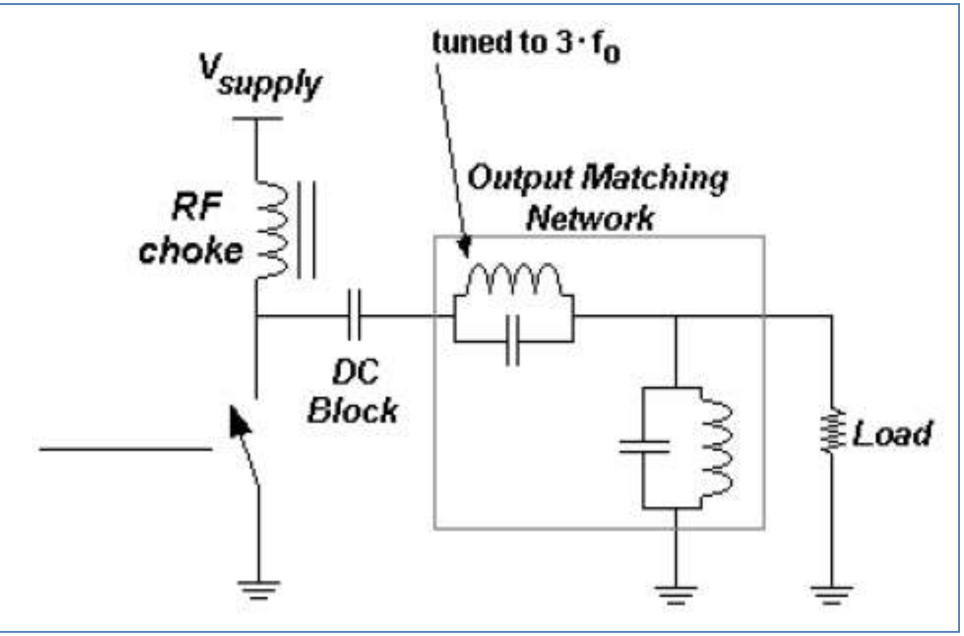


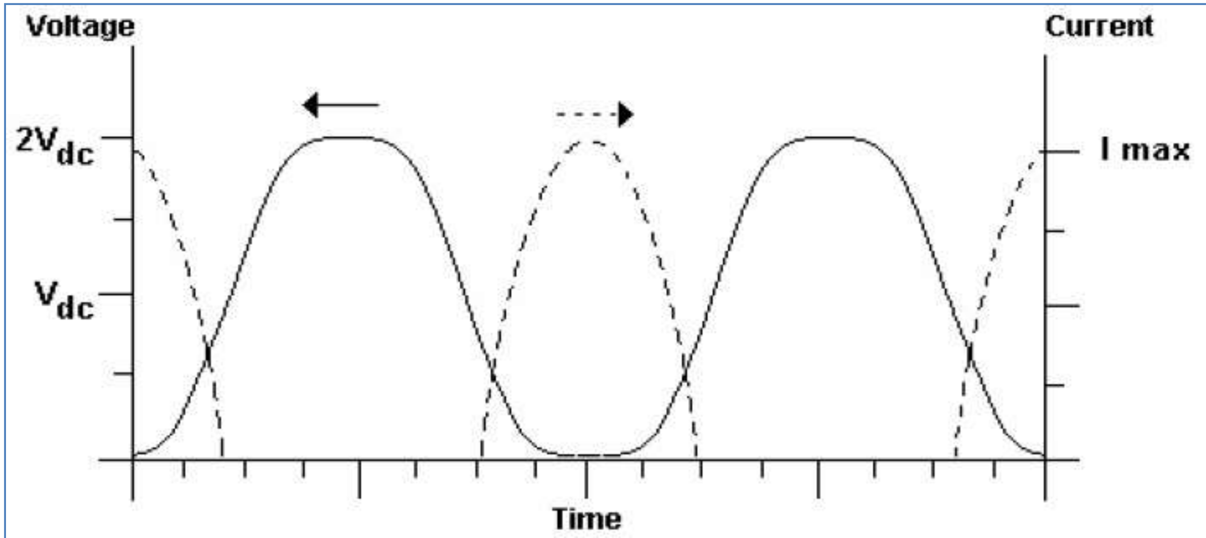


Class E

- Class-E employs a single transistor operated as a switch.
- In optimum class E, the drain voltage drops to zero and has zero slope just as the transistor turns on. The result is an ideal efficiency of 100 %, elimination of the losses associated with charging the drain capacitance.
- The Q of the output circuit is high enough so that the output current and output voltage consist of only the fundamental component

Class F

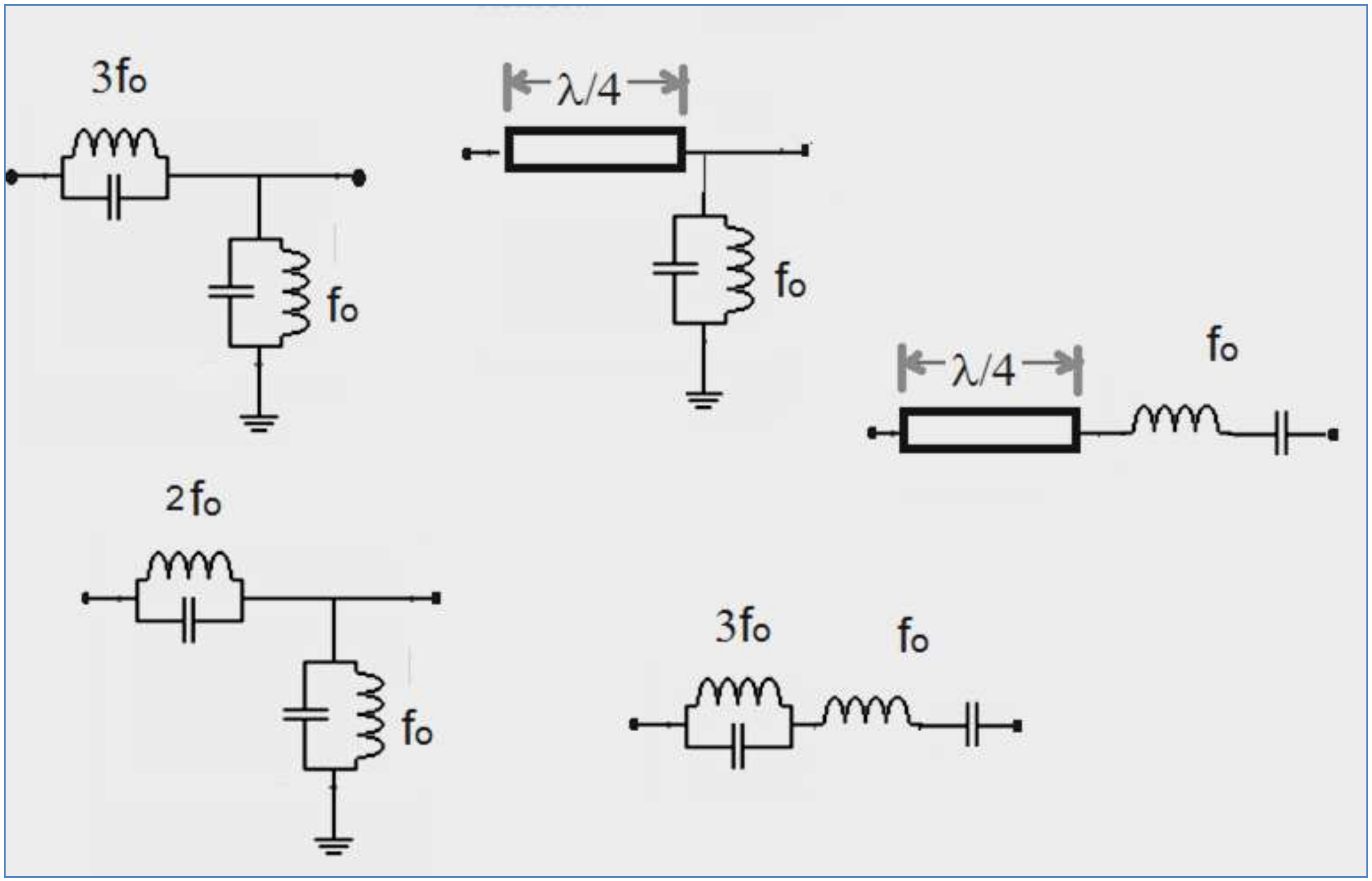

tuned to 3·f0
Vsupply
Output Matching
Network
RF
choke
DC
Block
Load


Voltage
Current
2Vdc
I max
Vdc
Time

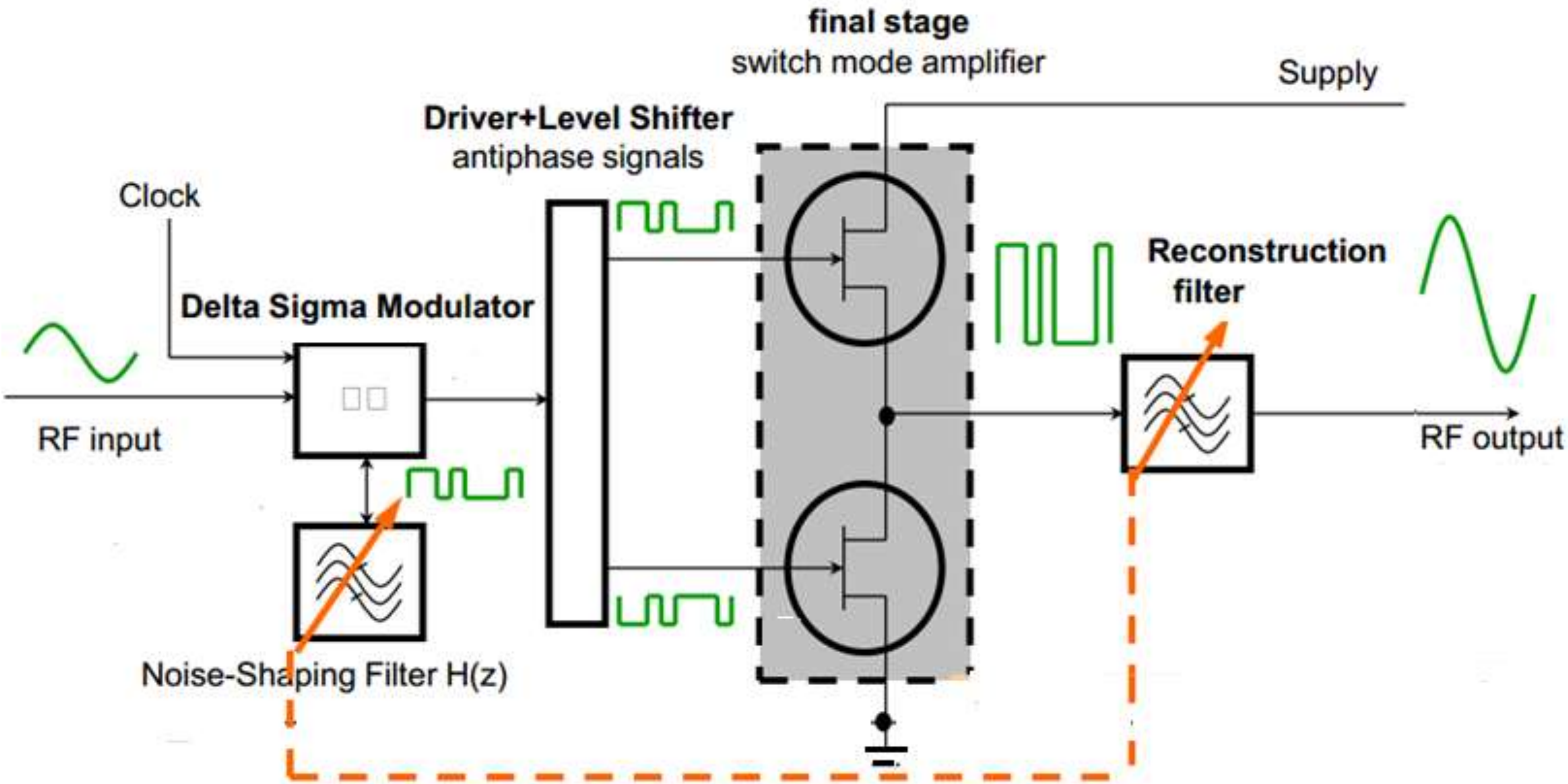
Class F

- Class-F boosts both efficiency and output by using harmonic resonators in the output network to shape the drain waveforms.
- The voltage waveform includes one or more odd harmonics and approximates a square wave, while the current includes even harmonics and approximates a half sine wave.
- Alternately (“inverse class F”), the voltage can approximate a half sine wave and the current a square wave.

Output matching networks for Class F variants

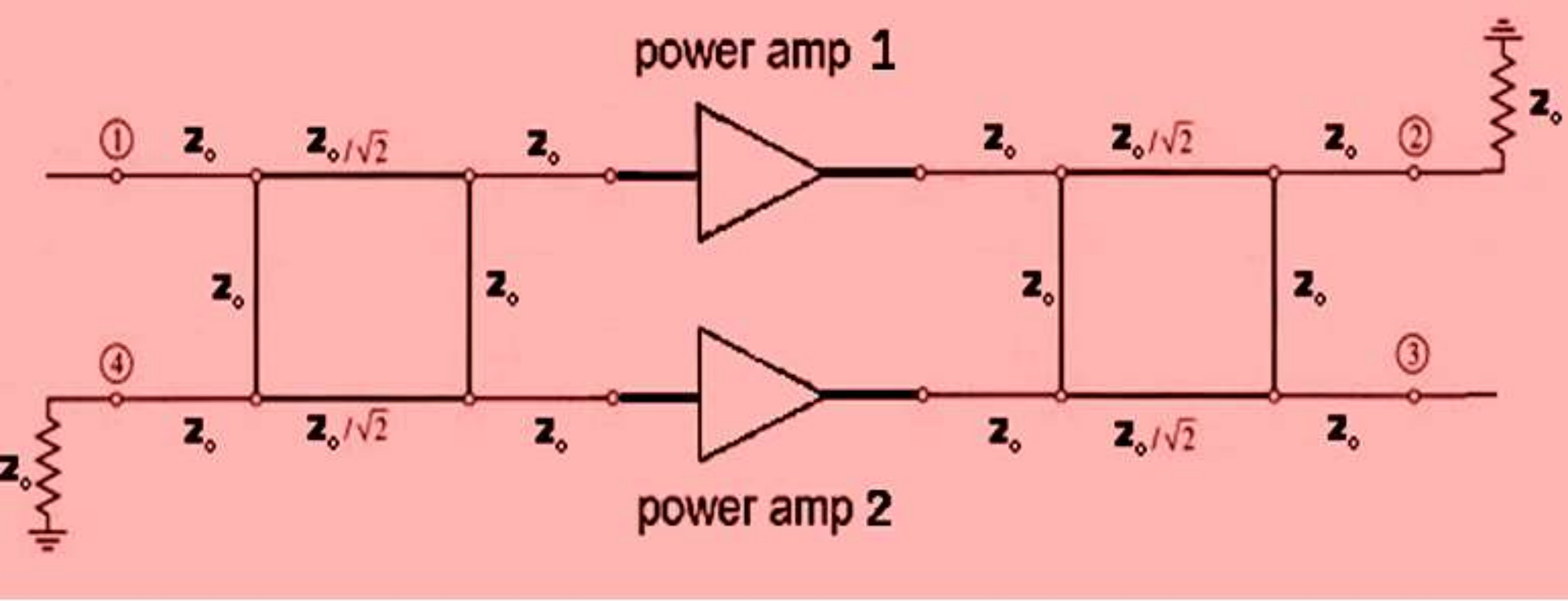


Class S amplifier architecture

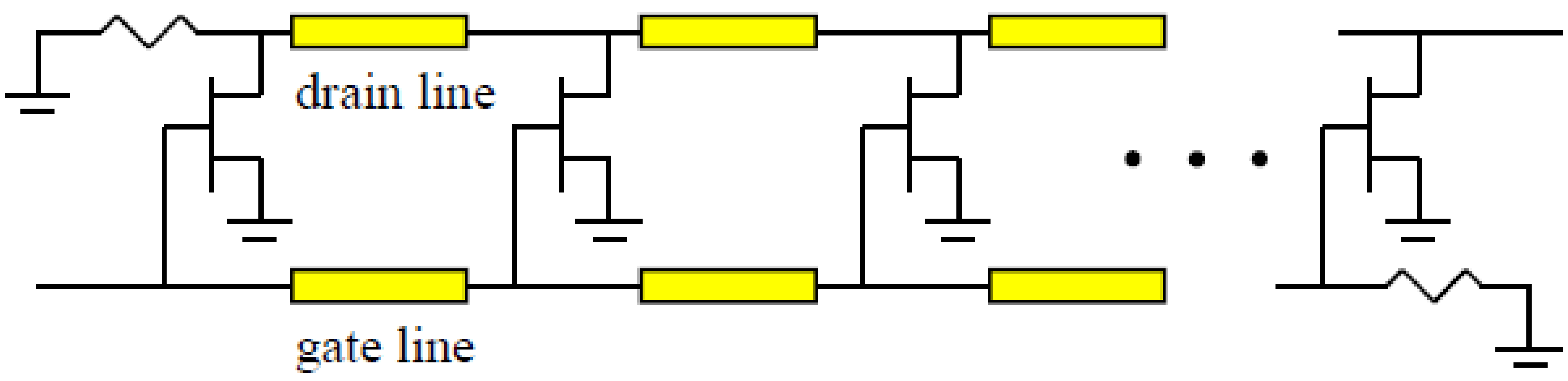


Class S amplification requires conversion of audio frequency signal into PWM signal

