Topics in Communications.

"Microwaves"

Spring 2017/2018

Dr. Yanal Al-Faouri

By: Mohammad Abu Hashya.

* * *

* Required Subjects in this CHAPTER:

- Diode (Varactor & schottky) => IL , Zd < reverse & forword , shunt & series configuration.
- The Job of R, L & C in certain circuit (for multiple choice questions).
- Transistor => [S] parameter, threshold freq.

Microwave Engineering

Chapter 11

Dr. Yanal Faouri

Email: y.faouri@ju.edu.jo

Active RF and Microwave Devices

- **Diodes and Diode Circuits**
- **Bipolar Junction Transistors**
- **Field Effect Transistors**

Introduction

- Active devices include diodes, transistors, and electron tubes, which can be used for signal detection, mixing, amplification, frequency multiplication, and switching, and as sources of RF and microwave signals.
- it will be adequate to work with the terminal characteristics of diodes and transistors using equivalent circuits or scattering parameters.
- These results will be used to study some basic diode detector and control circuits; for the design of amplifier, mixer, and oscillator circuits using diodes and transistors.

self reading:

History (1)

- The first detector diode was probably the “cat-whisker” crystal detector used in early radio work of the nineteenth century.
- The advent of electron tubes used as detectors and amplifiers later eliminated this component in most radio systems, but crystal diodes were used by Southworth in his 1930s experiments with waveguides since tube detectors could not operate at such high frequencies.
- Frequency conversion and heterodyning were also first developed for radio applications in the 1920s.
- But it was not until the 1960s that the subject of microwave semiconductor devices saw significant progress.
- The invention of the gallium arsenide field effect transistor (FET) in the late 1960s was one of the most far-reaching developments in modern microwave engineering.
- RF and microwave transistors are critical components in wireless systems, finding application as amplifiers, oscillators, switches, phase shifters, mixers, and active filters.

History (2)

- Monolithic microwave integrated circuits (MMICs) combine transmission lines, active devices, and other components on a semiconductor substrate.
- The first single-function MMICs were developed in the late 1960s, but more sophisticated circuits and subsystems, such as multistage FET amplifiers, transmit/receive radar modules, front ends for wireless products, and many other circuits, are now being fabricated as MMICs.
- The trend is toward MMICs having higher performance, lower power requirements, greater complexity, and lower cost.

5

- Three types of semiconductors typically used in active circuits:

✓**Si**: has highest thermal conductivity ⇒ high power devices.

✓**GaAs**: has highest electron mobility ⇒ low loss devices.

✓**InP** (Indium phosphide) maximum drift velocity of e^{-1} ⇒ highest frequency devices.

Properties of semiconductor	Si	GaAs	InP
Electron mobility [m^2/V_s]	0.07	0.45	0.32
Highest electron velocity (saturation velocity) [m/s]	10^5	$1.7x10^5$	$2.5x10^5$
Saturation field [V/m]	2000	400	1200
Thermal conductance [Wm/°C]	$1.45x10^{-2}$	$4.4x10^{-3}$	$6.8x10^{-3}$

Diodes and Diode Circuits

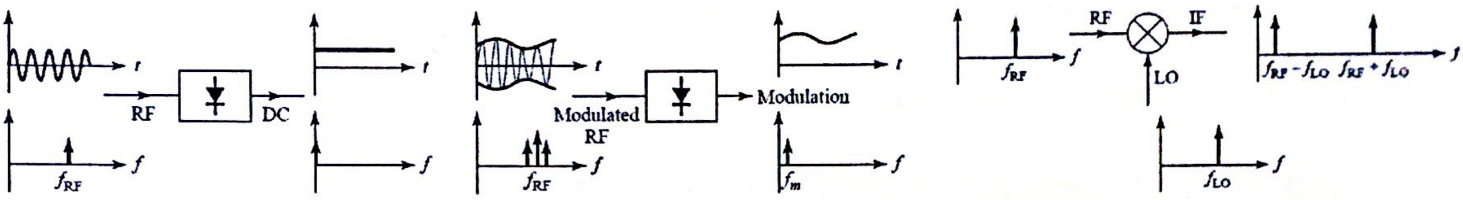
- A diode is a two-terminal semiconductor device having a nonlinear V–I relationship. This nonlinearity can be exploited for the useful functions of signal detection, demodulation, switching, frequency multiplication, and oscillation.
- RF and microwave diodes can be packaged as axial or beam lead components or as surface mountable chips, or be monolithically integrated with other components on a single semiconductor substrate.

Schottky Diodes and Detectors

- The classical pn junction diode commonly used at low frequencies has a relatively large junction capacitance that makes it unsuitable for high frequency application.
- The Schottky barrier diode, however, relies on a semiconductor–metal junction that results in a much lower junction capacitance, allowing operation at higher frequencies.
- Commercially available microwave Schottky diodes generally use **n-type gallium arsenide** (GaAs) material, while lower frequency versions may use **n-type silicon**.
- Schottky diodes are often biased with a small DC forward current, but can be used without bias.

Application of Schottky Diodes

- The primary application of Schottky diodes is in frequency conversion of an input signal.
- The three basic frequency conversion operations of rectification (conversion to DC), detection (demodulation of an amplitude-modulated signal), and mixing (frequency shifting).



Rectification   Detection   Mixing

V–I Characteristics of a Schottky Diode

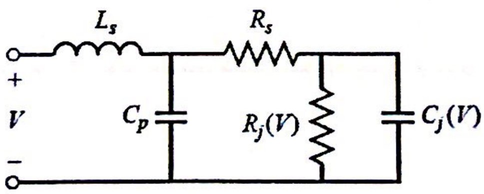
- A junction diode can be modeled as a nonlinear resistor, with a small-signal V–I relationship expressed as;

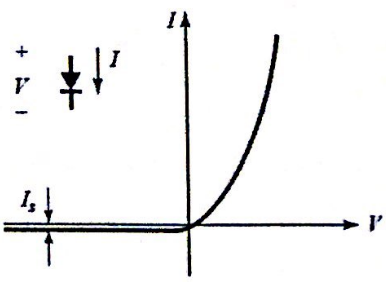
$$I(V) = I_s(e^{\alpha V} - 1) \qquad \alpha = q/nkT$$

q is the charge of an electron,
k is Boltzmann's constant,
T is temperature,
n is the ideality factor,
and I_s is the saturation current.

Typically, I_s is between 10^{-6} and 10^{-15} A, and $\alpha = q/nkT$ is approximately 1/(25 mV) for T = 290 K.
The ideality factor (n) depends on the structure of the diode, and can vary from about 1.05 for Schottky barrier diodes to about 2.0 for point-contact silicon diodes.

R_j is the junction resistance of the diode, $G_d = 1/R_j$ is called the dynamic conductance of the diode.





Small-Signal Approximation

- Small-signal approximation: Let the diode voltage be expressed as $V = V_0 + v$
- The small-signal approximation is based on the DC voltage–current relationship and shows that the equivalent circuit of a diode will involve a nonlinear resistance.
- In practice, however, the AC characteristics of a diode also involve reactive effects due to the structure and packaging of the diode.
- The leads or contacts of the diode package are modeled as a series inductance (L_s) and shunt capacitance, C_p.
- The series resistor (R_s) accounts for contact and current-spreading resistance.
- The junction capacitance (C_j) and the junction resistance (R_j) are **bias dependent**.

Parameters for Some Commercial Schottky Diodes

Schottky Diode	I_s (A)	R_s (Ω)	C_j (pF)	L_s (nH)	C_p (pF)
Skyworks SMS1546	3×10^{-7}	4	0.38	1.0	0.07
Skyworks SMS7630	5×10^{-6}	20	0.14	0.05	0.005
Avago HSMS2800	3×10^{-8}	30	1.6	—	—
Macom MA4E2054	3×10^{-8}	11	0.1	—	0.11

Diode Rectifiers

- In a rectifier application, a diode is used to convert a fraction of an RF input signal to DC power.
- Rectification is a very common function and is used for power monitors, automatic gain control circuits, and signal strength indicators.
- The total diode voltage consists of a DC bias voltage and a small-signal RF voltage ➔ $V = V_0 + v_0 \cos \omega_0 t$

- The output also contains AC signals of frequency ω_0, and $2\omega_0$ (as well as higher order harmonics), which are usually filtered out with a simple low-pass filter. A **current sensitivity**, β_i, can be defined as a measure of the change in the DC output current for a given RF input power.
- An **open-circuit voltage sensitivity**, β_v (typical range from 400 to 1500 mV/mW), can be defined in terms of the voltage drop across the junction resistance when the diode is open circuited.

$$\beta_i = \frac{\Delta I_{dc}}{P_{in}} = \frac{G'_d}{2G_d} \text{ A/W} \qquad \alpha G_d = G'_d \qquad \beta_v = \beta_i R_j$$