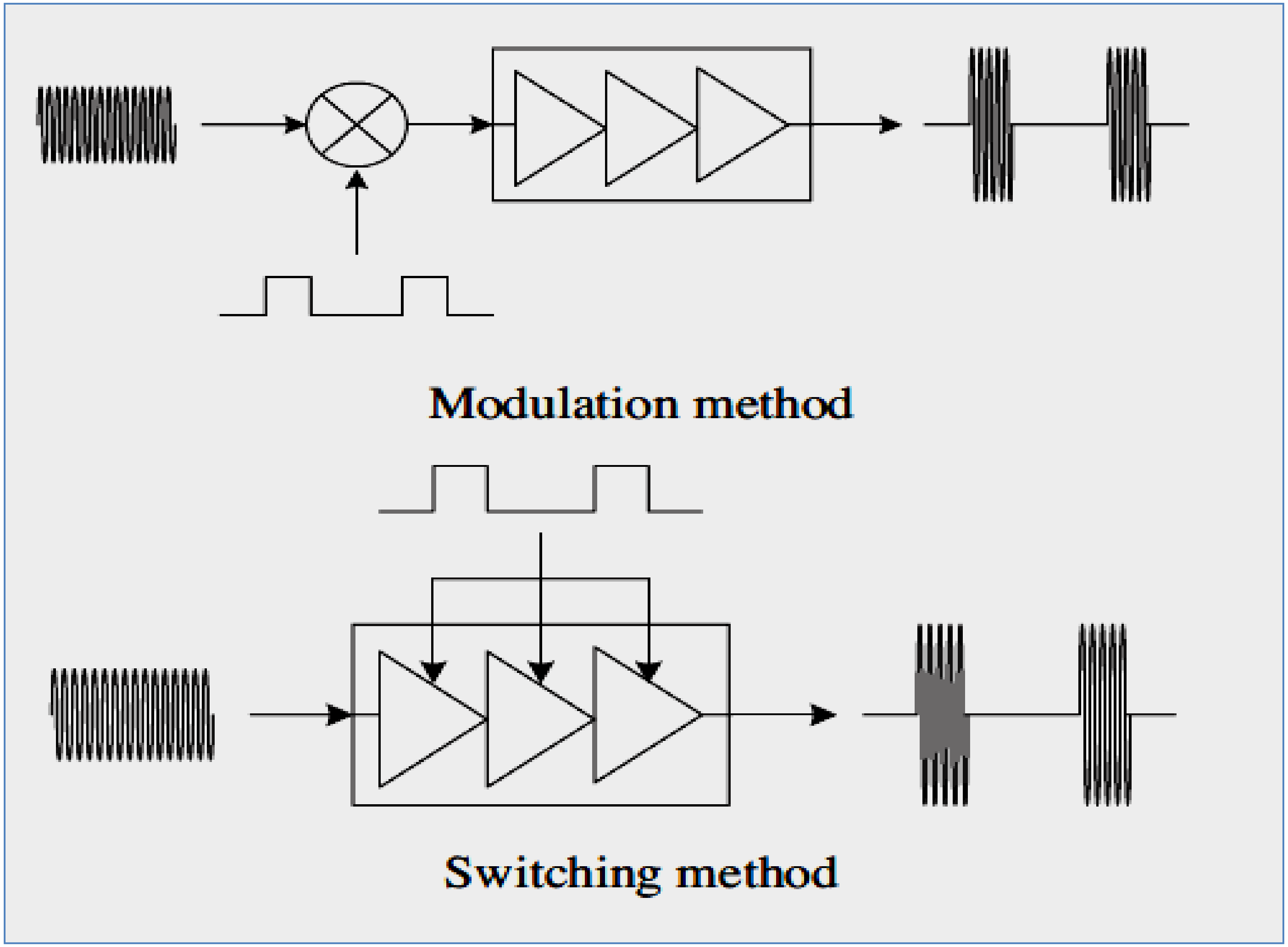
Balanced amplifier



Distributed amplifier



Pulsed power amplifiers



Practical considerations in designing a RF PCB

Designing for

cost-sensitive, high volume application

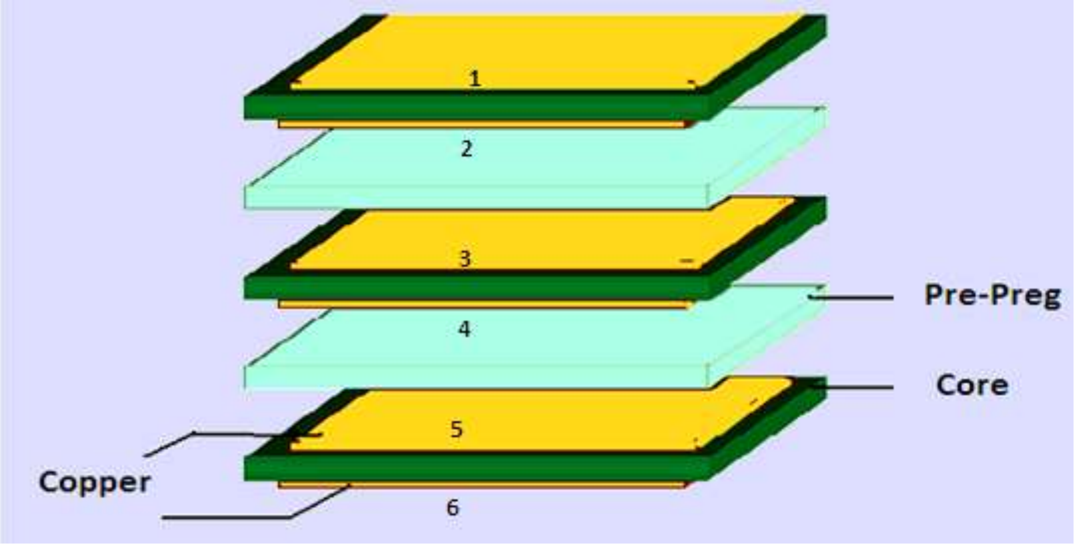
and

high performance, low volume application

Some Substrate Materials

Material	Dielectric Constant	Loss Tangent	Other Characteristics
RT Duroid 5880 Composite PTFE fiber glass	2.2	0.0009 @ 10 GHz	Low-cost "soft" substrate, widely used
RT Duroid 5870 Composite PTFE fiber glass	2.33	0.0012 @ 10 GHz	Low-cost "soft" substrate, widely used
CLTE Arlon	2.94 @ 10 GHz	0.0025 @ 10GHz	Stable Єr, low loss
RO4350B Rogers	3.48 @ 10 GHz	0.004 @ 10 GHz	Stable Єr, low loss, processing is similar to FR4
N7000-1 Nelco	3.9 @ 2.5GHz 3.8 @ 10 GHz	0.015@ 2.5 GHz 0.016 @ 10 GHz	High Tg (260 °C)
FR-4 Woven Glass/ Epoxy	4.7 @ 1 MHz, 4.3 @ 1GHz	0.030@ 1 MHz 0.020 @ 1 GHz	Inexpensive, unstable Єr, high loss
Alumina	9.0-10.0	<0.0015 to 25 GHz	Characteristics depend on manufacture, k=9.8 is more common
RT6010LM Rogers	10.2 @ 10 GHz	0.002 @ 10 GHz	High Єr, low loss

PCB stack-up



2-layer

1st Layer: Component, RF traces, Power , Signal, Ground
2nd Layer: RF Ground Plane

4-layer

1st Layer: Component and RF traces
2nd Layer: RF Ground Plane
3rd Layer: Power Plane
4th Layer: Ground Plane and Signal Routing

6-layer

1st Layer: Component and RF traces
2nd Layer: RF Ground Plane
3rd Layer: Signal
4th Layer: Signal
5th Layer: Power Plane
6th Layer: Ground Plane

Identifying components

Information from datasheets

Frequency range

RF power handling capability

Supply voltage

Absolute maximum ratings

Extent of pre-matching

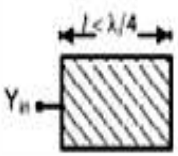
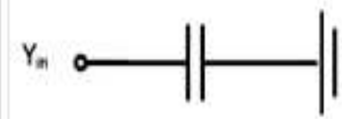
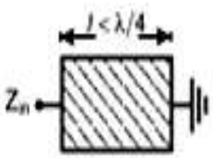
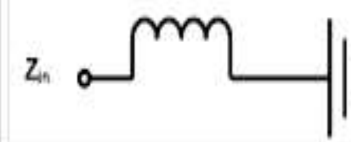
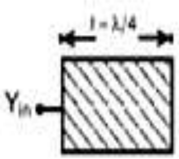
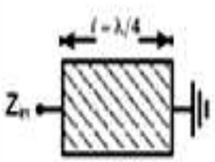
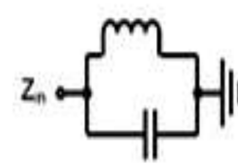
Case style

Operating temperature

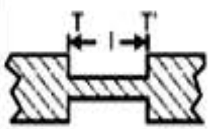
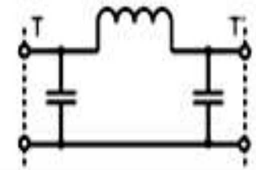
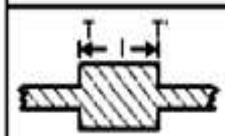
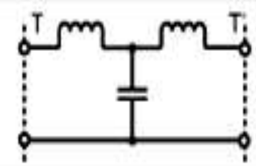
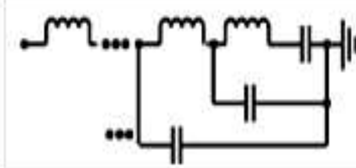
Heat dissipation capability

Layout and mounting recommendation

Transmission line patterns and uses

Transmission Line Pattern	Structure	Characteristics	Equation	Accuracy	Equivalent Circuit
	Short Stub	Inductive when $\beta \cdot l < \pi/2$	$X = Z_0 \cdot \tan(\beta \cdot l)$	Exact	
	Open Stub	Capacitive when $\beta \cdot l < \pi/2$	$B = Y_0 \cdot \tan(\beta \cdot l)$	Exact	
	Quarter-wave Open Stub	Equivalent to a series LC resonator	$L = (\pi \cdot Z_0)/(4 \cdot \omega_0)$ $C = 1/(\omega_0^2 \cdot L)$	Approximate; based on equating $dB/d\omega$ at resonance	
	Quarter-wave Short Stub	Equivalent to a parallel LC resonator	$C = (\pi \cdot Y_0)/(4 \cdot \omega_0)$ $L = 1/(\omega_0^2 \cdot C)$	Approximate; based on equating $dB/d\omega$ at resonance	

Transmission line patterns and uses

Transmission Line Pattern	Structure	Characteristics	Equation	Accuracy	Equivalent Circuit
	High Impedance Series Line	Equivalent to a series inductance	$X = Z_0 \cdot \tan(\beta \cdot l)$	Approximate; OK when $\beta \cdot l << \pi/4$	
	Low Impedance Series Line	Equivalent to a parallel capacitance	$B = Y_0 \cdot \tan(\beta \cdot l)$	Approximate; OK when $\beta \cdot l << \pi/4$	
	Radial Stub	Almost exclusively for a broadband short	No simple equation describes it adequately	No exact expression	

Impedance matching networks

- The impedance matching networks could be a combination of lumped elements and transmission lines.

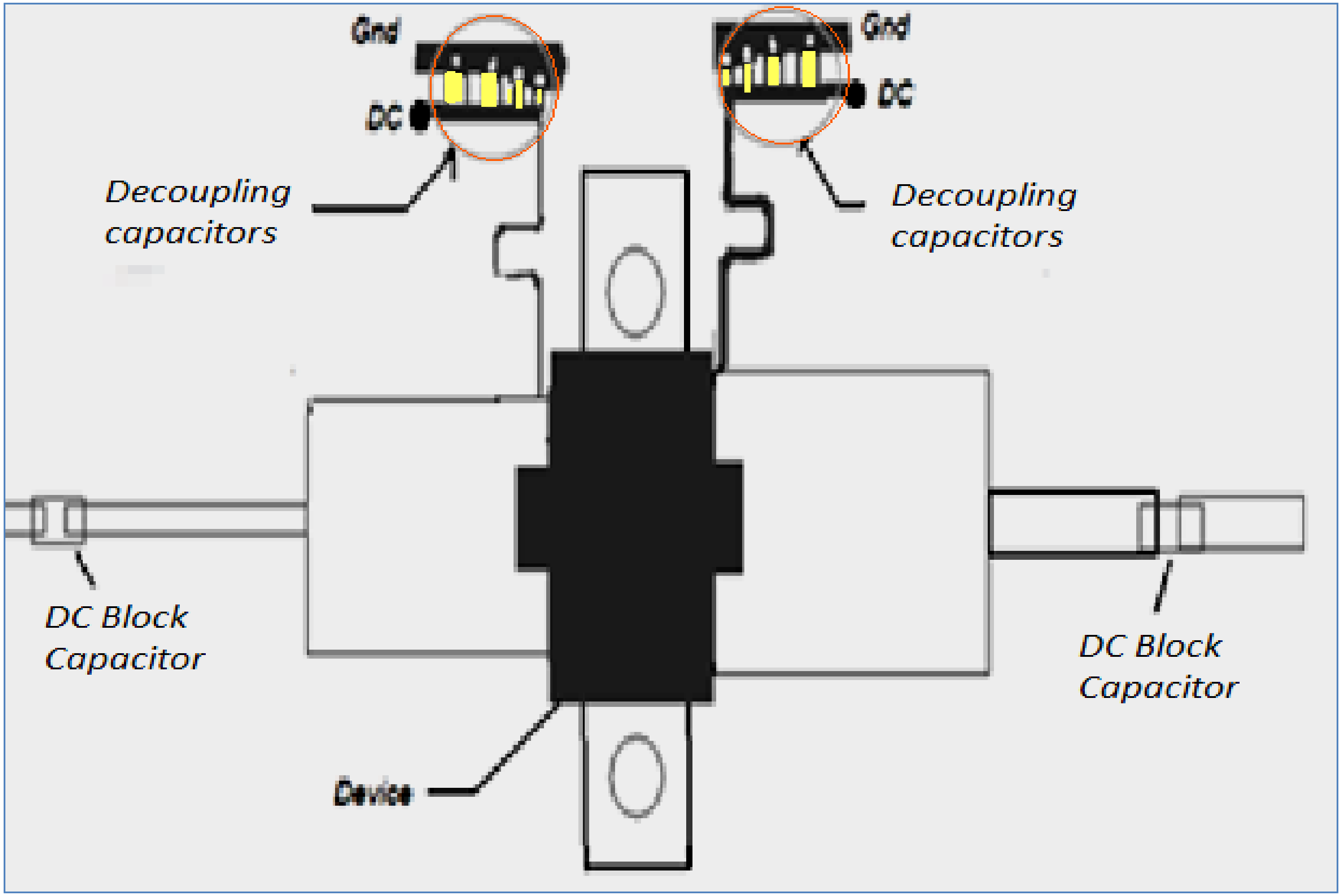
- The matching network topology has to be selected taking into account the bandwidth requirement, practically realizable/available component values.

- High Q lumped elements reduce losses in matching networks

Lumped component selection

- High Q inductors and capacitors
- The self-resonant frequency of the inductor should be higher than the frequency of operation.
- For DC blocking the series resonant frequency of the capacitor should be greater than the frequency of operation.

Bias decoupling



Decoupling capacitors

Useful Frequency Ranges of Capacitors

Component	Capacitance	Package	SRF	Useful Frequency Range*
Ultra-High Range	20pF	0402	2.5GHz	800MHz to 2.5GHz
Very High Range	100pF	0402	800MHz	250MHz to 800MHz
High Range	1000pF	0402	250MHz	50MHz to 250MHz
Midrange	1µF	0402	60MHz	100kHz to 60MHz
Low Range	10µF	0603	600kHz	10kHz to 600kHz

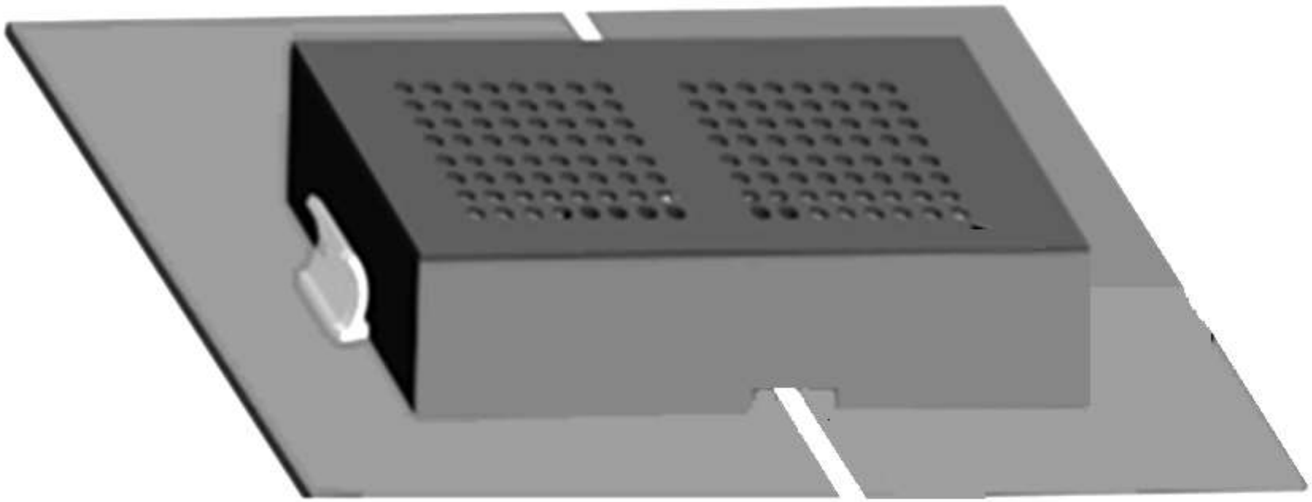
*Low end of useful frequency range defined as less than 5Ω of capacitive reactance.

EMI reduction

The electromagnetic emission/interference from the board can be reduced by

- Moving components on the PCB
- Adding/changing ground planes
- Reducing the length of noisy PCB traces and wires
- Using less noisy components
- Adding ferrite products
- Using special shielding techniques

Shielding



Board-level shields are the most cost-effective and generally the most efficient type of shielding.

Some do's and don'ts for designing RF PCBs

- Separate the analog, RF and digital parts
- Locate the microstrip lines on the top (component) side of the board
- Wherever possible the bottom side of the board should allow for a solid ground plane under the RF circuitry
- Ground plane vias around the RF circuits should be spaced closer than 1/20 of a wavelength as a minimum

Some do’s and don’ts for designing RF PCBs

- “Stitch” the top and bottom ground planes together with as many vias as possible

- Use solid ground plane under components that require heat sinking, with many closely spaced vias to transfer heat to all ground plane layers

- After finishing all layers, fill the empty spaces with copper pour that is connected to the ground.

Device structures and materials

Device Structures	Materials
HBT	SiGe
MOSFET	GaAs
MESFET	InGaAs
PHEMT	InGaP
HEMT	SiC
LDMOS FET	GaN

Comparison of semiconductor material properties

Material	Si	GaAs	4H-SiC	GaN
Relative dielectric constant ε_r	11.8	13.1	10	9
Bandgap E_g (eV)	1.1	1.42	3.26	3.39
Lattice constant (Å)	5.4	5.7	3.1	3.2
Electron mobility μ at 300 K (cm^2/Vs)	1350	8500	700	1200-2000
Saturated electron velocity v_{sat} ($10^7 cm/s$)	1	1	2	2.5
Breakdown field E_{br} (MV/cm)	0.3	0.4	3	3.3
Transit frequency f_T (GHz) FET	20	150	20	150
Thermal conductivity Θ (W/cmK)	1.5	0.43	3.3-4.5	1.3

PA device modeling

- Modeling from Simple DC and Linear Data
- Measurement-Based Models
- Table-Based Models
- Artificial Neural Network-Based Models
- Advanced Compact Models from NVNA Data
- X-Parameter-Based Models
- Closed-form physical equations

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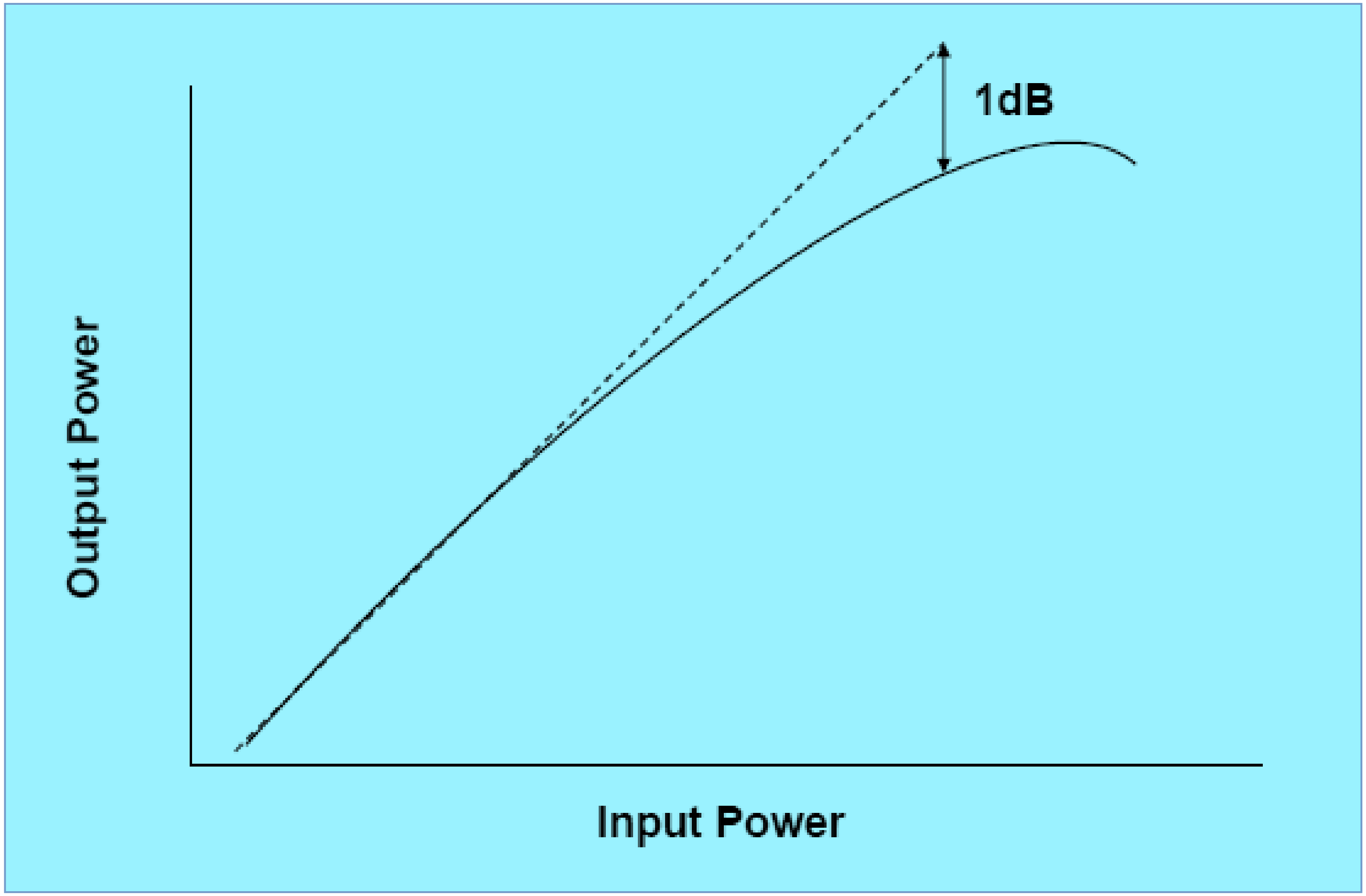
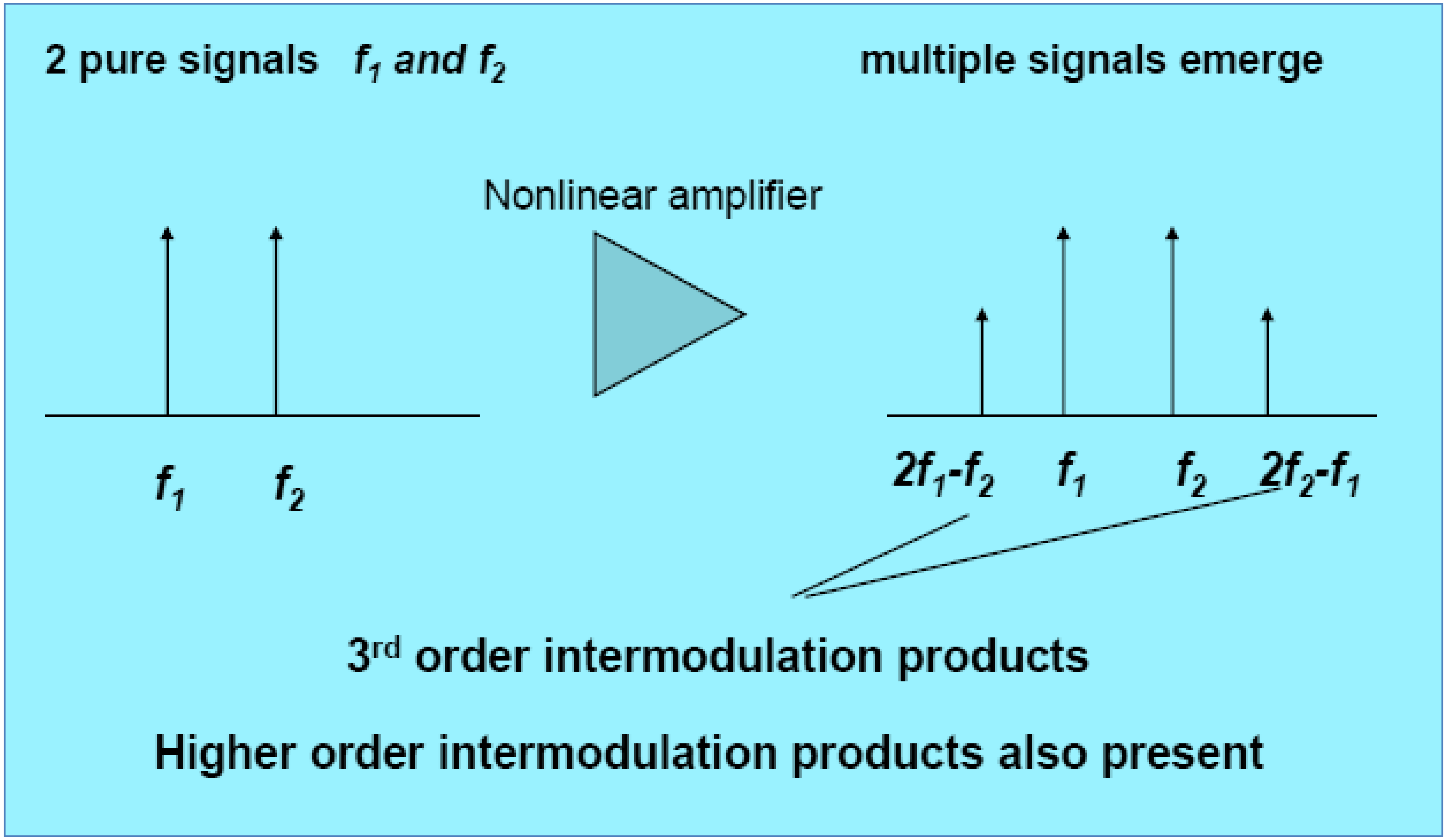
PA nonlinearities

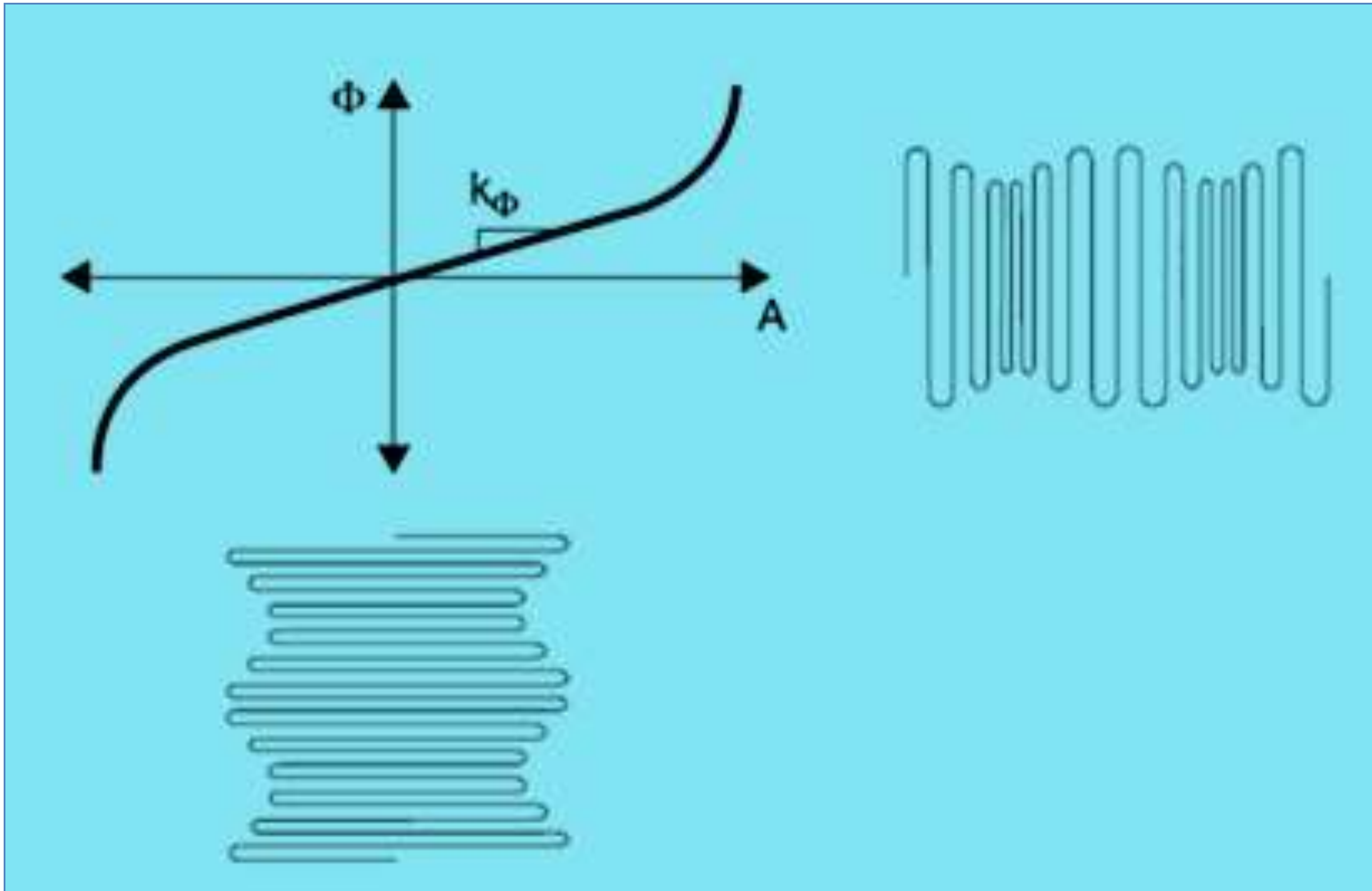
P1dB



3rd order intermodulation products



AM-PM distortion



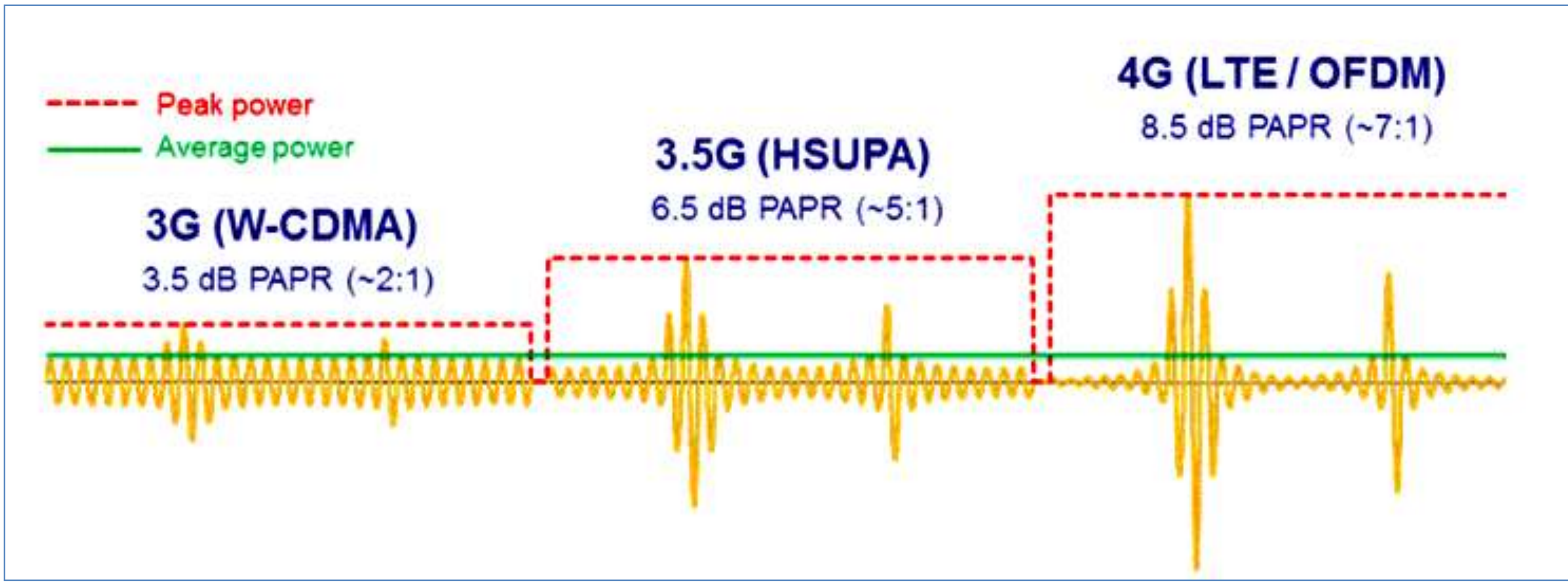
Operate PA in compression for efficiency

- This scheme is only applicable for constant amplitude schemes such as FM and GMSK where there is no amplitude element in the modulation.