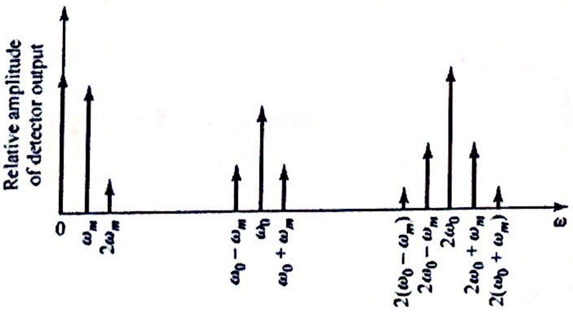
Diode Detectors

- In a detector application the nonlinearity of a diode is used to demodulate an amplitude modulated (AM) RF carrier.
- In this case, the diode voltage can be expressed as;

$$v(t) = v_0(1 + m \cos \omega_m t) \cos \omega_0 t$$

where ω_m is the modulation frequency, ω_0 is the RF carrier frequency ($\omega_0 >> \omega_m$), and m is defined as the modulation index ($0 \leq m \leq 1$)

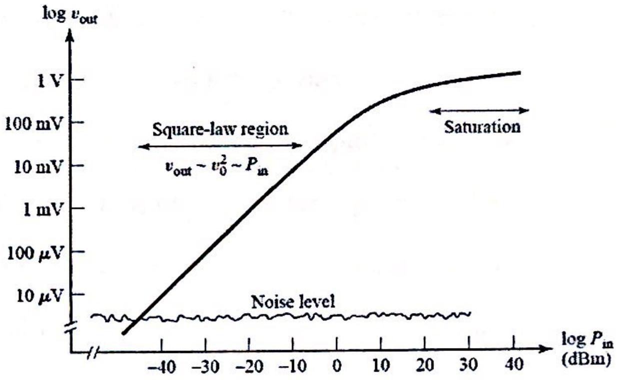
Output spectrum of a detected AM signal



Frequency	Relative Amplitude
0	$1 + m^2/2$
ω_m	$2m$
$2\omega_m$	$m^2/2$
$2\omega_0$	$1 + m^2/2$
$2\omega_0 \pm \omega_m$	m
$2(\omega_0 \pm \omega_m)$	$m^2/4$

Square-law Region for a Typical Diode Detector

- The desired demodulated output of frequency ω_m is easily separated from the undesired frequency components with a **low-pass filter**.
- Observe that the amplitude of this current is proportional to the square of the input signal voltage, and hence the input signal power.
- This square-law behavior is the usual operating condition for detector diodes, but it can be obtained only over a restricted range of input power.
- If the input power is too large, small-signal conditions will not apply, and the output will become saturated and approach a linear, and then a constant, i versus P characteristic.
- At very low signal levels the input signal will be lost in the noise floor of the device.

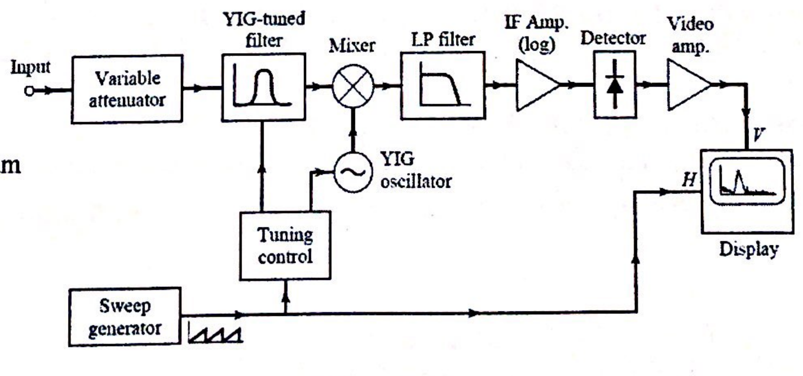


The output voltage can be considered as the voltage drop across a resistor in series with the diode

Spectrum Analyzer

- A spectrum analyzer gives a **frequency domain** representation of a signal, displaying the **average power density** versus frequency. Thus, its function is dual to that of an oscilloscope, which displays a time domain representation of a signal.
- A spectrum analyzer is basically a sensitive receiver that tunes over a specified frequency band and gives a video output that is proportional to the signal power in a narrow bandwidth.
- Spectrum analyzers are invaluable for measuring modulation products, harmonic and intermodulation distortion, noise, and interference effects.

Simplified Block Diagram of a Spectrum Analyzer



- A microwave spectrum analyzer can typically cover any frequency band in the range of several hundred megahertz to tens of gigahertz.
- The frequency resolution is set by the IF bandwidth, and is typically adjustable from about 100 Hz to 1 MHz.
- A sweep generator is used to repetitively scan the receiver over the desired frequency band by adjusting the local oscillator frequency, and to provide horizontal deflection of the display.
- An important part of a modern spectrum analyzer is the YIG-tuned bandpass filter at the input to the mixer.
- This filter is tuned along with the local oscillator, and acts as a pre-selector to reduce spurious intermodulation products.
- An IF amplifier with a logarithmic response is generally used to accommodate a wide dynamic range.
- Modern spectrum analyzers usually contain a computer to control the system and the measurement process.
- This improves performance and makes the analyzer more versatile, but can be a disadvantage in that the computer can sometimes remove the user from the physical reality of the measurement.

PIN Diodes and Control Circuits

- Switches are used extensively in microwave systems for directing signal or power flow between components.
- Switches can also be used to construct other types of control circuits, such as phase shifters and attenuators.
- Mechanical switches can be made in waveguide or coaxial form, and can handle high powers but are bulky and slow.
- PIN diodes, however, can be used to construct an electronic switching element easily integrated with planar circuitry and capable of high-speed operation.
- Switching speeds typically range from 1 to 10 μs, although speeds as fast as 20 ns are possible with careful design of the diode driving circuit.
- PIN diodes can also be used as power limiters, modulators, and variable attenuators.

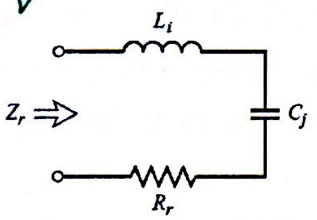
PIN Diode Characteristics

- A PIN diode contains an intrinsic (lightly doped) layer between the p and n semiconductor layers.
- When reverse biased, a small series junction capacitance leads to a relatively high diode impedance, while a forward bias current removes the junction capacitance and leaves the diode in a low-impedance state.
- These characteristics make the PIN diode a useful RF switching element.
- The parasitic inductance, Li , is typically less than 1 nH. The reverse resistance, Rr , is usually small relative to the series reactance due to the junction capacitance and is often ignored.
- The forward bias current is typically 10–30 mA, and the reverse bias voltage is typically 10–60 V.
- The bias voltages must be applied to the diode with RF chokes and DC blocks for isolation from the RF signal.

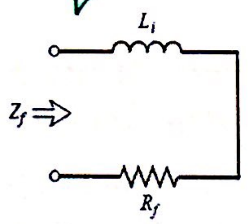
Parameters for Some Commercially Available PIN Diodes

memorize equ. cct in each case.

PIN Diode Equivalent circuits



Reverse Bias State

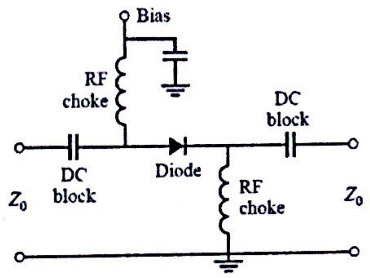


Forward Bias State

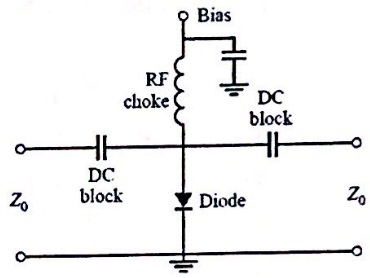
PIN Diode	R_f (Ω)	C_j (pF)
ASI 8001	3.0	0.03
Skyworks DSG9500	4.0	0.025
Infineon BA592	0.36	1.4
Microsemi UM9605	1.5	0.5

Single-pole PIN Diode Switches (1)

- A PIN diode can be used in either a series or a shunt configuration to form a single-pole, single-throw RF switch.
- These circuits are shown in the next figure along with the required bias networks.
- In the series configuration, the switch is ON when the diode is forward biased, while in the shunt configuration the switch is ON when the diode is reverse biased.
- In both cases, input power is reflected when the switch is in the OFF state.
- The DC blocking capacitors should have a relatively low impedance at the RF operating frequency, while the RF choke inductors should have a relatively high RF impedance.
- In some designs, high-impedance quarter-wavelength lines can be used in place of the chokes, to provide RF blocking.



Series Configuration



Shunt Configuration

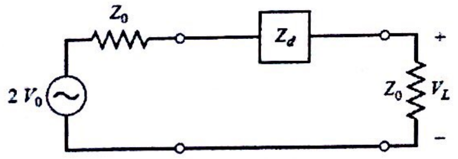
Single-pole PIN Diode Switches (2)

- An ideal switch would have zero insertion loss in the ON state, and infinite attenuation in the OFF state.
- Realistic switching elements, of course, result in some insertion loss for the ON state and finite attenuation for the OFF state.
- Knowing the diode parameters for the equivalent circuits allows the insertion loss for the ON and OFF states to be calculated for the series and shunt switches.

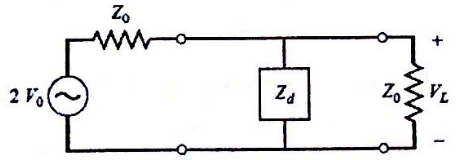
$$Z_d = \begin{cases} Z_r = R_r + j(\omega L_i - 1/\omega C_j) & \text{for reverse bias} \rightarrow \text{R, L \& C in series.} \\ Z_f = R_f + j\omega L_i & \text{for forward bias} \rightarrow \text{R \& L in series.} \end{cases}$$

$$\mathrm{IL} = -20 \log \left| \frac{V_L}{V_0} \right|$$

V_L the actual load voltage, and V_0, which is the load voltage that would appear if the switch Z_d (the diode impedance for either the reverse or forward bias state) were absent.



Series Switch



Shunt Switch

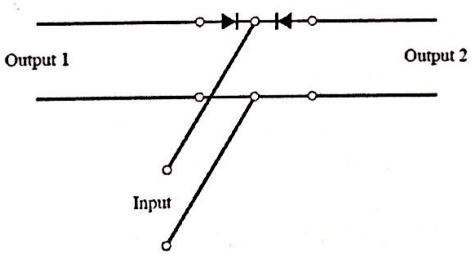
$$\mathrm{IL} = -20 \log \left| \frac{2Z_0}{2Z_0 + Z_d} \right| \quad \text{(series switch)}$$

$$\mathrm{IL} = -20 \log \left| \frac{2Z_d}{2Z_d + Z_0} \right| \quad \text{(shunt switch)}$$

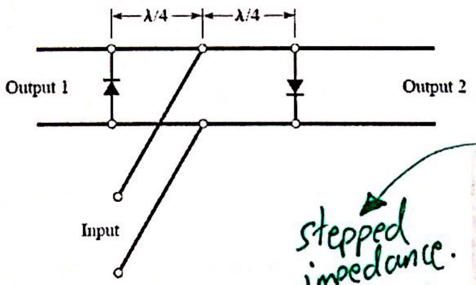
$Z_d \rightarrow$ diode $\rightarrow r$ (reverse).
$\rightarrow f$ (forward).

- The ON-state or OFF-state insertion loss of a switch can usually be improved by adding an external reactance in series or in parallel with the diode, to compensate for the diode reactance. This technique usually reduces the bandwidth, however.
- Several single-throw switches can be combined to form a variety of multiple-pole and/or multiple-throw configurations.
- In operation, one diode is forward biased in the low-impedance state, with the other diode reverse biased in the high-impedance state. The input signal is switched from one output to the other by reversing the diode bias states. The quarter-wave lines of the shunt circuit limit the bandwidth of this configuration.

Series SPDT PIN Switch



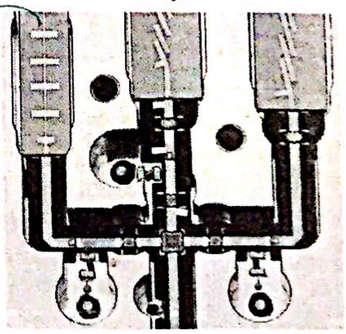
Shunt SPDT PIN Switch



- There are 3-Filters:

coupled resonators Filter

stepped impedance Filter.



Photograph of a SP3T GaAs PIN diode switch, operating from 6 to 27 GHz. The diode chips are 15 mils square.

Example: Single-Pole PIN Diode Switch

- A single-pole switch operating at 1.8 GHz is to be constructed using a Microsemi UM 9605 PIN diode with $C_j = 0.5$ pF and $R_f = 1.5\ \Omega$. What switch circuit (series or shunt) should be used to obtain the greatest ratio of off-to-on attenuation? Assume that $L_i = 0.5$ nH, $R_r = 2.0\ \Omega$, and $Z_0 = 50\ \Omega$

➔Solution:

$$Z_d = \begin{cases} Z_r = R_r + j(\omega L_i - 1/\omega C_j) & = 2.0 - j171.2\ \Omega \\ Z_f = R_f + j\omega L_i & = 1.5 + j5.6\ \Omega. \end{cases}$$

For the series circuit:

$$\text{IL}_{on} = -20\log\left|\frac{2Z_0}{2Z_0 + Z_f}\right| = 0.14 \text{ dB}$$

$$\text{IL}_{off} = -20\log\left|\frac{2Z_0}{2Z_0 + Z_r}\right| = 6.0 \text{ dB}$$

For the shunt circuit:

$$\text{IL}_{on} = -20\log\left|\frac{2Z_r}{2Z_r + Z_0}\right| = 0.11 \text{ dB}$$

$$\text{IL}_{off} = -20\log\left|\frac{2Z_f}{2Z_f + Z_0}\right| = 13.3 \text{ dB}$$

The shunt configuration has the greatest difference in attenuation between the ON and OFF states and has the lowest ON insertion loss

$$\frac{13.3}{0.11} > \frac{6.0}{0.14}$$

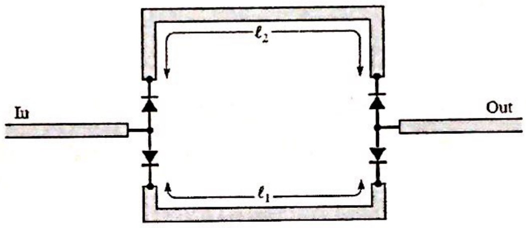
PIN Diode Phase Shifters

- Several types of microwave phase shifters can be constructed with PIN diode switching elements.
- Compared with ferrite phase shifters, diode phase shifters have the advantages of small size, integrability with planar circuitry, and high speed.
- The power requirements for diode phase shifters, however, are generally greater than those for a latching ferrite phase shifter because diodes require continuous bias current, while a latching ferrite device requires only a pulsed current to change its magnetic state.
- There are basically three types of PIN diode phase shifters: switched line, loaded line, and reflection.
- The switched-line phase shifter, is the most straightforward type, using two single-pole, double-throw switches to route the signal flow between one of two transmission lines of different length.
- The differential phase shift between the two paths is given by

$$\Delta\phi = \beta(\ell_2 - \ell_1)$$ β is the propagation constant of the line

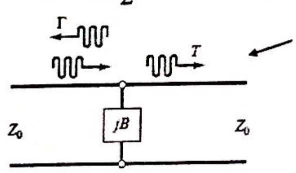
A Switched-line Phase Shifter

- If the transmission lines are TEM (or quasi-TEM, like microstrip), this phase shift is a linear function of frequency, which implies a true time delay between the input and output ports. This is a useful feature in wideband systems. This type of phase shifter is also inherently reciprocal, and so it can be used for both receive and transmit functions. The insertion loss of the switched line phase shifter is equal to the loss of the SPDT switches plus line losses.



Switched-line Phase Shifter

- Like many other types of phase shifters, the switched-line phase shifter is usually designed for discrete binary phase shifts of $\varphi = 180°$, $90°$, $45°$, etc.
- One potential problem with this type of phase shifter is that resonances can occur in the OFF line if its length is near a multiple of $\lambda/2$.
- The resonant frequency will be slightly shifted due to the series junction capacitances of the reversed biased diodes, so the lengths ℓ_1 and ℓ_2 should be determined with this effect taken into account.



A design that is useful for small amounts of phase shift (generally 45°, or less) is the loaded-line phase shifter

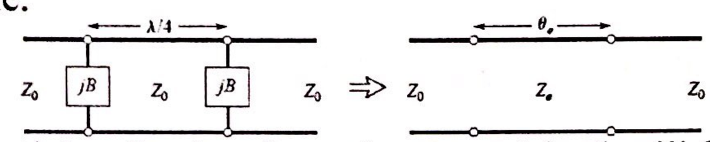
$$\Gamma = \frac{1-(1+jb)}{1+(1+jb)} = \frac{-jb}{2+jb}$$

$$T = 1 + \Gamma = \frac{2}{2+jb}$$

$$\Delta\phi = \tan^{-1}\frac{b}{2}$$

where $b = BZ_0$ is the normalized susceptance.
which can be made positive or negative, depending on the sign of b.
A disadvantage is the insertion loss that is inherently present due to the reflection from the shunt load.
In addition, **increasing** b to obtain a larger φ entails a **greater insertion loss**.

- The reflections from the shunt susceptance can be reduced by using the following circuit, where two shunt loads are separated by a $\lambda/4$ length of line.



$$\theta_e \simeq \frac{\pi}{2} + b$$

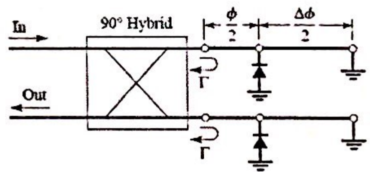
$$Z_e \simeq Z_0\left(1 + \frac{b}{2}\right)$$

- Then the partial reflection from the second load will be 180° out of phase with the partial reflection from the first load, leading to cancellation.
- The susceptance (B) can be implemented with a lumped inductor or capacitor, or with a stub, and switched between two states with an SPST diode switch.

Not included.

A Reflection Phase Shifter

A reflection-type phase shifter using a quadrature hybrid:
Operation Principle:
An input signal divides equally between the two right-hand ports of the hybrid.
The diodes are both biased in the same state (forward or reverse biased), so the waves reflected from the two terminations will add in phase at the indicated output port.
Turning the diodes on or off changes the total path length for both reflected waves by φ, producing a phase shift of φ at the output.
Ideally, the diodes would look like short circuits in their ON state, and open circuits in their OFF state, so that the reflection coefficients at the right side of the hybrid can be written as $\Gamma = e^{-j(\phi + \pi)}$ for the diodes in their ON state, and $\Gamma = e^{-j(\phi + \Delta\phi)}$ for the diodes in their OFF state.
There is an infinite number of choices of line lengths that give the desired φ (i.e., the value of $\varphi/2$ is a degree of freedom).