- Does not provide high efficiency for high peak to average power levels signals as amplifier will not run in compression all the time.

- Can use digital pre-distortion to provide linearity where some amplitude components are available.

Higher data rates lead to “peakier waveforms

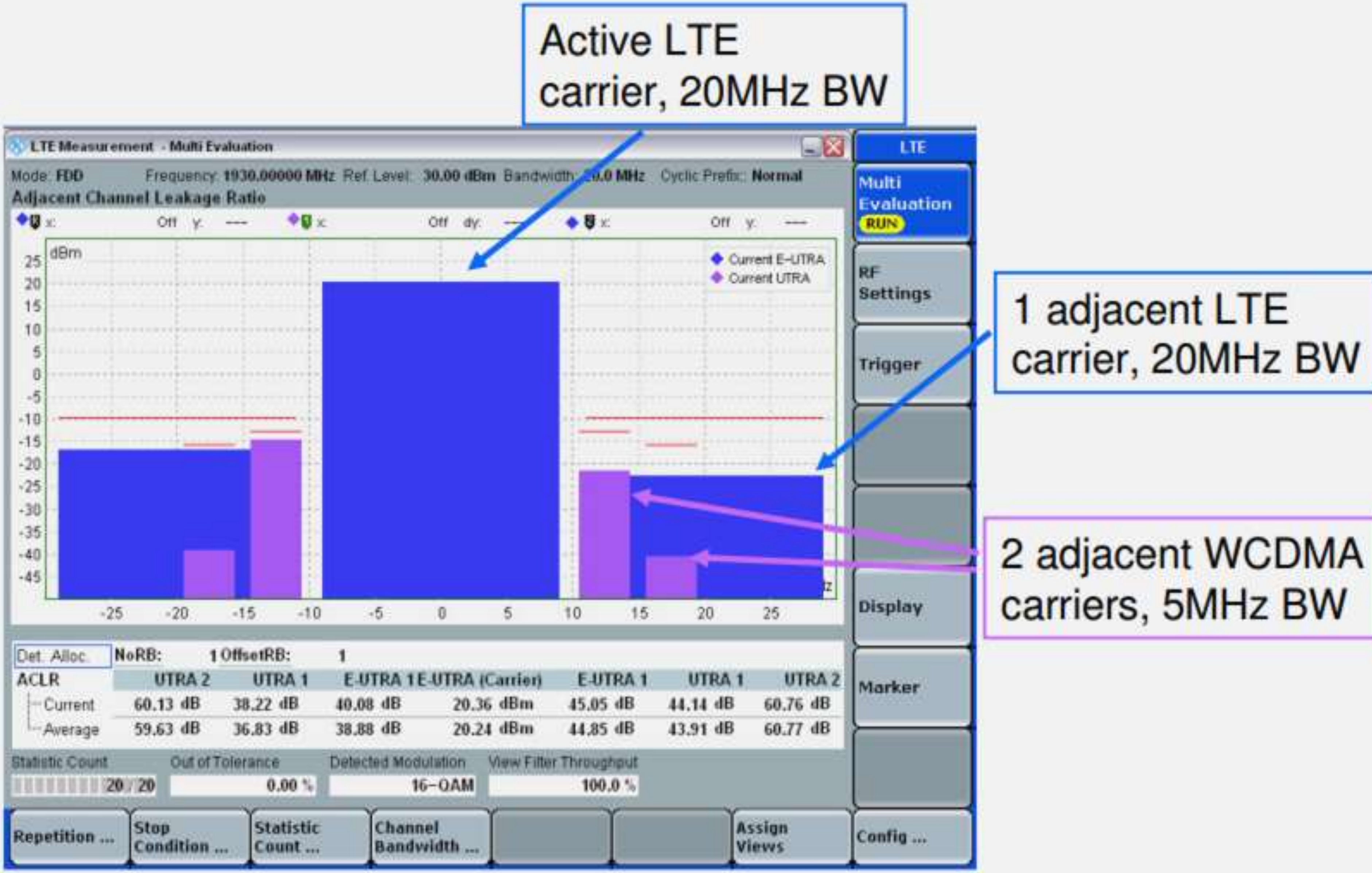


PAPR

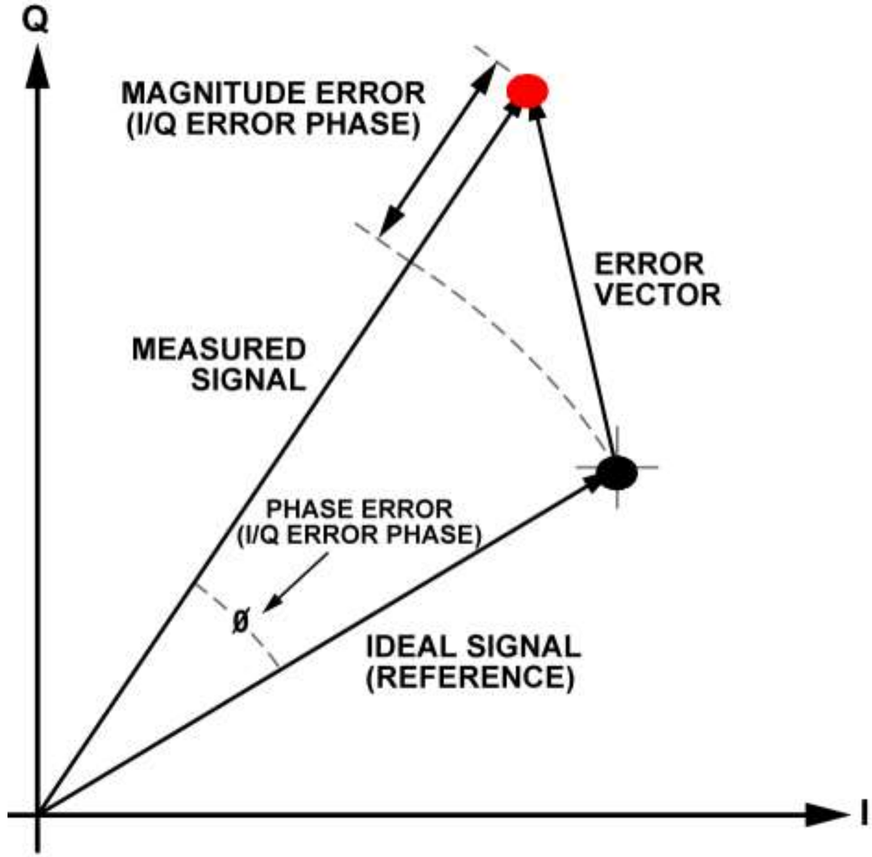
	Standard	Launched	Typ. Carrier BW (MHz)	Typ. Spectral Efficiency (bps/Hz)	Approx. PAPR(dB)
2G cellular	GSM	1991	0.2	0.17	0.0
Digital TV	DVB-H	2007	8	0.28	8.0
2.75G cellular	GSM + EDGE	2003	0.2	0.33	3.5
3G cellular	WCDMA FDD	2001	5	0.51	7.0
Digital TV	DVB-T	1997	8	0.55	8.0
Wi-Fi	IEEE 802.11a/g	2003	20	0.90	9.0
WiMAX	IEEE 802.16d	2004	20	1.20	8.5
Wi-Fi	IEEE 802.11n	2007	20	2.40	9.0
3.5G cellular	HSDPA	2007	5	2.88	8.0
3.9G cellular	LTE	2009	20	8.00	10.0

ACLR/ACPR

- Adjacent Channel Leakage power Ratio (ACLR) is the ratio of the filtered mean power centred on the assigned channel frequency to the filtered mean power centred on an adjacent channel frequency.


Active LTE carrier, 20MHz BW
1 adjacent LTE carrier, 20MHz BW
2 adjacent WCDMA carriers, 5MHz BW
LTE Measurement - Multi Evaluation
Mode: FDD Frequency: 1930.00000 MHz Ref. Level: 30.00 dBm Bandwidth: 20.0 MHz Cyclic Prefix: Normal
Adjacent Channel Leakage Ratio
dBm
Current E-UTRA
Current UTRA
Det. Alloc. NoRB: 1 OffsetRB: 1
ACLR UTRA 2 UTRA 1 E-UTRA 1 E-UTRA (Carrier) E-UTRA 1 UTRA 1 UTRA 2
Current 60.13 dB 38.22 dB 40.08 dB 20.36 dBm 45.05 dB 44.14 dB 60.76 dB
Average 59.63 dB 36.83 dB 38.88 dB 20.24 dBm 44.85 dB 43.91 dB 60.77 dB
Statistic Count 20 / 20
Out of Tolerance 0.00 %
Detected Modulation 16-QAM
View Filter Throughput 100.0 %
LTE
Multi Evaluation RUN
RF Settings
Trigger
Display
Marker
Config ...
Repetition ...
Stop Condition ...
Statistic Count ...
Channel Bandwidth ...
Assign Views

EVM



The EVM result is defined as the square root of the ratio of the mean error vector power to the mean reference power expressed in percent.

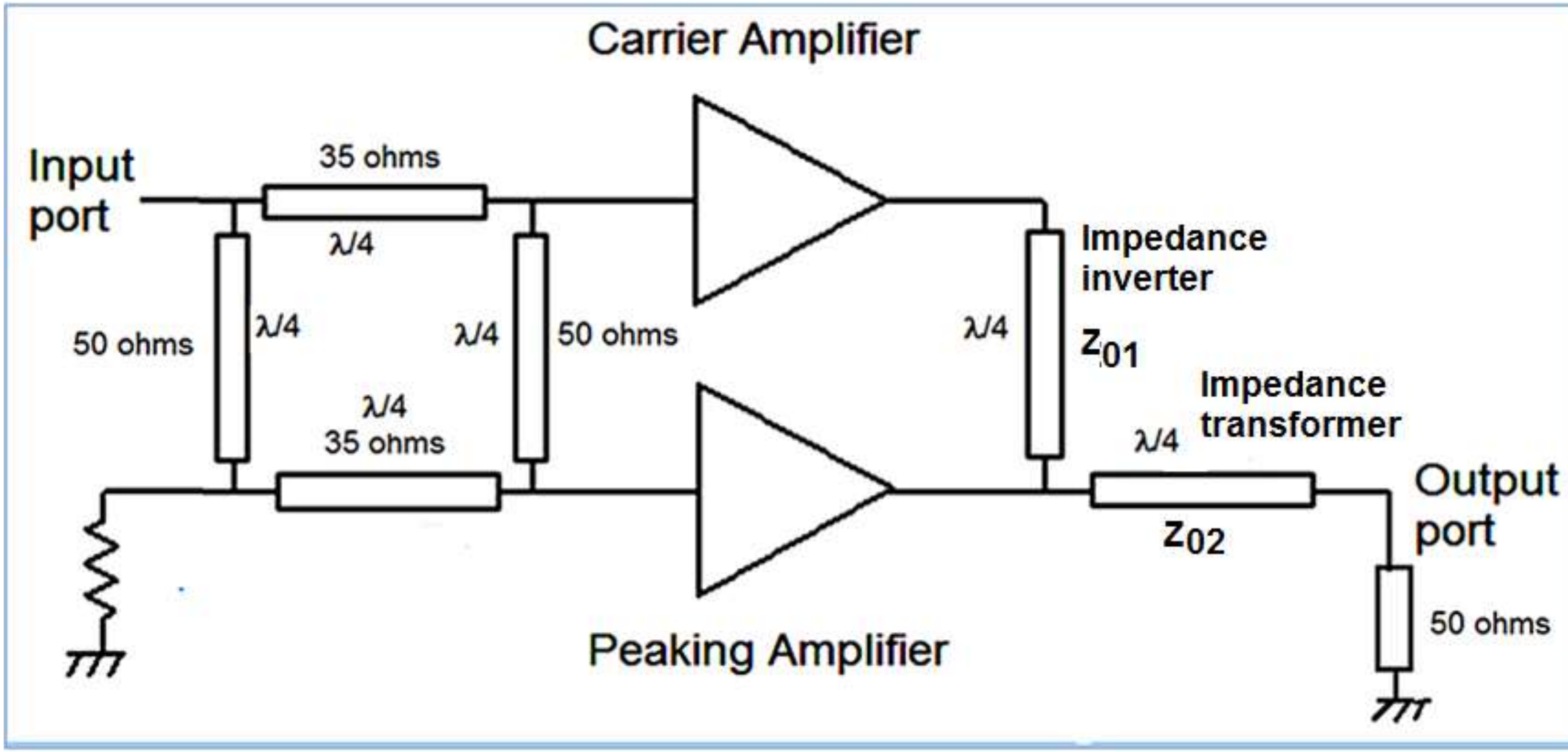
The **error vector magnitude** or **EVM** (sometimes also called **receive constellation error** or **RCE**) is a measure used to quantify the performance of a digital radio transmitter or receiver.

PA efficiency enhancement techniques

DC Tracking

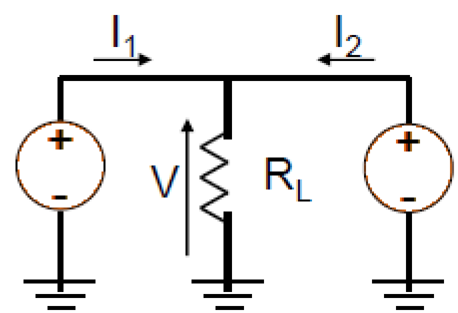
- The PA voltage is adjusted in line with the modulation level, often using protocol commanded power supply voltage level changes
- Deals only with average power
- Does not address broadband issues
- Does not address instantaneous peaks and troughs

Doherty amplifier

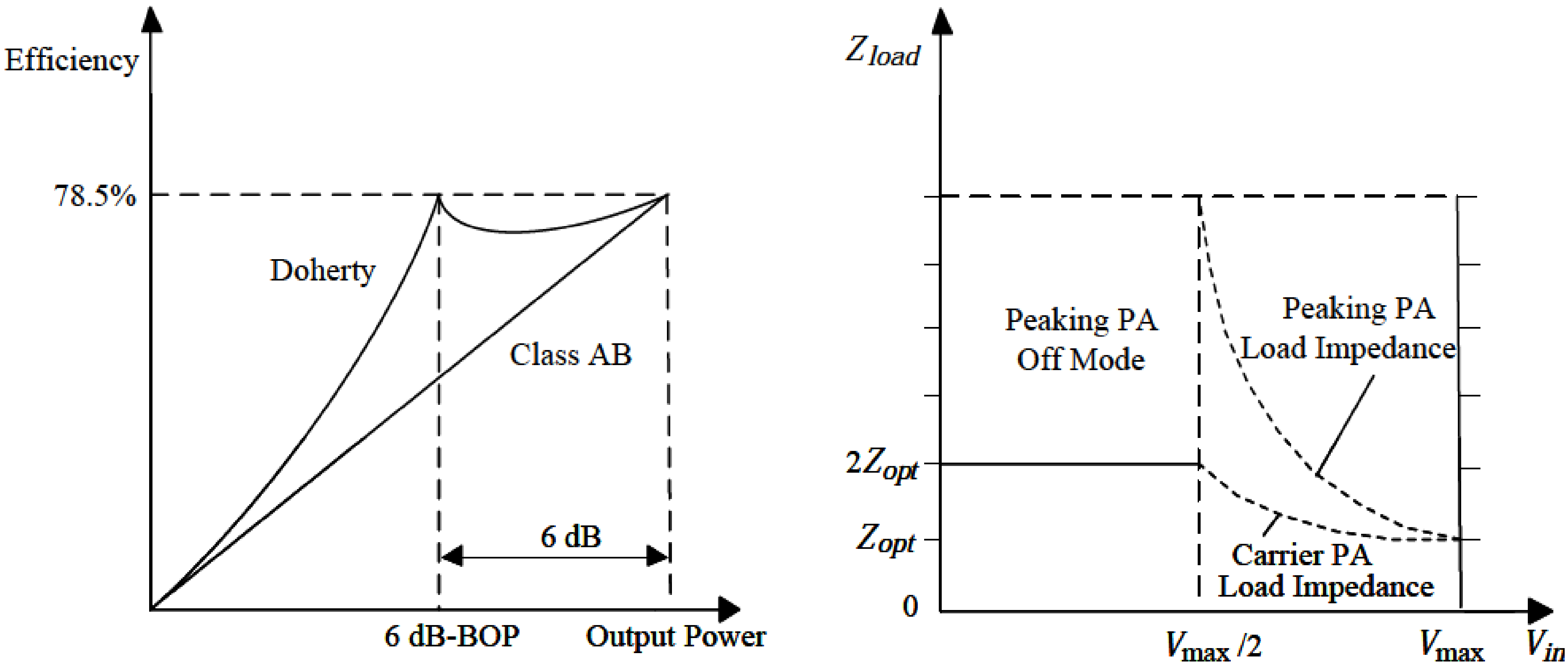


Load modulation characteristic of Doherty amplifier

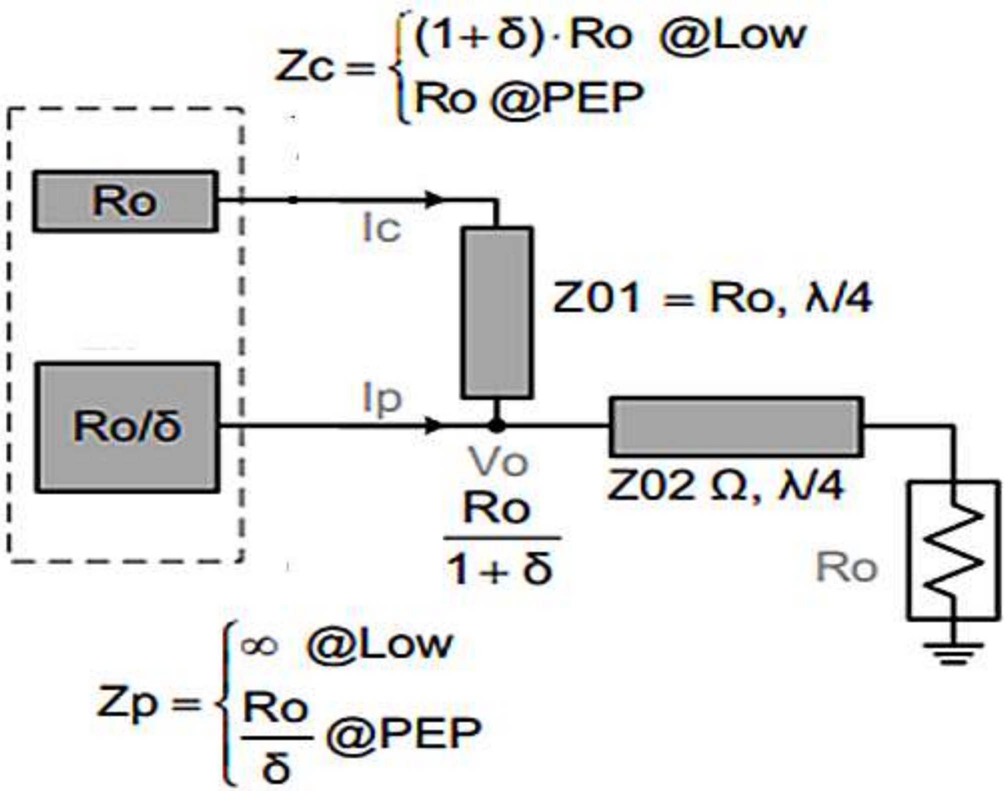
- For a low input drive signal, the peaking amplifier is off and looks like an open circuit. As the drive level increases, the peaking amplifier begins to conduct more and more, feeding current into the output circuit. The carrier amplifier output sees its load impedance drop as input RF power increases. i.e. the peaking amplifier provides a form of fundamental active load pulling to the carrier amplifier



Symmetric Doherty PA



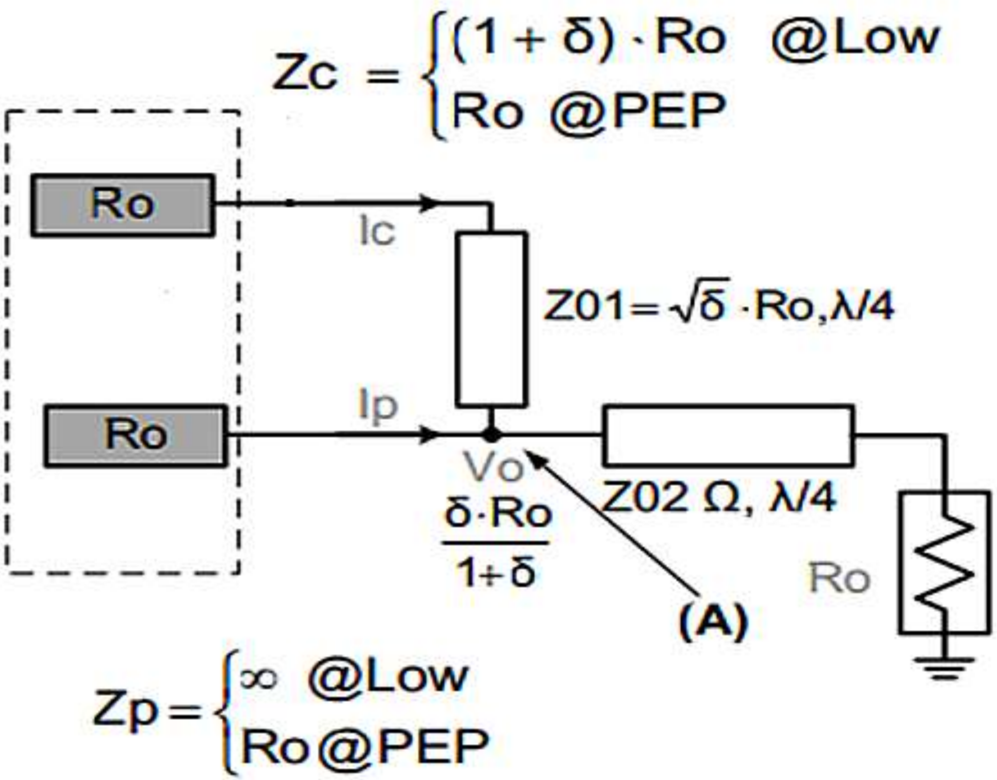
Output combiner variant 1



where δ is the size ratio of the peaking PA over the carrier PA.

The output impedance of the peaking PA is R_0/δ at the peak power region

Output combiner variant 2



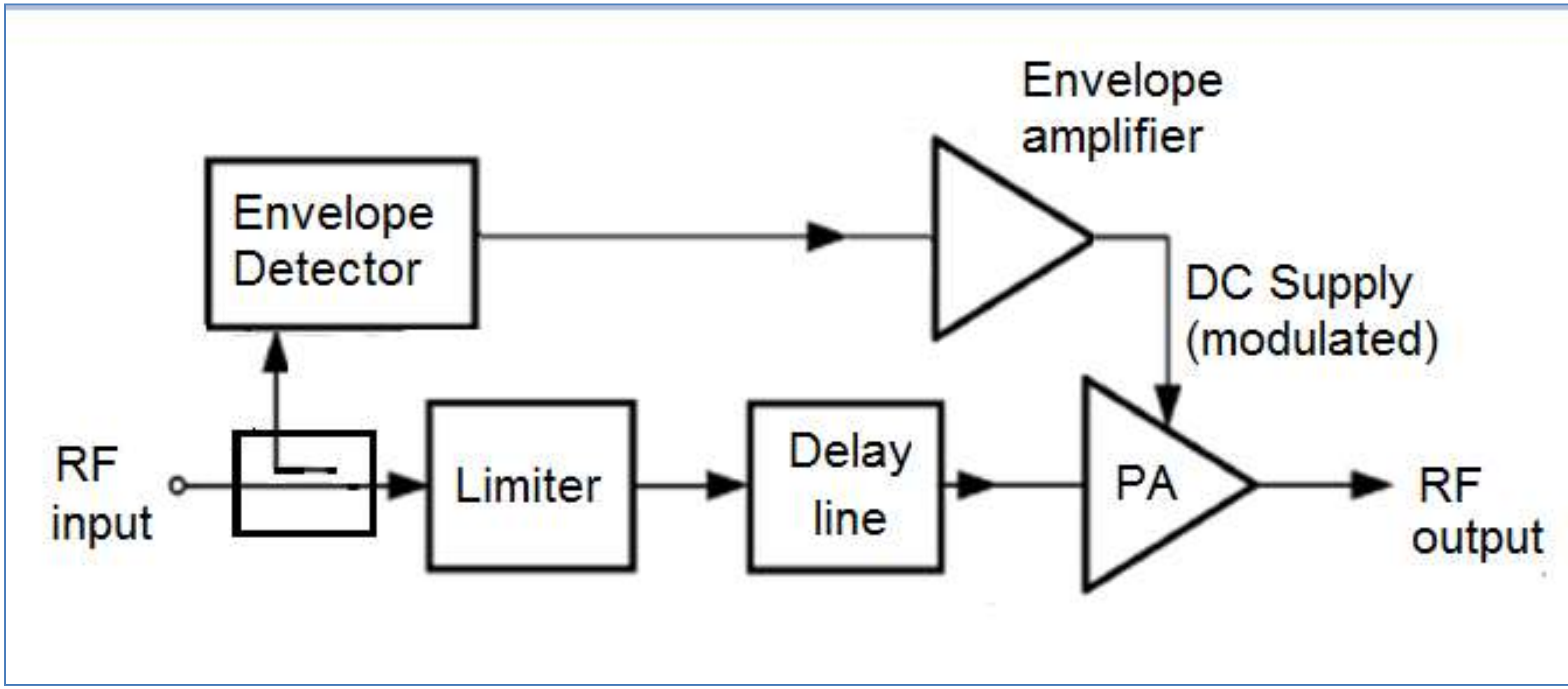
$$Z_{01} = \sqrt{\delta} \cdot R_0,$$

$$Z_{02} = \sqrt{\frac{\delta}{1+\delta}} \cdot R_0$$

where δ is the size ratio of the peaking PA over the carrier PA.

The output impedance of the peaking PAs becomes R_0 (same as load)

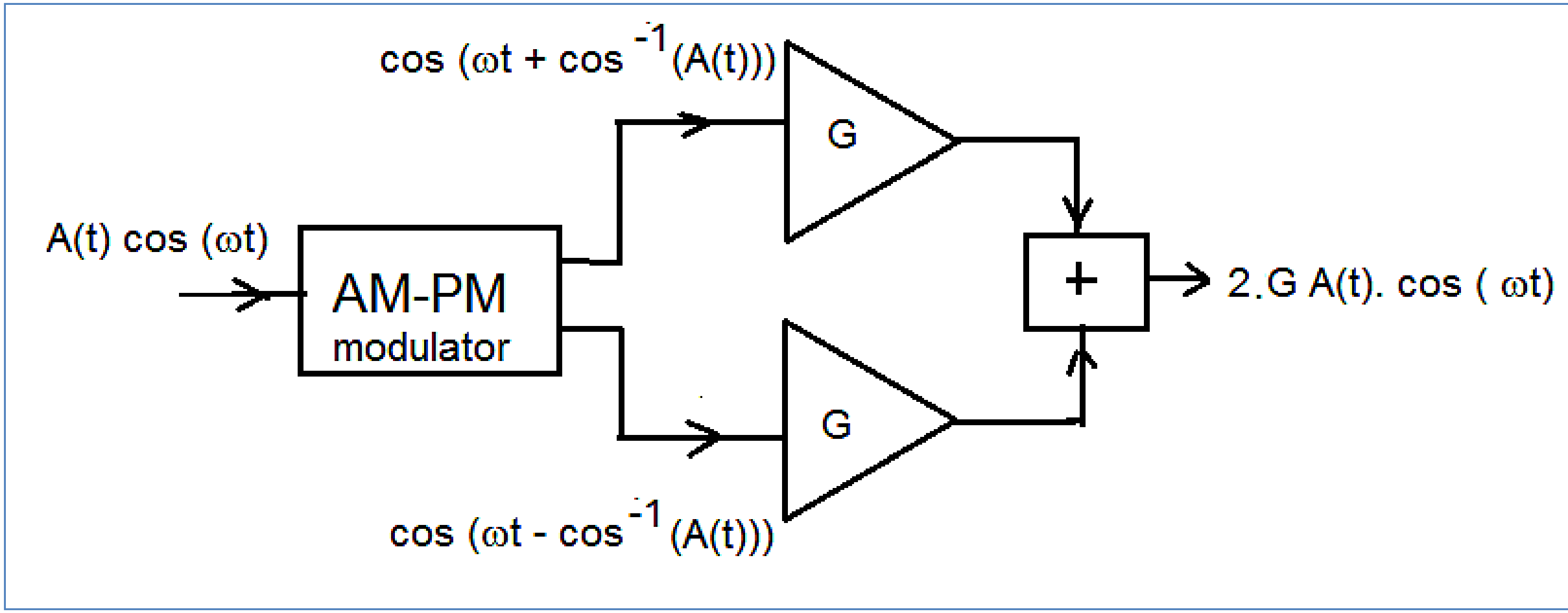
Envelope elimination and restoration



The limiter eliminates the possibility of AM-PM distortion in the PA which can be a saturated /switching amplifier. The envelope amplitude is recovered at the output using a DC supply modulated by output of envelope detector

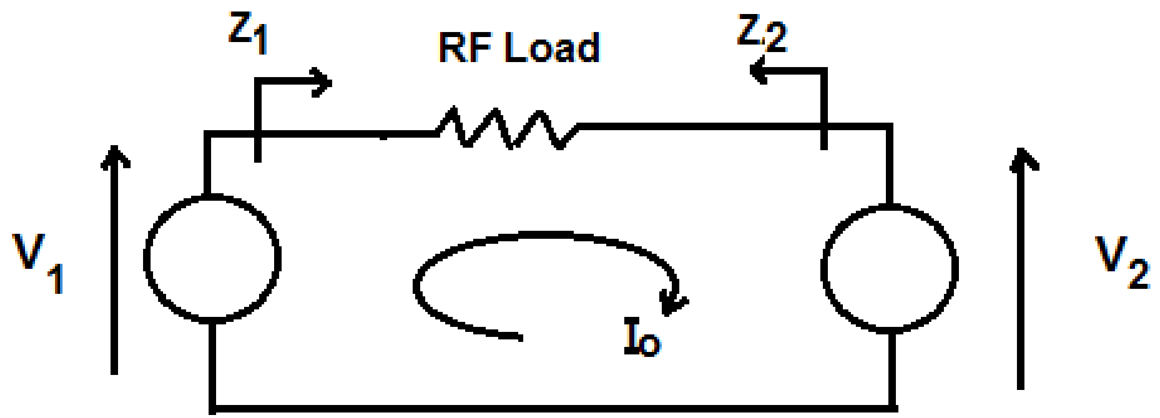
Chireix's Outphasing amplifier or LINC

LINC (**LI**near amplification with **N**onlinear **C**omponents)



A complex modulated signal is split into **two constant envelope, phase-modulated signals** that are amplified by **high efficiency nonlinear (saturated) amplifiers** and then combined at the output. The range of load modulation for both amplifiers extends from R/2 to ∞.

Load-pulling in Outphasing amplifier

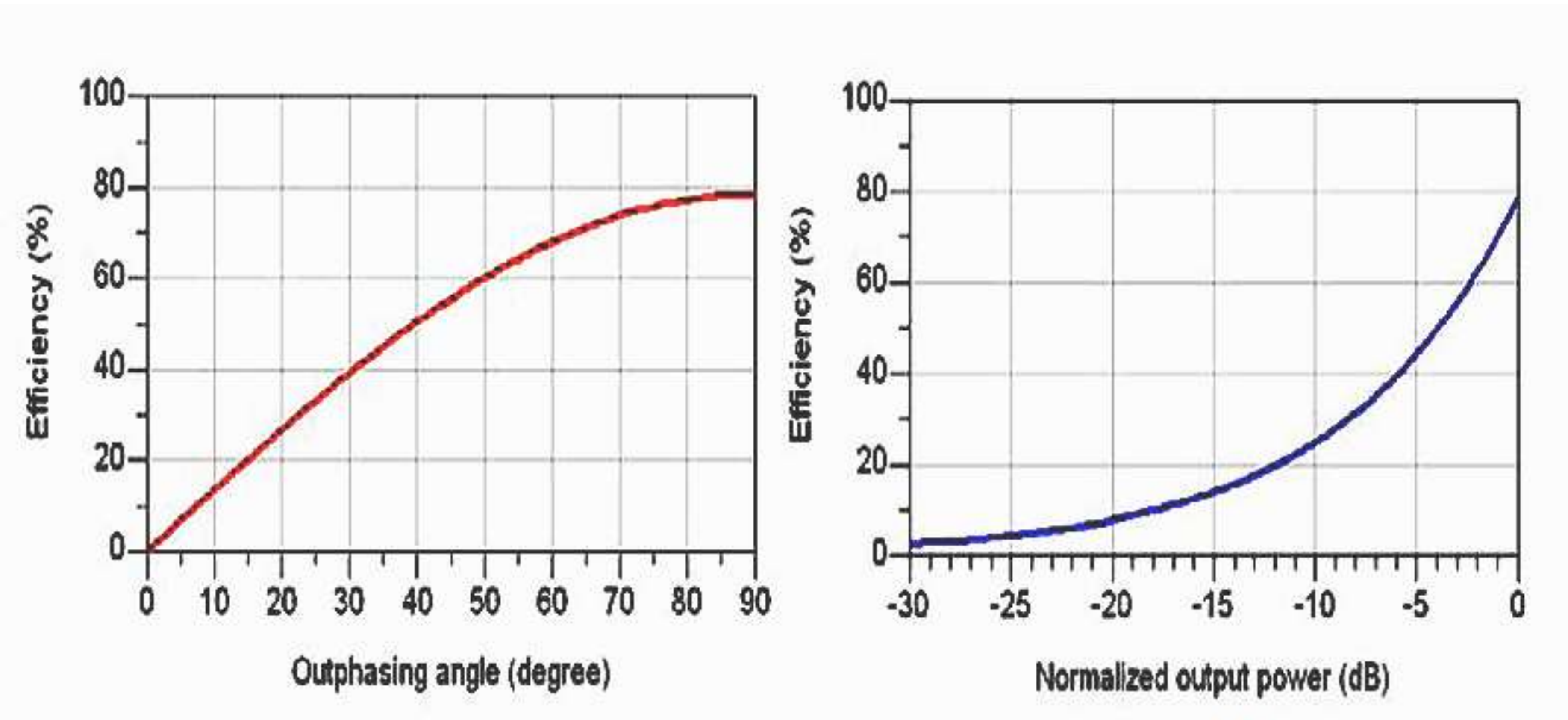


$$\mathbf{Z}_1 = \frac{\mathbf{V}_1}{\frac{\mathbf{V}_1 - \mathbf{V}_2}{R_L}} = \frac{\cos\phi + j\sin\phi}{2j\sin\phi} R_L = \frac{R_L}{2}(1 - j\cot\phi)$$