The bandwidth is optimized if the reflection coefficients for the two states are phase conjugates. Thus, if φ = 90°, the best bandwidth will be obtained for φ = 45°

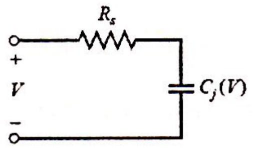
The insertion loss is limited by the loss of the hybrid, as well as by the forward and reverse resistances of the diodes.
Impedance transformation sections can be used to improve performance in this regard.

Varactor Diodes

- The PIN diode has a junction capacitance that can be switched on or off with bias voltage.
- This effect can be enhanced by tailoring the size and doping profile of the intrinsic layer of the diode to provide a desired junction capacitance versus junction voltage (C vs. V) behavior when reverse biased.
- Such a device is called a varactor diode, and it produces a junction capacitance that varies smoothly with bias voltage, thus providing an electrically adjustable reactive circuit element.
- One of the most common applications of varactor diodes is to provide electronic frequency tuning of the local oscillator in a multichannel receiver, such as those used in cellular telephones, wireless local area network radios, and television receivers.
- This is accomplished by using a varactor diode in the resonant circuit of a transistor oscillator, and controlling the DC reverse bias voltage applied to the diode.
- The nonlinearity of varactor diodes also makes them useful for frequency multipliers.
- Varactor diodes are generally made from silicon for RF applications, and gallium arsenide for microwave applications.

Varactor Diode Equivalent Circuit

- A simplified equivalent circuit for a **reverse-biased** varactor diode is:



- The junction capacitance is dependent on the (negative) junction bias voltage (V) according to

$$C_j(V) = \frac{C_0}{(1 - V/V_0)^\gamma}$$

where C_0 is the junction capacitance with no bias; V_0= 0.5 V for silicon diodes, and V_0= 1.3 V for GaAs diodes. The exponent γ depends on the doping profile of the intrinsic layer of the diode. An ideal hyper-abrupt varactor diode has $\gamma = 0.5$; many practical diodes have an exponent of about $\gamma = 0.47$, although the value can be as high as 1.5 or 2.0 for some diodes.

In the equivalent circuit, R_s is the series junction and contact resistance, typically on the order of a few ohms.

A typical GaAs varactor diode may have C_0= 0.5 – 2.0 pF, and a junction capacitance that varies from about 0.1 to 2.0 pF as the bias voltage ranges from −20 to 0 V.

Parasitic reactances due to the diode package should be included in a realistic design.

Self Reading:

Other Diodes (1)

Gunn diodes: The operation of a Gunn diode is based on the transferred electron effect (also known as the Gunn effect), which was discovered by J. B. Gunn in 1963.

Practical Gunn diodes typically use GaAs or InP materials in a specially doped bulk form, as opposed to a traditional pn junction.

The Gunn diode has an I –V characteristic that exhibits a negative differential resistance (negative slope) that can be used to generate RF power directly from a DC source when properly biased.

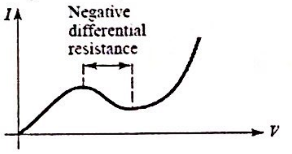
The region of negative differential resistance (negative slope) corresponds to the operating point of the device.

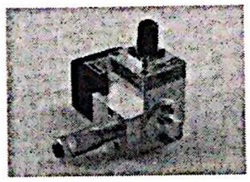
Gunn diodes can produce continuous power of up to several hundred milliwatts, at frequencies from 1 to 200 GHz, with efficiencies ranging from 5% to 15%.

Oscillator circuits using Gunn diodes require a high-Q resonant circuit or cavity, which is often tuned mechanically.

Electronic tuning by bias adjustment is limited to 1% or less, but varactor diodes are sometimes included in the resonant circuit to provide a greater range of electronic tuning.

Gunn diode sources are used extensively in low-cost applications such as traffic radars, motion detectors for door openers and security alarms, and test and measurement systems.





Other Diodes (2)

IMPATT diodes: An **imp**act **a**valanche and **t**ransit **t**ime (IMPATT) diode has a physical structure similar to a PIN diode, but is operated with a relatively high voltage (70–100 V) to produce a reverse-biased avalanche breakdown current.

It exhibits a negative resistance over a broad frequency band that can extend into the submillimeter range, and it can be used to directly convert DC to RF power.

IMPATT sources are generally noisier than Gunn diodes but are capable of higher powers and higher DC-to-RF conversion efficiencies.

IMPATTs also have better temperature stability than Gunn diodes. Typical IMPATTs operate at frequencies from 10 to 300 GHz, with efficiencies ranging up to 15%.

IMPATT diodes are among the few practical solid-state devices that can provide fundamental frequency power above 100 GHz.

IMPATT devices can also be used for frequency multiplication and amplification.

Silicon IMPATT diodes can provide CW power ranging from 10 W at 10 GHz to 1 W at 94 GHz, with efficiencies typically below 10%.

GaAs IMPATTs can provide CW power ranging from 20 W at 10 GHz to 5 mW at 130 GHz.

Pulsed operation generally results in higher powers and higher efficiencies. Because of the low efficiency of these devices, thermal considerations are a limiting factor for both CW and pulsed operation.

IMPATT oscillators can be mechanically or electrically tuned.

A disadvantage of IMPATT oscillators is that their AM noise level is generally higher than that of other sources.

Other Diodes (3)

- **Tunnel diodes:**
- The tunnel diode, invented by L. Esaki in 1957, is a pn junction diode with a doping profile that allows electron tunneling through a narrow energy band gap, leading to negative resistance at high frequencies.
- Tunnel diodes can be used for oscillators as well as amplifiers.
- Before high-frequency transistors were available, tunnel diodes provided the only means of high-frequency amplification with a solid-state device.
- Such an amplifier employs the diode in a one-port reflection circuit, where the negative RF resistance of the device produces a reflection coefficient with a magnitude greater than unity, and therefore amplification of an incident signal.
- Such amplifiers have been made obsolete by modern RF and microwave transistors, but tunnel diodes are still used in some applications today.

Other Diodes (4)

- **BARITT diodes:**
- A barrier injection transit time (BARITT) diode has a structure similar to a junction transistor without a base contact.
- Like the IMPATT diode, it is a transit time device.
- It generally has a lower power capability than the IMPATT diode, but the advantage of lower AM noise.
- This makes it useful for local oscillator applications at frequencies up to 94 GHz.
- BARITT diodes are also useful for detector and mixer applications.

Transistors

- Transistors are three-terminal semiconductor devices, and can be categorized as either:
- *junction transistors* or
- *field effect transistors*.
- Junction transistors include

✓ *bipolar junction transistors* (BJTs) that use a single semiconductor material (usually silicon).

✓ *heterojunction bipolar transistors* (HBTs) that use compound semiconductors such as GaAs, indium phosphide (InP), or silicon germanium (SiGe), often in conjunction with thin layers of other materials (e.g., aluminum).

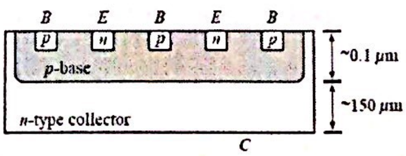
- Both *npn* and *pnp* configurations are possible, but most RF junction transistors are usually of the *npn* type due to higher electron mobility at higher frequencies.

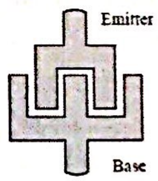
- Bipolar just use one material (Silicon or Germanium).
- HBT use compound (GaAs or SiGe).

Bipolar Junction Transistors (Not important).

- is one of the oldest and most popular active RF devices in use today because of its low cost and good operating performance in terms of frequency range, power capacity, and noise characteristics.
- Silicon junction transistors are useful for amplifiers up to the range of 2–10 GHz, and in oscillators up to about 20 GHz.
- Bipolar transistors typically have very low (1/ f) noise characteristics, making them well suited for oscillators with low-phase noise.
- Bipolar junctiontransistors are sometimes preferred over FETs at frequencies below about 2–4 GHz because of higher gain and lower cost, and the possibility of biasing with a single power supply.
- Bipolar transistors are subject to shot noise as well as thermal noise effects, so their noise figure is not as good as that of FETs.

Construction of a typical silicon bipolar transistor having multiple fingers for the base and emitter electrodes.

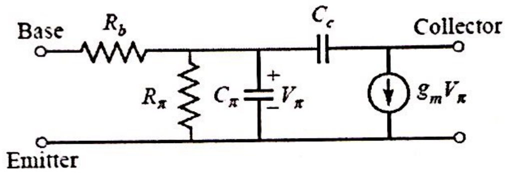




Small-signal equivalent circuit model for an RF BJT

- The BJT is current driven, with the base current modulating the collector current.
- The upper frequency limit of the bipolar transistor is controlled primarily by the base length, which is typically on the order of 0.1 μm.
- Hybrid-π model, is popular because of its similarity to the equivalent circuit of a FET, and because of its utility in circuit analysis.