$$\mathbf{Z}_2 = \frac{\mathbf{V}_2}{-(\frac{\mathbf{V}_1 - \mathbf{V}_2}{R_L})} = \frac{\cos\phi - j\sin\phi}{-2j\sin\phi} R_L = \frac{R_L}{2}(1 + j\cot\phi)$$

where $\phi = \sin^{-1} A(t)$

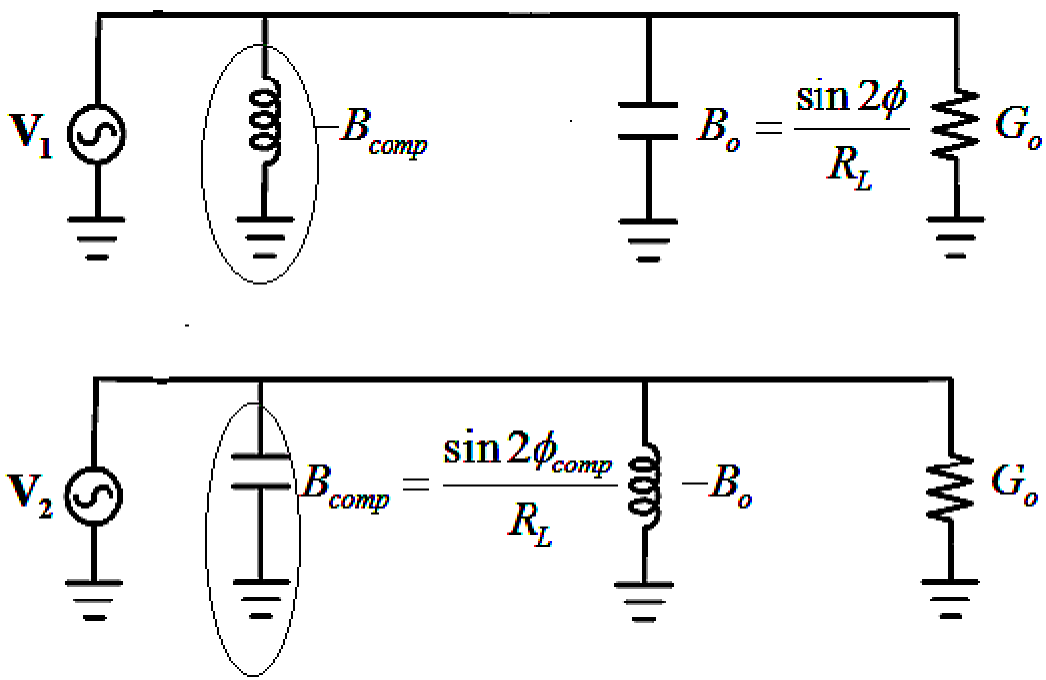
Theoretical efficiency of outphasing amplifier



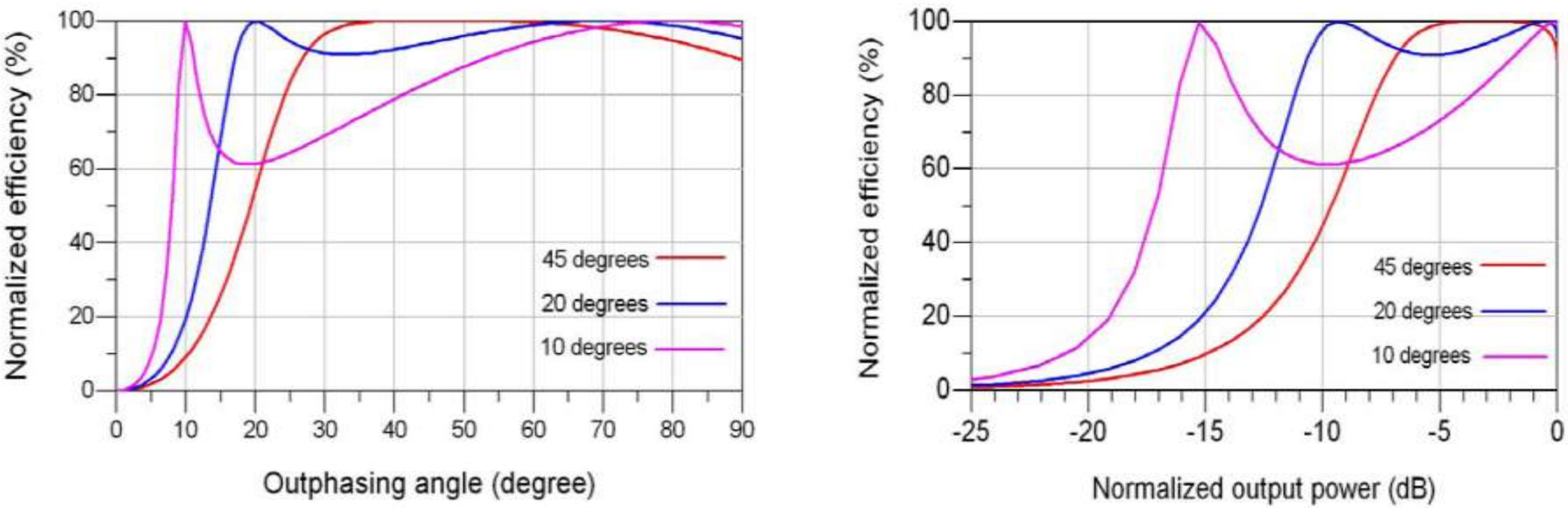
Efficiency drops as the outphasing angle ϕ decreases

Load compensation

- Load compensation method compensates for the susceptance component by adding a proper shunt reactance

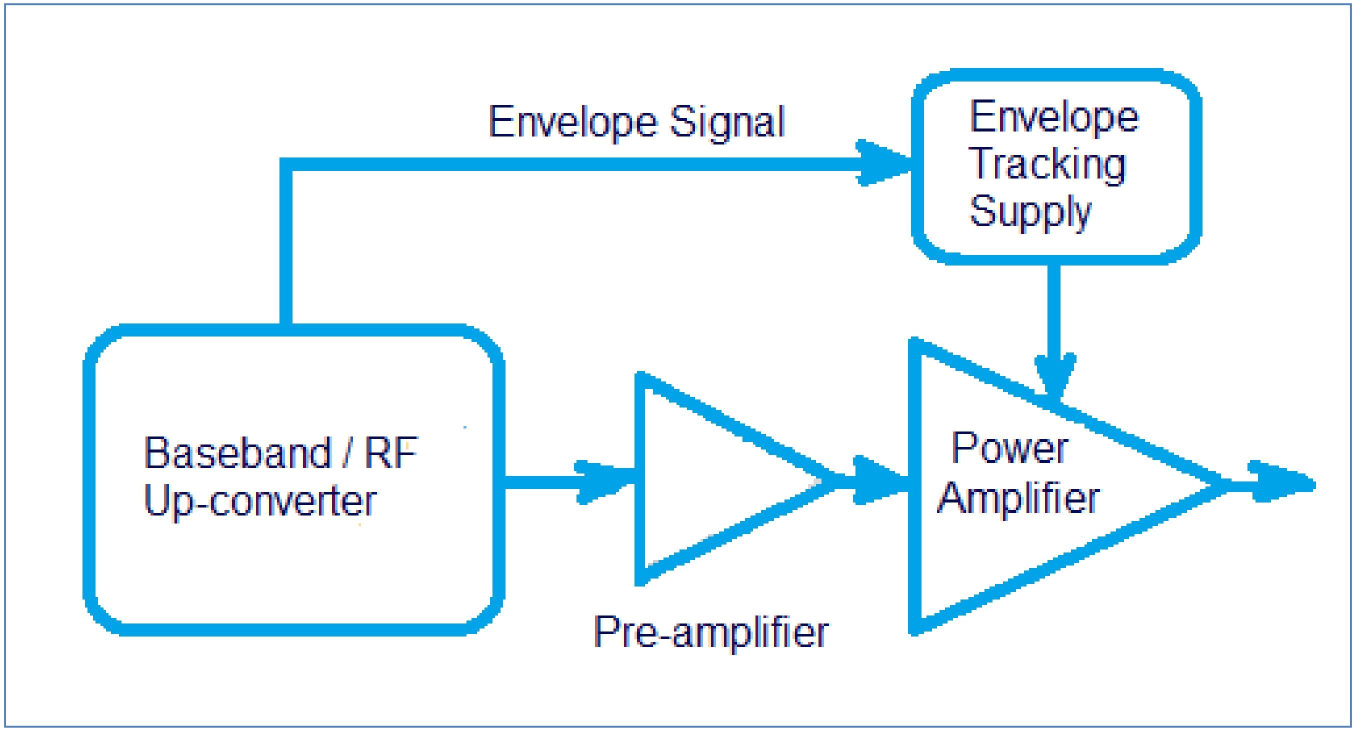


Efficiency improvement with load compensation

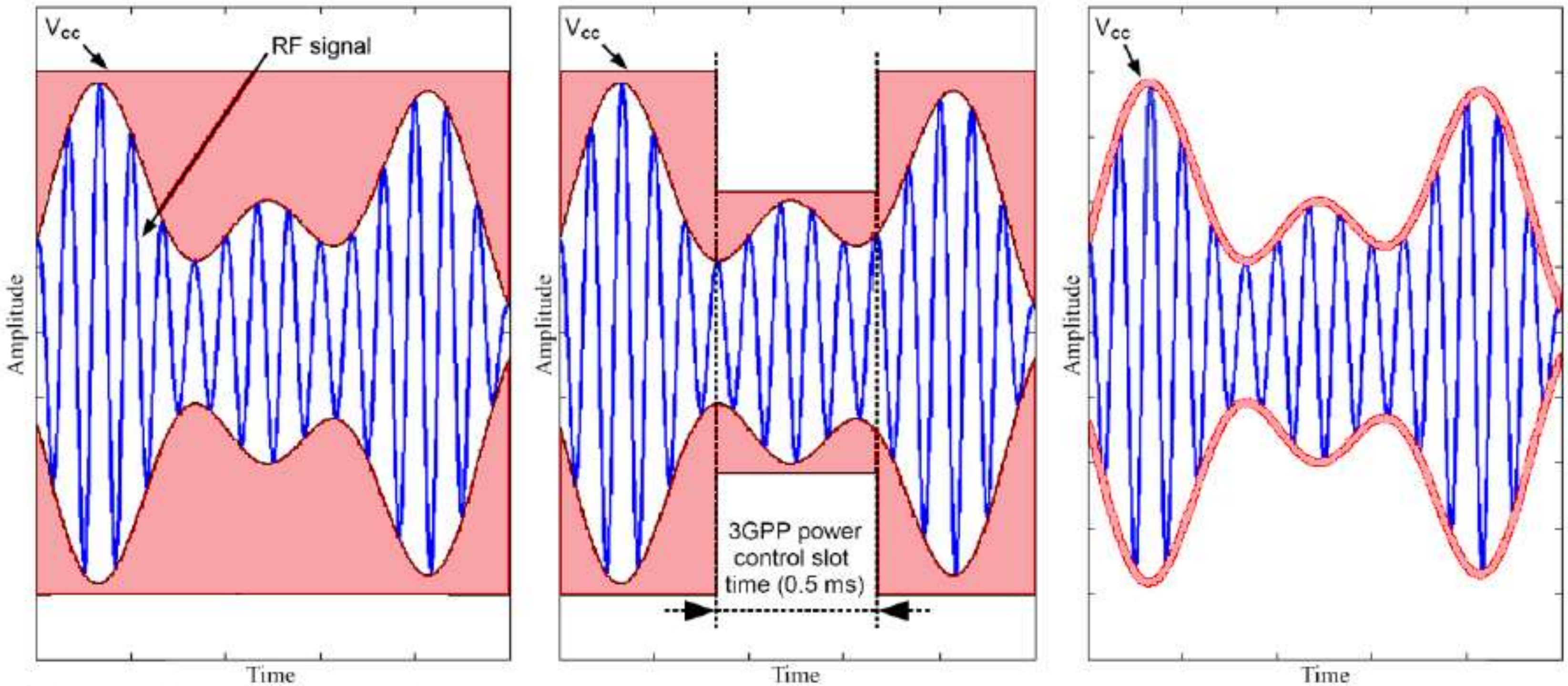


Better overall efficiency can be achieved if the load can be adaptively compensated according to the change of the outphasing angle.

Envelope Tracking System



Average power and envelope tracking



Constant collector voltage

Average power tracking

Envelope tracking

Envelope tracking -variants

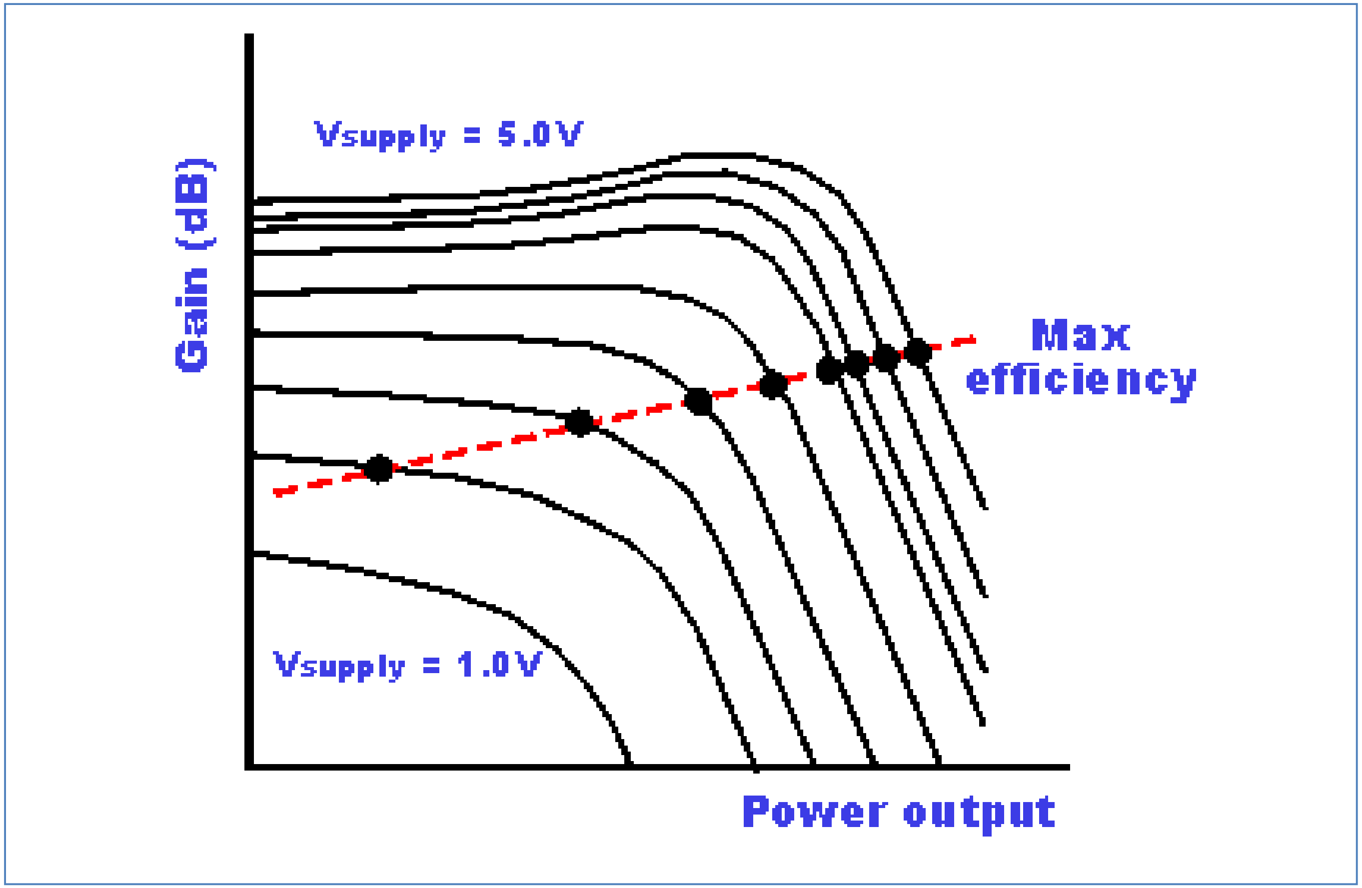
- When the supply voltage tracks the instantaneous envelope modulation signal, it is called Wide Bandwidth ET (WBET)
- When the supply voltage tracks the long-term average of the input envelope power, it is called Average ET (AET)
- When the supply voltage switches to different step levels according to the input envelope power, it is called Step ET (SET)