This model does not include parasitic resistances and inductances due to the base and emitter leads



Simplified hybrid-π equivalent circuit for a microwave bipolar junction transistor in the common emitter configuration

- In many cases the capacitor (C_c) between the base and collector in the hybrid-π model, has a relatively small value and may be ignored.
- This has the effect of making $S_{12} = 0$, implying that power only flows in one direction through the device (from port 1 to port 2); such a device is called **unilateral**. This approximation is often used to simplify analysis.
- The hybrid-π model is roughly based on the physics of the junction transistor, and can be **useful** under circumstances where the element values of the model are fairly **constant** over a range of operating **bias conditions**, **load conditions**, and **frequency**.
- Otherwise, the element values become frequency, bias, or load **dependent**, in which case the hybrid-π model (or any other equivalent circuit model) becomes much less useful.

- It is simpler to treat the transistor as a two-port network, characterized by two-port parameters. In practice, scattering parameters, measured under typical operating conditions, are usually used for this purpose and are supplied by the device manufacturer

- Note that there are relatively large mismatches at the base (port 1) and the collector (port 2), and that the gain (given roughly by $|S_{21}|$) drops quickly with an increase in frequency.
- Also note that $|S_{12}|$ is relatively small (particularly at low frequencies), making the device approximately unilateral.

$S_{21} > S_{12}$ → since the source @ port 1 most of the time.

Frequency (GHz)	S_{11}	S_{12}	S_{21}	S_{22}
0.1	$0.78\angle -33°$	$0.03\angle 71°$	$12.7\angle 155°$	$0.93\angle -17°$
0.5	$0.46\angle -113°$	$0.08\angle 52°$	$6.3\angle 104°$	$0.53\angle -38°$
1.0	$0.38\angle -158°$	$0.11\angle 54°$	$3.5\angle 80°$	$0.40\angle -43°$
2.0	$0.40\angle 157°$	$0.19\angle 56°$	$1.9\angle 52°$	$0.33\angle -63°$
4.0	$0.52\angle 117°$	$0.38\angle 45°$	$1.1\angle 14°$	$0.33\angle -127°$

Scattering Parameters for an NPN Silicon BJT (NEC NE 58219, $V_{ce} = 5.0$ V, $I_c = 5.0$ mA, common emitter)

*changing in S due to change in operating frequancy.

41

- The upper frequency limit (f_T) defined as the **threshold frequency** can be estimated when the short-circuit current gain (G_I^{SC}) of the transistor is unity.
- If we assume an input current I_{in} at the base, and ignore the series base resistance R_b (typically small), and the shunt resistance R_π (typically large), then the voltage across the capacitor C_π is $V_\pi = I_{in}/j\omega C_\pi$
- The output short-circuit current at the collector is $I_{out} = g_m V_\pi$, so the short-circuit current gain is

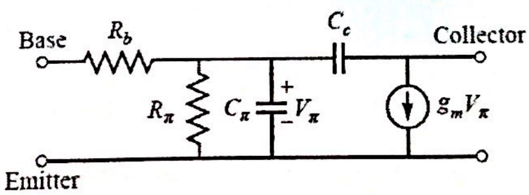
$$G_I^{sc} = \left|\frac{I_{out}}{I_{in}}\right| = \frac{g_m}{\omega C_\pi}$$

→ write it on formula sheet.

- The current gain is seen to decrease with frequency, and is unity at the **threshold frequency** (f_T)

$$f_T = \frac{g_m}{2\pi C_\pi}$$

⇒ we have it when the gain is equal to unity.

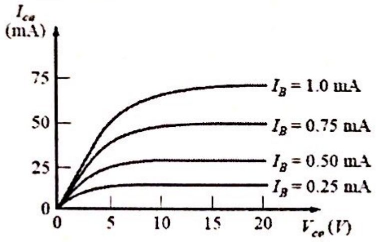


$$G = 1 = \frac{g_m}{2\pi f_T C_\pi}$$

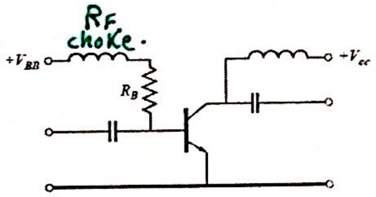
$$\Rightarrow f_T = \frac{g_m}{2\pi C_\pi} \#$$

42

- The biasing point for the transistor depends on the application and type of device, with **low collector currents** generally giving the **best noise figure**, and **higher collector currents** giving the **best power gain**.

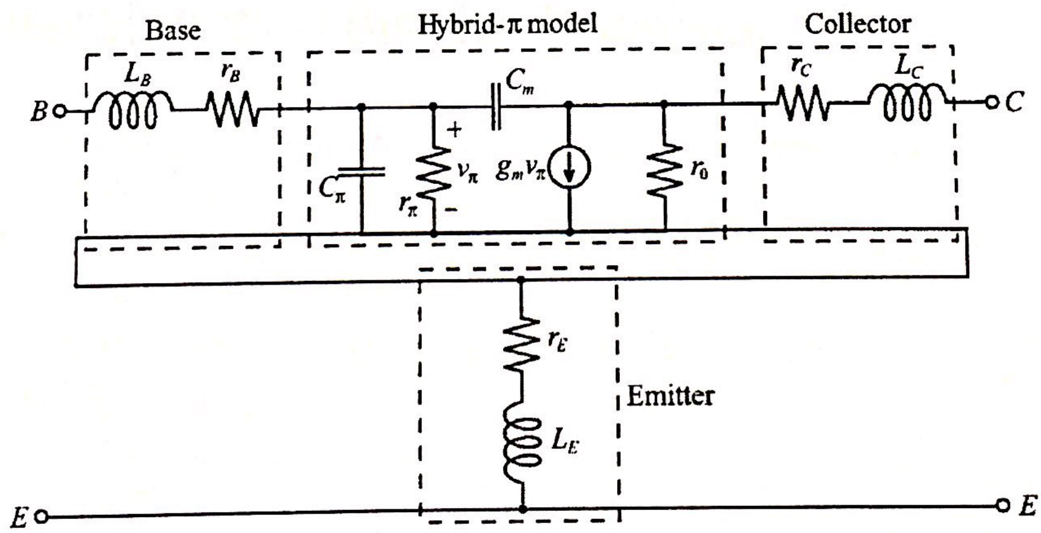


A typical DC operating characteristics for a BJT



A typical bias and decoupling circuit for a bipolar transistor in a common emitter configuration

AC model (hybrid) of a BJT used in Microwave amplifier:



Typical Values: C_m=16 pF, C_π=37 pF, L_E=0.5 nH, L_B=1.1 nH, L_C=1.1 nH, r_E=1.5 Ω, r_C=1.5 Ω, r_B=125 Ω,

Heterojunction Bipolar Transistor (1)

- This structure offers much improved performance at high frequencies.
- Some HBTs can operate at frequencies exceeding 100 GHz, and recent developments with HBTs using SiGe have demonstrated that these devices are useful in low-cost circuits operating at frequencies of 60 GHz or higher.
- Since the HBT is similar in structure and operation to the BJT, the equivalent circuit model of BJT can be used for both transistor types.
- As with BJTs, equivalent circuit models may have limited applicability when attempting to model HBTs over a range of operating conditions, so scattering parameter data, measured for a particular bias point, may be more useful.
- High levels of monolithic integration are easy and inexpensive with SiGe HBTs, so this technology is proving to be very useful for low-cost millimeter wave circuits for both defense and commercial applications

Heterojunction Bipolar Transistor (2)

- The scattering parameters at several frequencies for a popular microwave HBT.
- Observe that $|S_{21}|$ decreases much less rapidly with frequency when compared with the BJT.
- The device also is seen to be approximately unilateral, as $|S_{12}|$ is relatively small.

Frequency (GHz)	S_{11}	S_{12}	S_{21}	S_{22}
1.0	$0.91\angle{-44°}$	$0.06\angle{68°}$	$3.92\angle{149°}$	$0.93\angle{-17°}$
2.0	$0.75\angle{-86°}$	$0.10\angle{46°}$	$3.39\angle{120°}$	$0.79\angle{-31°}$
4.0	$0.59\angle{-144°}$	$0.11\angle{29°}$	$2.18\angle{82°}$	$0.64\angle{-43°}$
6.0	$0.54\angle{176°}$	$0.11\angle{34°}$	$1.64\angle{57°}$	$0.58\angle{-53°}$

Note that reduction in S_{21} for Hetrojunction is Less than reduction in S_{21} for NPN silicon BJT.
since HBT is made from more than one material.

Field Effect Transistors

- In contrast to BJTs, field effect transistors (FETs) are monopolar, with only one carrier type (holes or electrons) providing current flow through the device: n-channel FETs employ electrons, while p-channel devices use holes.
- In addition, while a BJT is a current controlled device, an FET is a voltage-controlled device, having a source-to-drain characteristic that is similar to that of a voltage-dependent variable resistor.
- Field effect transistors can take many forms, including the MESFET (metal semiconductor FET), the MOSFET (metal oxide semiconductor FET), the HEMT (high electron mobility transistor), and the PHEMT (pseudomorphic HEMT).

History of FET → self reading.

- The first junction FETs were developed in the 1950s, while the HEMT was proposed in the early 1980s.
- GaAs MESFETs are among the most commonly used transistors for microwave and millimeter wave applications, being usable at frequencies up to 60 GHz or more.
- Even higher operating frequencies can be obtained with GaAs HEMTs.
- GaAs MESFETs and HEMTs are especially useful for low-noise amplifiers since these transistors have lower noise figures than any other active devices.
- Recently developed gallium nitride (GaN) HEMTs are very useful for high power RF and microwave amplifiers.
- CMOS FETs are increasingly being used for RF integrated circuits, offering high levels of integration at low cost and low power requirements, for commercial wireless applications.

Performance Characteristics of Microwave Transistors

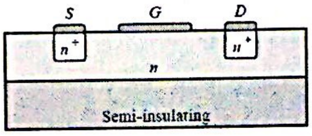
Device Semiconductor	BJT Si	HBT SiGe	CMOS Si	MESFET GaAs	HEMT GaAs	HEMT GaN
Frequency range (GHz)	10	30	20	60	100	10
Typical gain (dB)	10–15	10–15	10–20	5–20	10–20	10–15
Noise figure (dB)	2.0	0.6	1.0	1.0	0.5	1.6
(frequency, GHz)	(2)	(8)	(4)	(10)	(12)	(6)
Power capacity	High	Medium	Low	Medium	Medium	High
Cost	Low	Medium	Low	Medium	High	Medium
Single-polarity supply	Yes	Yes	Yes	No	No	No

we will focus on these.

Metal Semiconductor Field Effect Transistor

- One of the most important developments in microwave technology has been the GaAs metal semiconductor field effect transistor (MESFET), as this device permitted the first practical solid-state implementation of amplifiers, oscillators, and mixers at microwave frequencies, leading to key applications in radar, GPS, remote sensing, and wireless communications.
- GaAs MESFETs can be used at frequencies well into the millimeter wave range, with high gain and low noise figure, often making them the device of choice for hybrid and monolithic integrated circuits at frequencies above 10 GHz.

- The gate junction is formed as a Schottky barrier.
- The desirable gain and noise features of this transistor are a result of the higher electron mobility of GaAs compared to silicon, and the absence of shot noise.
- The device is biased with a drain-to-source voltage (V_{ds}) and a gate-to-source voltage (V_{gs}).
- In operation, electrons are drawn from the source to the drain by the positive V_{ds} supply voltage. An applied signal voltage on the gate then modulates these majority electron carriers, producing voltage amplification.
- The maximum frequency of operation is limited by the gate length; present FETs have gate lengths on the order of 0.2–0.6 μm, with corresponding upper frequency limits of 100 to 50 GHz.

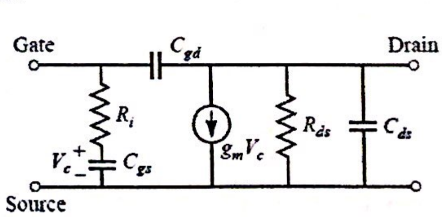


n-channel GaAs MESFET

Small-signal Equivalent Circuit for a Microwave MESFET for a Common-Source Configuration

- This model does not include package parasitics, which typically introduce small series resistances and inductances at the three terminals due to ohmic contacts and bonding leads.
- The dependent current generator $g_m V_c$ depends on the voltage across the gate-to-source capacitor (C_{gs}) leading to a value of $|S_{21}| > 1$ under normal operating conditions (where port 1 is at the gate, and port 2 is at the drain).
- The reverse signal path, given by S_{12}, is due solely to the capacitance C_{gd}

R_i (series gate resistance)	$= 7\ \Omega$
R_{ds} (drain-to-source resistance)	$= 400\ \Omega$
C_{gs} (gate-to-source capacitance)	$= 0.3$ pF
C_{ds} (drain-to-source capacitance)	$= 0.12$ pF
C_{gd} (gate-to-drain capacitance)	$= 0.01$ pF
g_m (transconductance)	$= 40$ mS

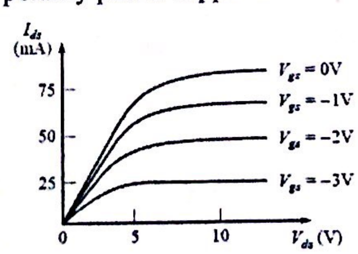


- To determine the upper frequency of operation for a MESFET.
- For a FET, the short-circuit current gain, G_I^{sc} is defined as the ratio of drain current to gate current when the output is short circuited.
- For the unilateral case, where $C_{gd} = 0$, the short circuit current gain is

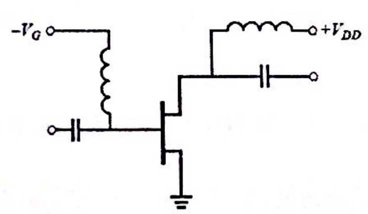
$$G_I^{sc} = \left|\frac{I_d}{I_g}\right| = \left|\frac{g_m V_c}{I_g}\right| = \frac{g_m}{\omega C_{gs}}$$

$$\boxed{f_T = \frac{g_m}{2\pi C_{gs}}}$$ a result that is equivalent to a bipolar junction transistor

- For proper operation, the transistor must be biased at an appropriate operating point.
- This depends on the application (low noise, high gain, high power), the class of the amplifier (class A, class AB, class B), and the transistor.
- For low-noise design, the drain current is generally chosen to be about 15% of I_{dss} (the saturated drain-to-source current).
- High-power circuits generally use higher values of drain current.
- DC bias voltage must be applied to both the gate and drain, without disturbing the RF signal paths.
- This can be done with biasing and decoupling circuitry for a dual-polarity supply.
- The RF chokes provide a very low DC resistance for biasing, and a very high impedance at RF frequencies to isolate the signal from the bias supply.
- Similarly, the input and output decoupling capacitors block DC from the input and output lines while allowing passage of RF signals.
- More sophisticated bias circuits can provide compensation for temperature and device variations, and may work with single-polarity power supplies.



DC characteristics of an n-channel GaAs MESFET

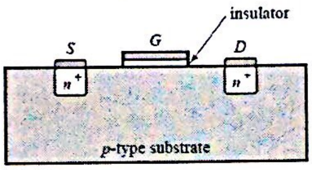


biasing and decoupling circuitry

self reading

Metal Oxide Semiconductor Field Effect Transistor (MOSFET)

- The silicon metal oxide semiconductor field effect transistor (MOSFET) is the most common type of FET, being used extensively in analog and digital integrated circuits.
- It consists of a lightly doped p substrate, and differs from a MESFET by having a thin insulating layer (SiO2) between the gate contact and the channel region.
- Because the gate is insulated, it does not conduct DC bias current.
- MOSFETs can be used at frequencies into the UHF range, and can provide powers of several hundred watts when devices are packaged in parallel.
- Laterally diffused MOSFETs (LDMOS) have direct grounding of the source, and can operate at low microwave frequencies with high powers.
- These devices are commonly used for high-power transmitters for cellular base stations at 900 and 1900 MHz.



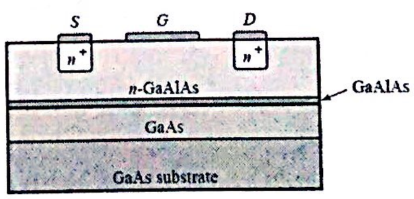
Cross section of an n-channel MOSFET

- High-density integrated circuits typically use complementary MOS (CMOS), where both n-channel and p-channel devices are used.
- This technology is very mature, and has the advantages of low power requirements and low unit cost.
- Most RF and microwave MOSFETs use n-channel silicon devices, although GaN devices are possible.
- The small-signal equivalent circuit for a MOSFET is the same as that of the MESFET.
- Scattering parameters are available for most nMOS devices intended for high-frequency applications.

High Electron Mobility Transistor (HEMT)

- The high electron mobility transistor (HEMT) is a heterojunction FET, meaning that it does not use a single semiconductor material, but instead is constructed with several layers of compound semiconductor materials.
- These may include transitions between gallium aluminum arsenide (GaAlAs), GaAs, gallium indium arsenide (GaInAs), and similar
- compounds.
- These structures result in high carrier mobility—about twice that found in a standard MESFET.
- GaAs HEMTs can operate at frequencies above 100 GHz.
- The cross section of a HEMT device consists of semi-insulating GaAs substrate, followed by an undoped GaAs layer, and then a very thin undoped GaAlAs layer. This is topped with an n-doped GaAlAs layer.
- To reduce thermal and mechanical stress the layers usually have matched crystal lattices.
- Several variations on this device are possible, including the use of different compound semiconductors, and the pseudomorphic HEMT, which uses a lattice mismatch between the layers.
- The relatively complicated structure of the HEMT requires sophisticated fabrication techniques, leading to a relatively high cost.
- The HEMT is also referred to in the literature as a MODFET (modulation-doped FET), a TEGFET (two-dimensional electron gas FET), and an SDFET (selectively doped FET).

- A relatively new type of HEMT uses GaN and aluminum gallium nitride (AlGaN) on a silicon or SiC substrate.
- GaN HEMTs operate with drain voltages in the range of 20–40 V, and can deliver powers up to 100 W at frequencies in the low microwave range, making these devices popular for high-power transmitters.
- The equivalent circuit model of Figure MESFET can also be used for HEMTs, and the DC bias characteristics of a HEMT are similar to those of the MESFET.