Envelope Tracking

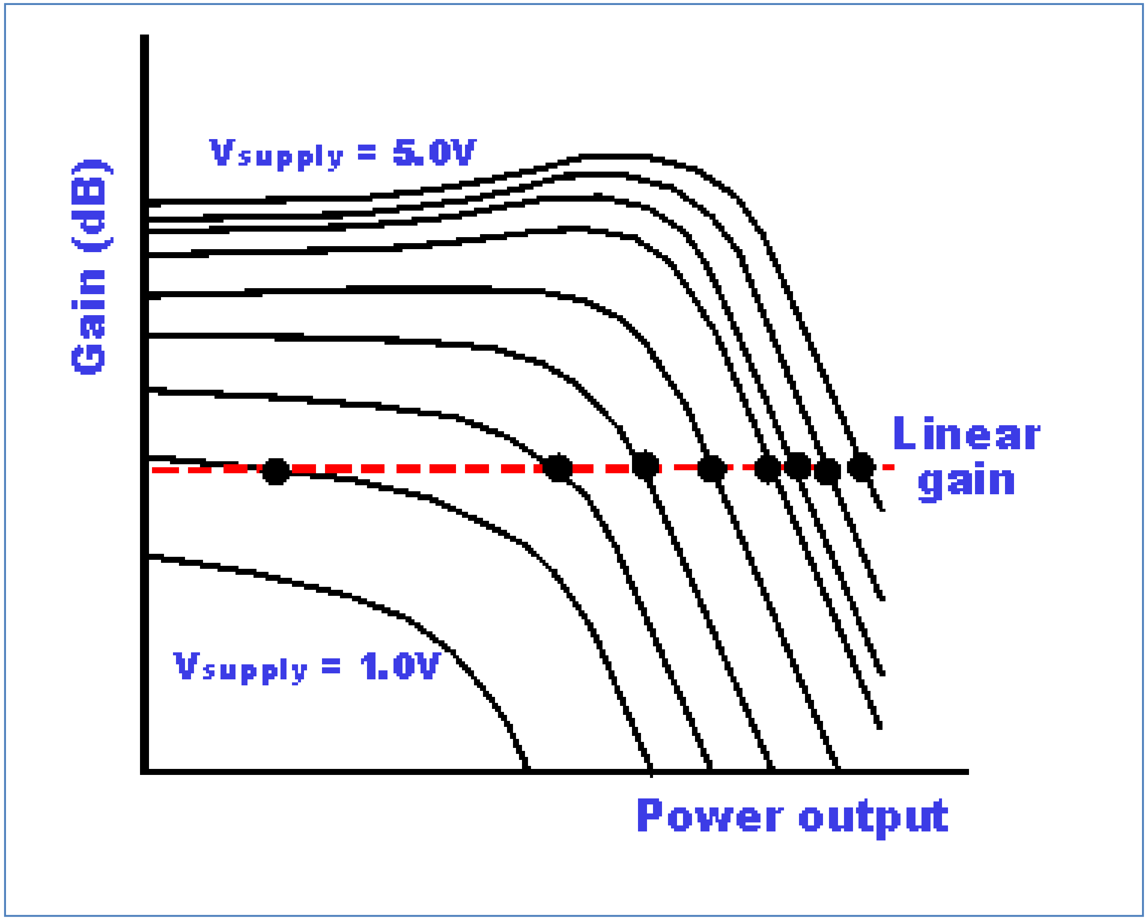
$$\boxed{Envelope = \sqrt{I^2 + Q^2}}$$

Envelope tracking requires very fast , high-bandwidth power supply. A supply bandwidth of 40-60 MHz is required to track a 20 MHz RF channel.

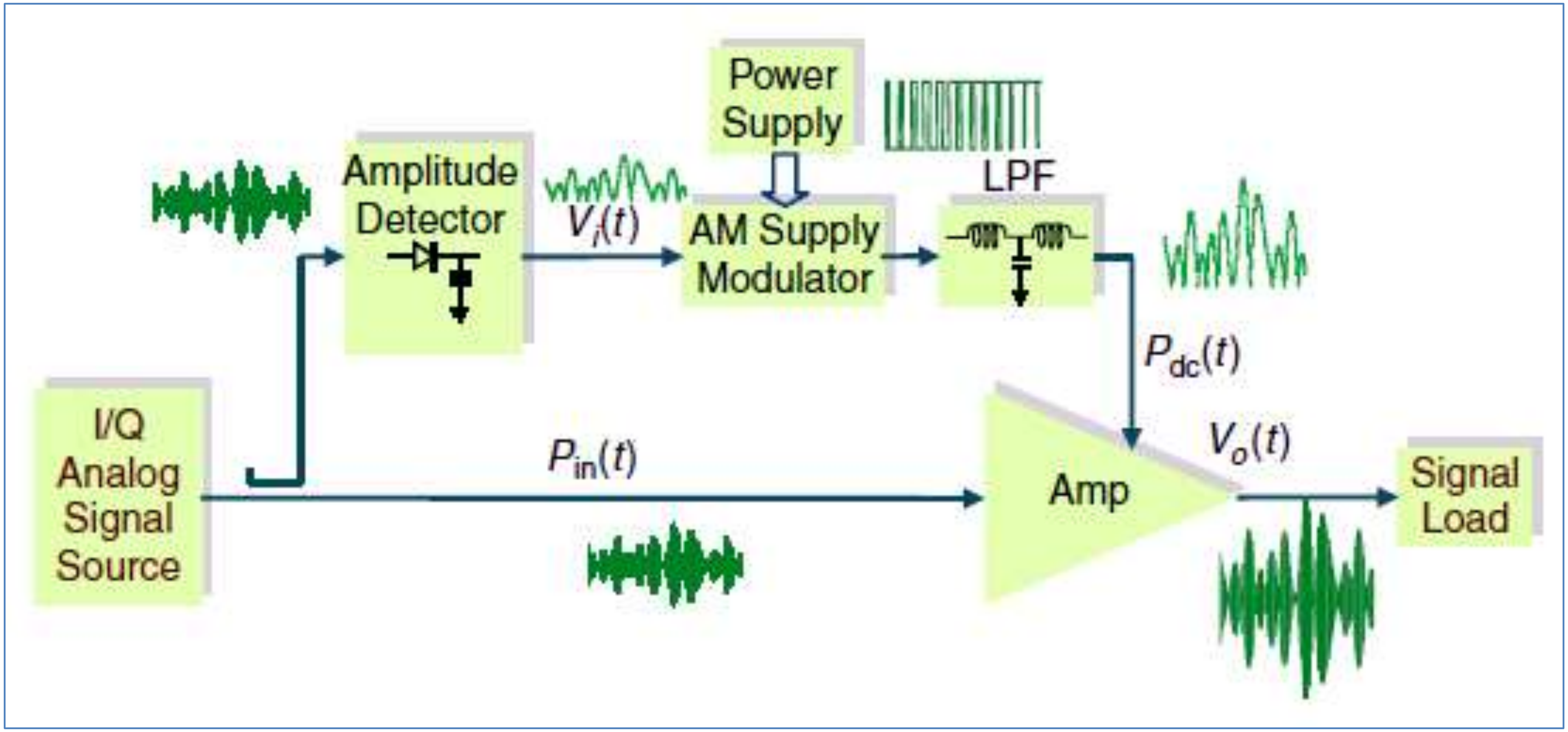
Maximum efficiency envelope tracking



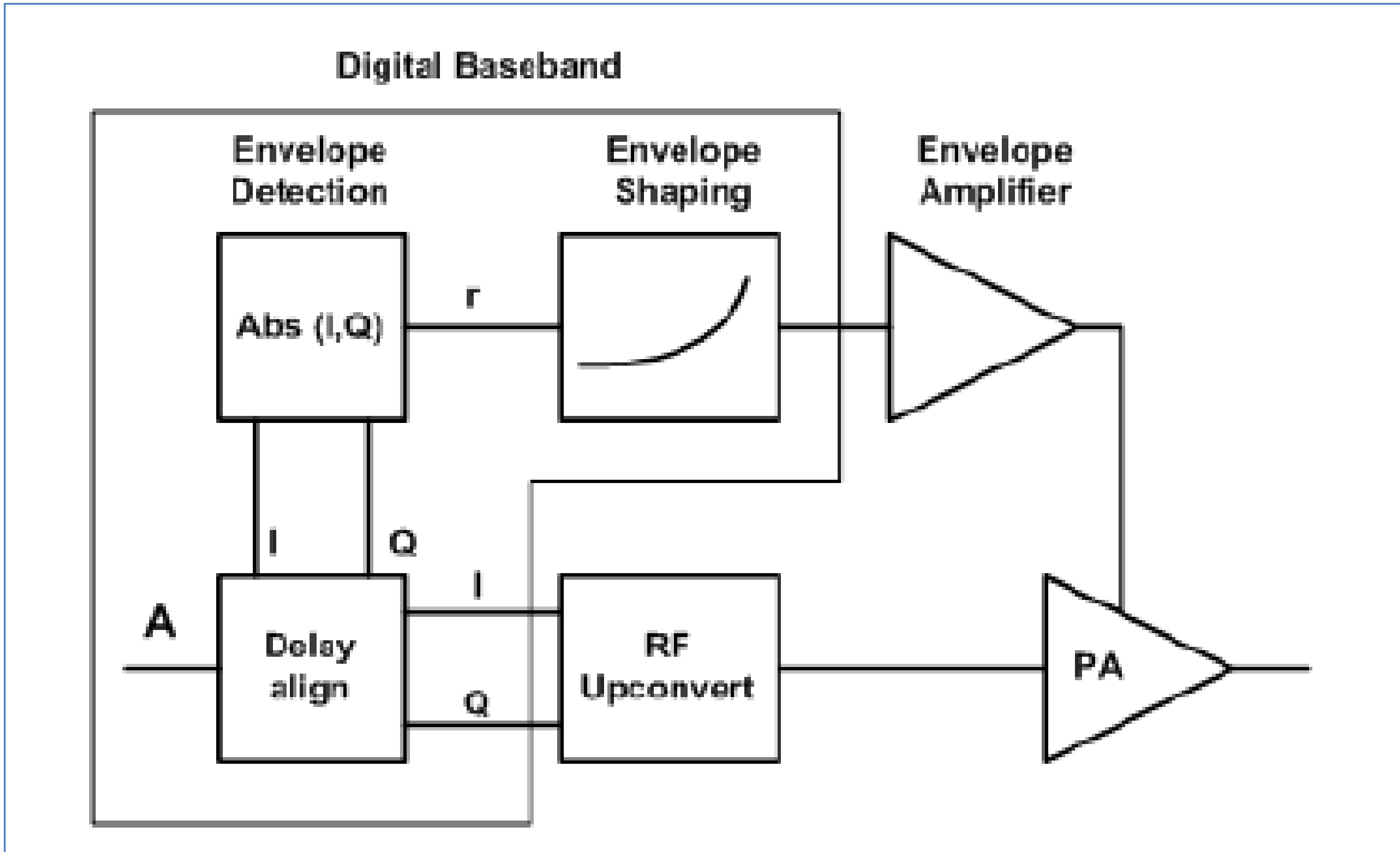
Linear gain envelope tracking



Analog ET system



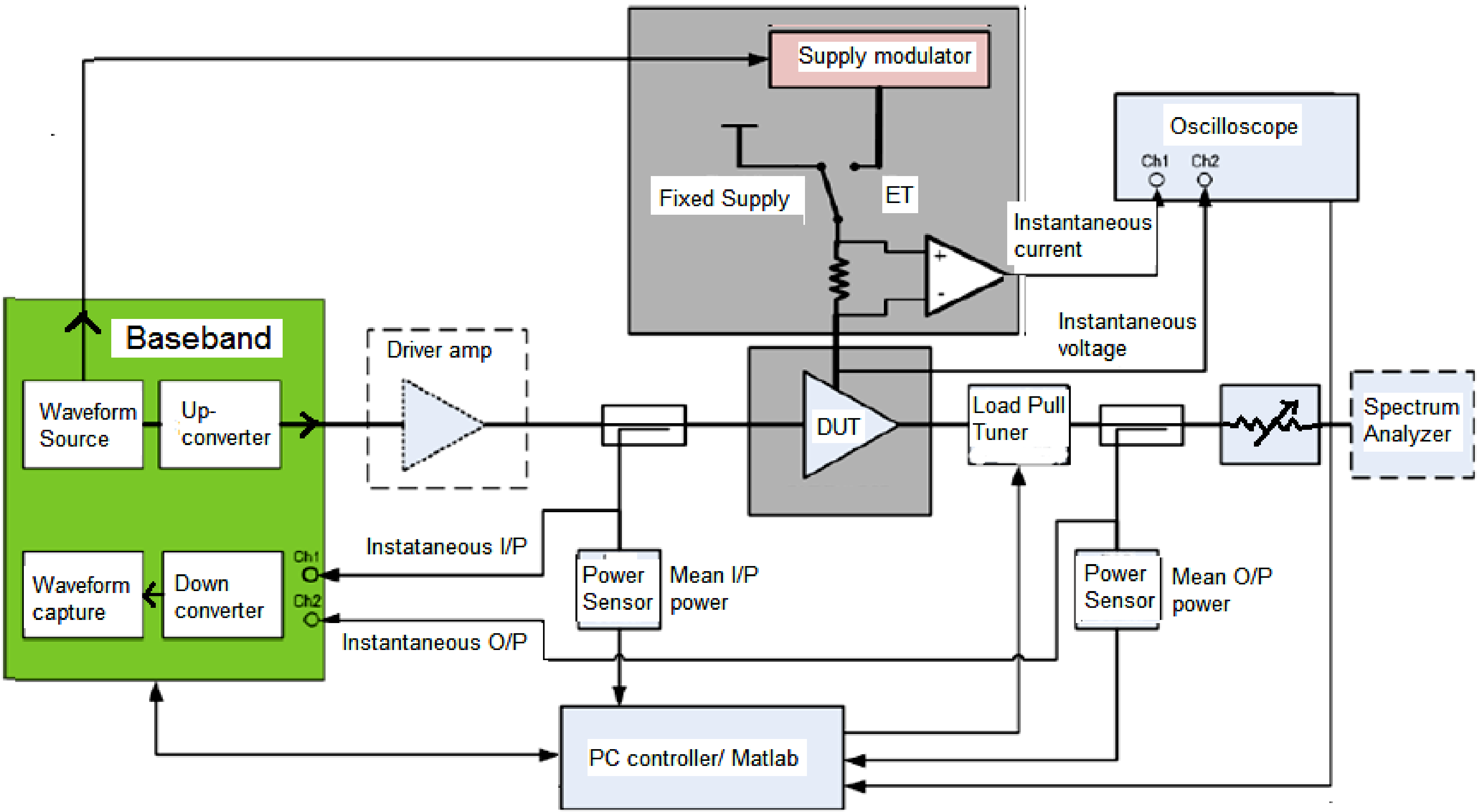
Digital ET system



PA Characterization methods

Test methodology	PA current measurement	Supply impedance	Supply bandwidth requirements	Correlation with ET operation	Parameters measured
Swept CW testing	Bench PSU	Low (decoupling Capacitor)	Low (Bench PSU)	Poor, due to PA die heating	Gain (AM:AM), Efficiency
Pulsed RF /DC testing	Instrumentation grade current probe, ~5 us resolution	Low (decoupling Capacitor)	Low (Bench PSU)	Good, if short pulses (~10 us, 10% duty cycle).	Gain (AM:AM), Efficiency
Dynamic supply modulation	Challenging – high BW with high common mode voltage current sense	Requires low impedance dynamic supply (no decoupling)	High (~60 MHz BW)	Excellent	Gain (AM:AM), Phase (AM:PM), Efficiency

ET PA Characterization bench



ET architectures

- A simple ET architecture could be a solution that it only tracks the envelope signal coarsely with a fast DC-DC buck converter.

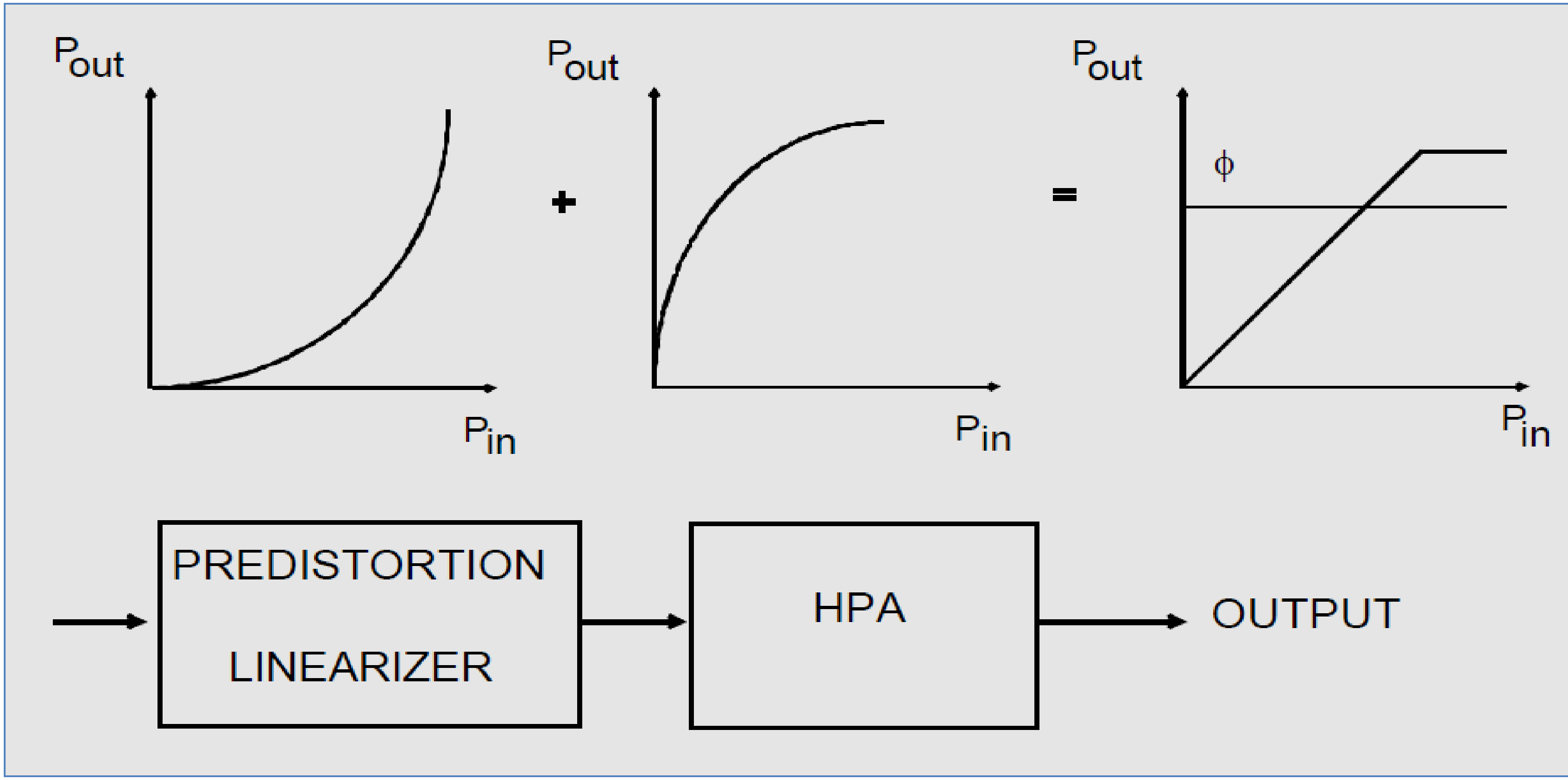
- A complex solution could include very accurate and fast envelope signal tracking

Application of ET

- Envelope tracking can be applied to the design of cellular terminals operating at 0.5W to broadcast transmission systems operating at 10KW.

PA Linearization methods

Predistortion linearizers



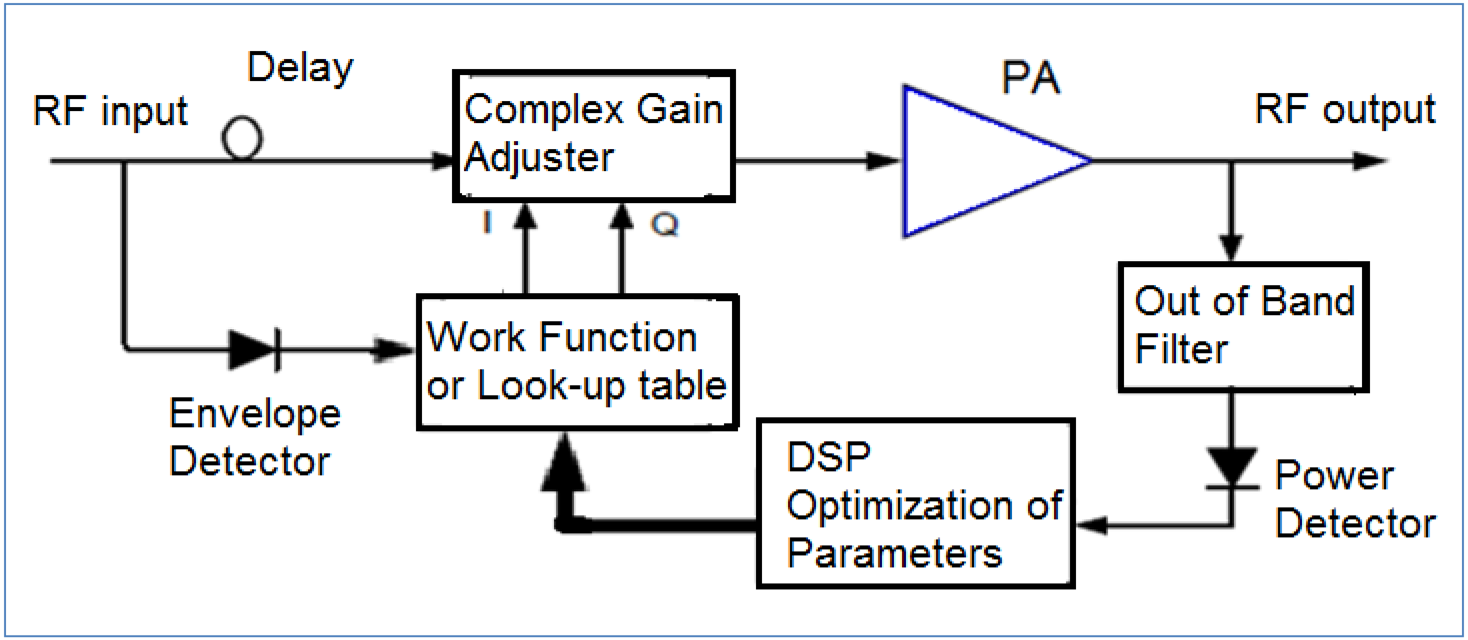
Predistortion linearizers generate a response opposite to an PA's response in magnitude and phase

RF predistortion techniques

There are three distinct RF predistortion techniques.

- The **work function-based approach** utilizes a low-order polynomial to fit the AM/AM and AM/PM characteristics of the power amplifier.
- The **look-up table technique** fits the power amplifier's characteristics more accurately. However, it requires a more sophisticated adaptation technique.
- The **analog nonlinearity technique** uses diodes to generate IM distortion. This IM distortion is then phased and attenuated to make it anti-phase with the distortion created by the power amplifier.

RF Predistorter



The predistorter consists of a complex gain adjuster that controls the amplitude and phase of the input signal. The extent of predistortion is controlled by two nonlinear work functions that interpolate the AM/AM and AM/PM nonlinearities of the power amplifier.

Advantages of RF-based predistortion

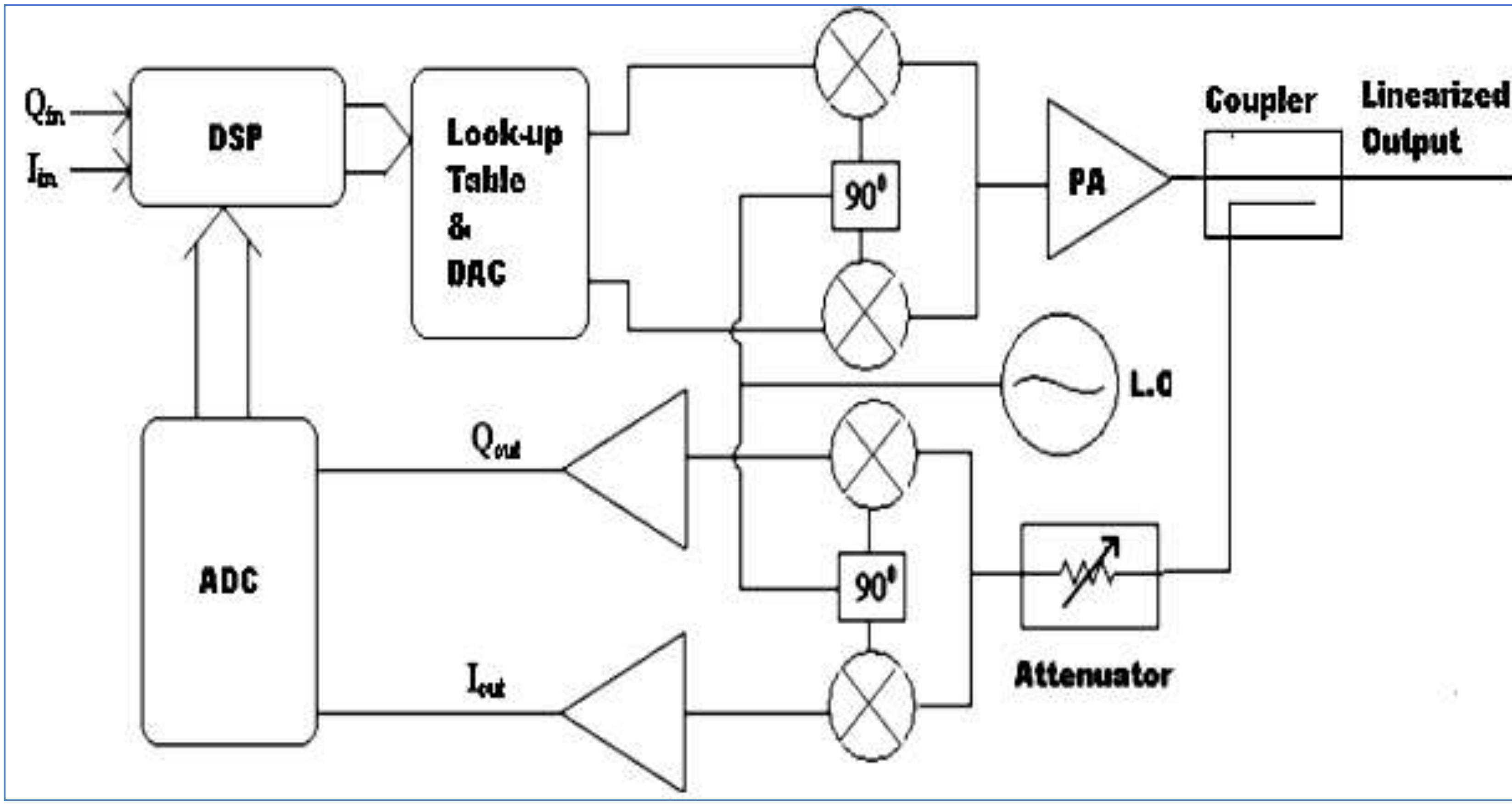
- Correction is applied before the power amplifier where insertion loss is less critical.
- The correction architecture has a moderate bandwidth.

Digital predistortion techniques

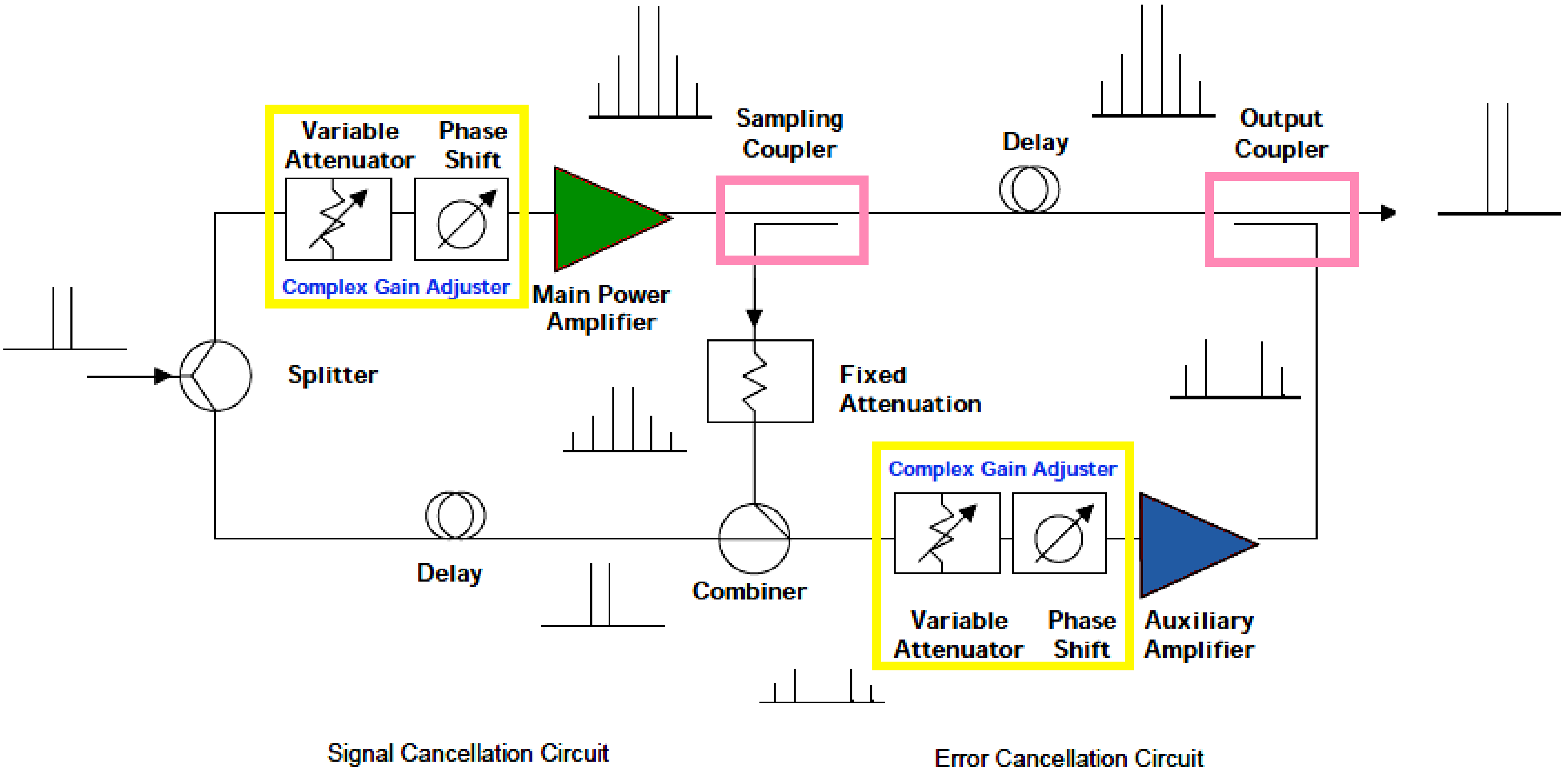
- Complex vector mapping look up table (LUT), Complex Gain LUT and Cartesian feedback techniques falls under the class of **static predistortion technique**

- Secant Method and Linear Convergence Methods are associated with **adaptation- based predistortion technique**

Digital adaptive predistortion



Feedforward amplifier

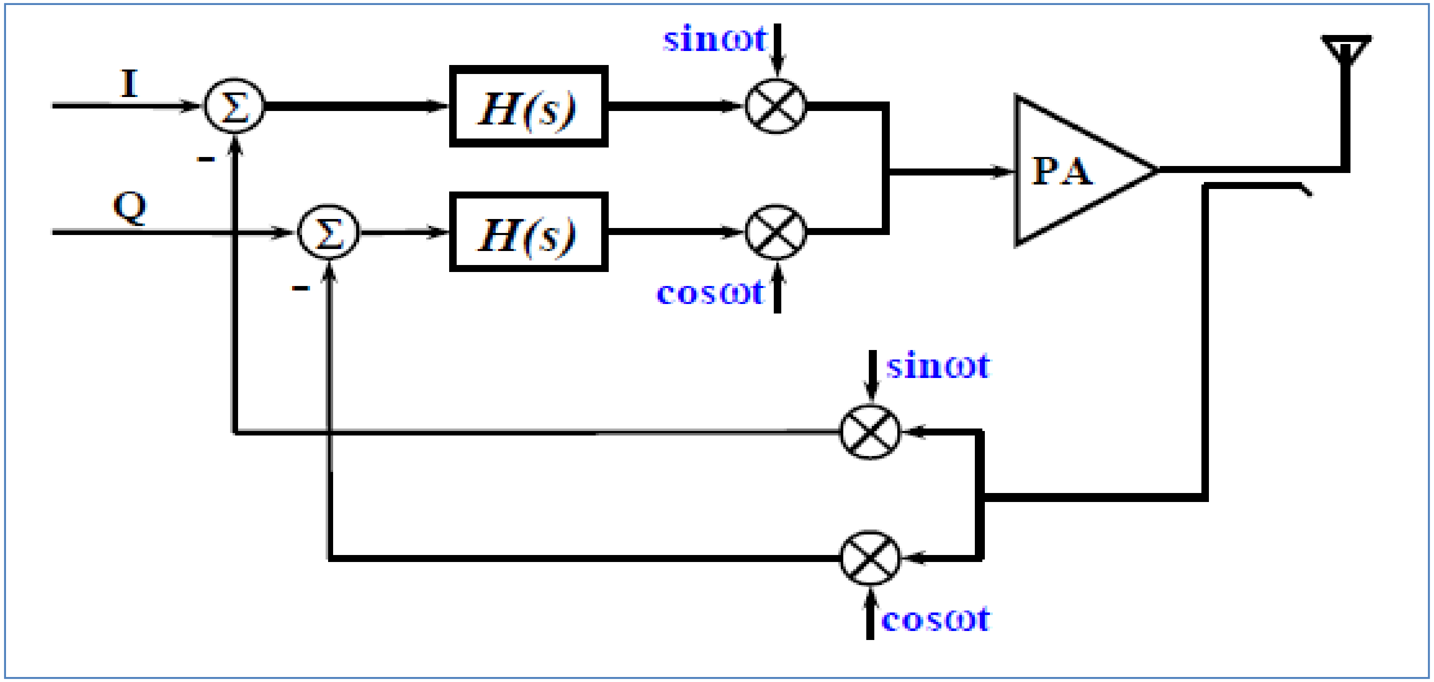


Adaptive Feedforward Linearization

- **Feedforward linearization simultaneously offers wide bandwidth and good IMD suppression**

- High complexity
- Automatic adaptation is essential to maintain performance

Cartesian feedback



The system block H(s) represents the loop driver amplifiers, which provide the loop gain as well as the dynamics introduced by the compensation strategy. Cartesian feedback has relatively low complexity, offers reasonable IMD suppression.
Stability considerations limit the bandwidth to a few hundred KHz

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