Cross section of an n-channel HEMT

Scattering Parameters for a GaN HEMT (Cree CGH21120, V_{DD}= 328 V, I_D= 500 mA, common source)

- The scattering parameters for a medium power GaN HEMT

Frequency (GHz)	S_{11}	S_{12}	S_{21}	S_{22}
0.5	$0.96\angle 180°$	$0.007\angle -16°$	$3.67\angle 68°$	$0.72\angle -174°$
1.0	$0.95\angle 172°$	$0.008\angle -35°$	$2.03\angle 44°$	$0.78\angle -172°$
2.0	$0.78\angle 153°$	$0.014\angle -83°$	$2.09\angle -17°$	$0.91\angle -174°$
4.0	$0.88\angle -51°$	$0.008\angle 79°$	$0.84\angle 88°$	$0.88\angle 171°$

Topics in Communications.
"Microwaves"

Spring 2017/2018

Dr. Yanal Al-Faouri

By: Mohammad Abu Hashya.

* * *

Microwave Engineering

Chapter 12

Dr. Yanal Faouri

Email: y.faouri@ju.edu.jo

Microwave Amplifier Design

- Two-Port Power Gains
- Stability
- Single-Stage Transistor Amplifier Design
- Low Noise Amplifier Design
- Power Amplifiers

Introduction

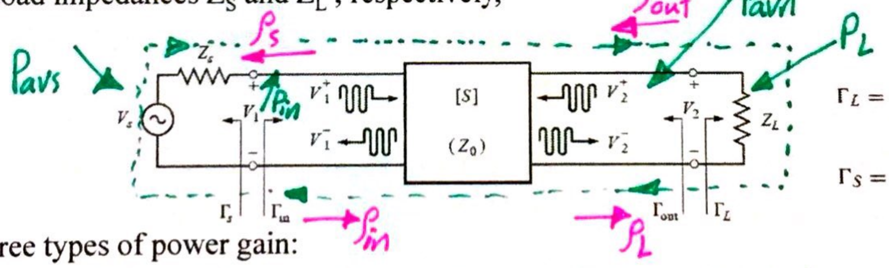
- Signal amplification is one of the most basic circuit functions in modern RF and microwave systems.
- Early microwave amplifiers relied on tubes, such as klystrons and traveling-wave tubes, or solid-state reflection amplifiers based on the negative resistance characteristics of tunnel or varactor diodes.
- However, due to the dramatic improvements and innovations in solid-state technology that have occurred since the 1970s, most RF and microwave amplifiers today use transistor devices such as Si BJTs, GaAs or SiGe HBTs, Si MOSFETs, GaAs MESFETs, or GaAs or GaN HEMTs.
- Microwave transistor amplifiers are rugged, low-cost, and reliable and can be easily integrated in both hybrid and monolithic integrated circuitry.
- Transistor amplifiers can be used at frequencies in excess of 100 GHz in a wide range of applications requiring small size, low noise figure, broad bandwidth, and medium to high power capacity.

for $\rho_s = 0$ (choose $Z_S = Z_0$)

Two-Port Power Gains

$[S]$ related to Z_0

- Consider an arbitrary two-port network, characterized by its scattering matrix [S], connected to source and load impedances Z_S and Z_L, respectively,



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

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

- There are three types of power gain:
- **Power gain** = $G = P_L/P_{in}$ is the ratio of power dissipated in the load Z_L to the power delivered to the input of the two-port network. This gain is independent of Z_S, although the characteristics of some active devices may be dependent on Z_S.
- **Available power gain** = $G_A = P_{avn}/P_{avs}$ is the ratio of the power available from the two-port network to the power available from the source. This assumes conjugate matching of both the source and the load, and depends on Z_S, but not Z_L.
- **Transducer power gain** = $G_T = P_L/P_{avs}$ is the ratio of the power delivered to the load to the power available from the source. This depends on both Z_S and Z_L.
- The gain is maximized ($G = G_A = G_T$) when the input and output are both conjugately matched to the two-port device.

we want to have two matching networks as will be shown in the next slides.

- In general, the input impedance of the terminated two-port network will be mismatched with a reflection coefficient given by Γ_{in}, which can be determined using a signal flow graph or by the scattering parameters by using $V_2^+ = \Gamma V_2^-$, we have:

$$V_1^- = S_{11}V_1^+ + S_{12}V_2^+ = S_{11}V_1^+ + S_{12}\Gamma_L V_2^-,$$

$$V_2^- = S_{21}V_1^+ + S_{22}V_2^+ = S_{21}V_1^+ + S_{22}\Gamma_L V_2^-.$$

$$\Gamma_{in} = \frac{V_1^-}{V_1^+} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0}$$

if $\rho_{in} < 1$
$\Rightarrow \rho_L < 1$

$$\Gamma_{out} = \frac{V_2^-}{V_2^+} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S}$$

$$V_1 = V_S\frac{Z_{in}}{Z_S + Z_{in}} = V_1^+ + V_1^- = V_1^+(1 + \Gamma_{in}).$$ VDR

$$Z_{in} = Z_0\frac{1 + \Gamma_{in}}{1 - \Gamma_{in}}$$

$$V_1^+ = \frac{V_S}{2}\frac{(1 - \Gamma_S)}{(1 - \Gamma_S\Gamma_{in})}$$

$$P_{in} = \frac{1}{2Z_0}|V_1^+|^2(1 - |\Gamma_{in}|^2) = \frac{|V_S|^2}{8Z_0}\frac{|1 - \Gamma_S|^2}{|1 - \Gamma_S\Gamma_{in}|^2}(1 - |\Gamma_{in}|^2)$$

$$P_L = \frac{|V_2^-|^2}{2Z_0}(1 - |\Gamma_L|^2)$$

$$P_L = \frac{|V_1^+|^2}{2Z_0}\frac{|S_{21}|^2(1 - |\Gamma_L|^2)}{|1 - S_{22}\Gamma_L|^2} = \frac{|V_S|^2}{8Z_0}\frac{|S_{21}|^2(1 - |\Gamma_L|^2)|1 - \Gamma_S|^2}{|1 - S_{22}\Gamma_L|^2|1 - \Gamma_S\Gamma_{in}|^2}.$$

$$G = \frac{P_L}{P_{in}} = \frac{|S_{21}|^2(1 - |\Gamma_L|^2)}{(1 - |\Gamma_{in}|^2)|1 - S_{22}\Gamma_L|^2}$$

$$P_{avs} = P_{in}\Big|_{\Gamma_{in}=\Gamma_S^*} = \frac{|V_S|^2}{8Z_0}\frac{|1 - \Gamma_S|^2}{(1 - |\Gamma_S|^2)}$$ max power transfer.

$$P_{avn} = P_L\Big|_{\Gamma_L=\Gamma_{out}^*} = \frac{|V_S|^2}{8Z_0}\frac{|S_{21}|^2(1 - |\Gamma_{out}|^2)|1 - \Gamma_S|^2}{|1 - S_{22}\Gamma_{out}^*|^2|1 - \Gamma_S\Gamma_{in}|^2}\Bigg|_{\Gamma_L=\Gamma_{out}^*}$$

$$P_{avn} = \frac{|V_S|^2}{8Z_0}\frac{|S_{21}|^2|1 - \Gamma_S|^2}{|1 - S_{11}\Gamma_S|^2(1 - |\Gamma_{out}|^2)}$$

Power Gain in terms of S-Parameters:

The power gain ➔ $G = \frac{P_L}{P_{in}} = \frac{|S_{21}|^2(1 - |\Gamma_L|^2)}{(1 - |\Gamma_{in}|^2)|1 - S_{22}\Gamma_L|^2}$

- The available and transducer power gains will be:

$$G_A = \frac{P_{avn}}{P_{avs}} = \frac{|S_{21}|^2(1 - |\Gamma_S|^2)}{|1 - S_{11}\Gamma_S|^2(1 - |\Gamma_{out}|^2)} \qquad G_T = \frac{P_L}{P_{avs}} = \frac{|S_{21}|^2(1 - |\Gamma_S|^2)(1 - |\Gamma_L|^2)}{|1 - \Gamma_S\Gamma_{in}|^2|1 - S_{22}\Gamma_L|^2}$$

- A special case of the transducer power gain occurs when both the input and output are matched for zero reflection (in contrast to conjugate matching). Then $\Gamma_L = \Gamma_S = 0$,

$$G_T = |S_{21}|^2$$

- Another special case is the unilateral transducer power gain, G_{TU}, where $S_{12} = 0$ (or is negligibly small) results in $|\Gamma_{in}| = |S_{11}|$.
- This nonreciprocal characteristic is approximately true for many transistors devices. $G_{TU} = \frac{|S_{21}|^2(1 - |\Gamma_S|^2)(1 - |\Gamma_L|^2)}{|1 - S_{11}\Gamma_S|^2|1 - S_{22}\Gamma_L|^2}$

$S_{12} = 0 \Leftarrow$ unilateral انتبه للحالات الخاصة اذا ذكرها بالسؤال مثلاً
وشوف التغيير اللي راح يصير.

To train on using equations.

Example: Comparison of Power Gain Definitions

- A silicon bipolar junction transistor has the following scattering parameters at 1.0 GHz, with a 50 Ω reference impedance:

$$S_{11} = 0.38\angle{-158^\circ} \quad S_{12} = 0.11\angle 54^\circ \quad S_{21} = 3.50\angle 80^\circ \quad S_{22} = 0.40\angle{-43^\circ}$$

The source impedance is Z_S = 25 Ω and the load impedance is Z_L = 40 Ω. Compute G, G_A and G_T

- Solution:

$$\Gamma_S = \frac{Z_S - Z_0}{Z_S + Z_0} = \frac{25 - 50}{25 + 50} = -0.333$$

$$\Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0} = \frac{40 - 50}{40 + 50} = -0.111$$

$$\Gamma_{in} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} = 0.365\angle - 152^\circ$$

$$\Gamma_{out} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S} = 0.545\angle - 43^\circ.$$

$$G = \frac{|S_{21}|^2(1 - |\Gamma_L|^2)}{(1 - |\Gamma_{in}|^2)|1 - S_{22}\Gamma_L|^2} = 13.1$$

lost power @ the load.

$$G_A = \frac{|S_{21}|^2(1 - |\Gamma_S|^2)}{|1 - S_{11}\Gamma_S|^2(1 - |\Gamma_{out}|^2)} = 19.8$$

lost power @ the source.

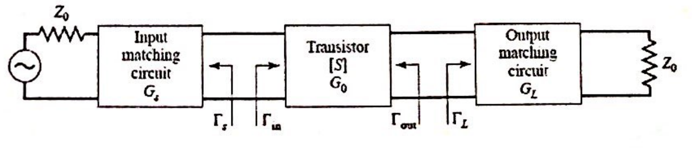
$$G_T = \frac{|S_{21}|^2(1 - |\Gamma_S|^2)(1 - |\Gamma_L|^2)}{|1 - \Gamma_S\Gamma_{in}|^2|1 - S_{22}\Gamma_L|^2} = 12.6$$

lost power @ the load & the source.

Best to have three of them equaled.

General Transistor Amplifier Circuit

- A single-stage microwave transistor amplifier can be modeled by using two matching networks on both sides of the transistor to transform the input and output impedance Z_0 to the source and load impedances Z_S and Z_L.
- The most useful gain definition for amplifier design is the **transducer power gain,** which accounts for both source and load mismatch.
- Separate effective gain factors for the input (source) matching network, the transistor itself, and the output (load) matching network as follows:



$$G_S = \frac{1 - |\Gamma_S|^2}{|1 - \Gamma_{in}\Gamma_S|^2}$$

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

$$G_L = \frac{1 - |\Gamma_L|^2}{|1 - S_{22}\Gamma_L|^2}$$

The overall transducer gain is then $G_T = G_S G_0 G_L$ (as ratio)

or $G_T(dB) = G_S(dB) + G_0(dB) + G_L(dB$ (as in dB values.

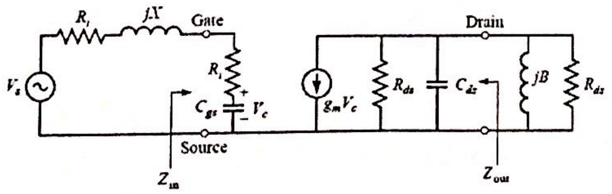
- The effective gains G_S and G_L of the matching networks may be greater than unity. This is because the unmatched transistor would incur power loss due to reflections at the input and output of the transistor, and the matching sections can reduce these losses.

- If the transistor is unilateral, so that $S_{12} = 0$ (or is small enough to be ignored), then $\Gamma_{in} = S_{11}$, $\Gamma_{out} = S_{22}$, and the unilateral transducer gain reduces to $G_{TU} = G_S G_0 G_L$, where:

$$G_S = \frac{1 - |\Gamma_S|^2}{|1 - S_{11}\Gamma_S|^2}$$

• self reading •

- The above results have been derived using the scattering parameters of the transistor, but it is possible to obtain alternative expressions for gain in terms of the equivalent circuit parameters of the transistor.
- As an example, consider the evaluation of the unilateral transducer gain for a conjugately matched FET using the equivalent circuit (with $C_{gd} = 0$).
- Setting the series source inductive reactance $X = 1/\omega C_{gs}$ will make $Z_{in} = Z_S^*$, and setting the shunt load inductive susceptance $B = -\omega C_{ds}$ will make $Z_{out} = Z_L^*$; this effectively eliminates the reactive elements from the transistor equivalent circuit.
- Then by voltage division $V_C = V_S/2j\omega R_i C_{gs}$, and the gain can be easily evaluated as:



$$G_{TU} = \frac{P_L}{P_{avs}} = \frac{\frac{1}{8}|g_m V_c|^2 R_{ds}}{\frac{1}{8}|V_S|^2/R_i} = \frac{g_m^2 R_{ds}}{4\omega^2 R_i C_{gs}^2} = \frac{R_{ds}}{4R_i}\left(\frac{f_T}{f}\right)^2$$

The gain of a conjugately matched FET amplifier drops off as $1/f^2$, or 6 dB per octave

Stability

- Conditions for a transistor amplifier to be stable:
- There are two types of stability:
- **Unconditional stability**: The network is unconditionally stable if $|\Gamma_{in}| < 1$ and $|\Gamma_{out}| < 1$ for all passive source and load impedances (i.e., $|\Gamma_S| < 1$ and $|\Gamma_L| < 1$).
- **Conditional stability**: The network is conditionally stable if $|\Gamma_{in}| < 1$ and $|\Gamma_{out}| < 1$ only for a certain range of passive source and load impedances. This case is also referred to as **potentially unstable**.

→ stable at certain values & unstable at other certain values.

- The stability condition of an amplifier circuit is usually **frequency dependent** since the input and output matching networks generally depend on frequency.
- It is therefore possible for an amplifier to be stable at its design frequency but unstable at other frequencies. Careful amplifier design should consider this possibility.
- The rigorous general treatment of stability requires that the network scattering parameters (or other network parameters) have no poles in the right-half complex frequency plane, in addition to the conditions that $|\Gamma_{in}| < 1$ and $|\Gamma_{out}| < 1$.
- This can be a difficult assessment in practice, but for the special case considered here, where the scattering parameters are known to be **pole free** (as confirmed by measurability), the following stability conditions are adequate.

Stability Circles

- For an amplifier to be unconditionally stable:

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

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

- If the device is unilateral ($S_{12} = 0$), these conditions reduce to the simple results that $|S_{11}| < 1$ and $|S_{22}| < 1$ are sufficient for unconditional stability.
- Finding this range for Γ_S and Γ_L can be facilitated by using the **Smith chart** and plotting the input and output stability circles.
- The stability circles are defined as the loci in the Γ_L (or Γ_S) plane for which $|\Gamma_{in}| = 1$ (or $|\Gamma_{out}| = 1$).
- The stability circles then define the boundaries between stable and potentially unstable regions of Γ_S and Γ_L.
- Γ_S and Γ_L must lie on the Smith chart ($|\Gamma_S| < 1$, $|\Gamma_L| < 1$ for passive matching networks).

Derivation for the Output Stability Circle

- Starting with the condition that $|\Gamma_{in}| = 1$:

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

Define Δ as the determinant of the scattering matrix:

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

$$|S_{11} - \Delta\Gamma_L| = |1 - S_{22}\Gamma_L|$$

Squaring both sides then simplify through completing the square

this is the radius of the circle that has center (C).

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

center. radius.

In the complex plane, an equation of the form $|\Gamma - C| = R$ represents a circle having a center at C (a complex number) and a radius R (a real number).

this is the center (complex number).

$C = |C| \angle \theta_c$

"write this" Equation on Formula Sheet.

- Defines the **output** stability circle with a center C_L and radius R_L, where:

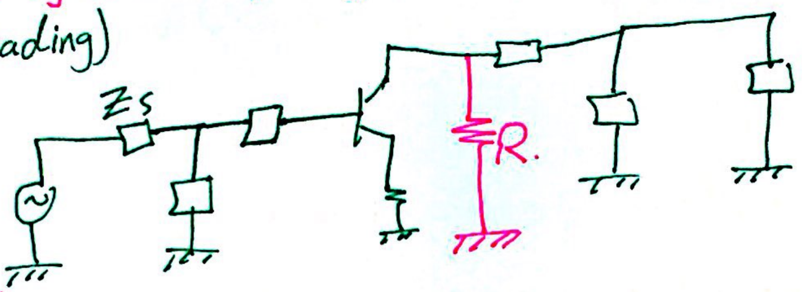
$$C_L = \frac{(S_{22} - \Delta S_{11}^*)^*}{|S_{22}|^2 - |\Delta|^2} \quad \text{(center)} \qquad R_L = \left| \frac{S_{12}S_{21}}{|S_{22}|^2 - |\Delta|^2} \right| \quad \text{(radius)}$$

- Similar results can be obtained for the **input** stability circle by interchanging S_{11} and S_{22}

$$C_S = \frac{(S_{11} - \Delta S_{22}^*)^*}{|S_{11}|^2 - |\Delta|^2} \quad \text{(center)} \qquad R_S = \left| \frac{S_{12}S_{21}}{|S_{11}|^2 - |\Delta|^2} \right| \quad \text{(radius)}$$

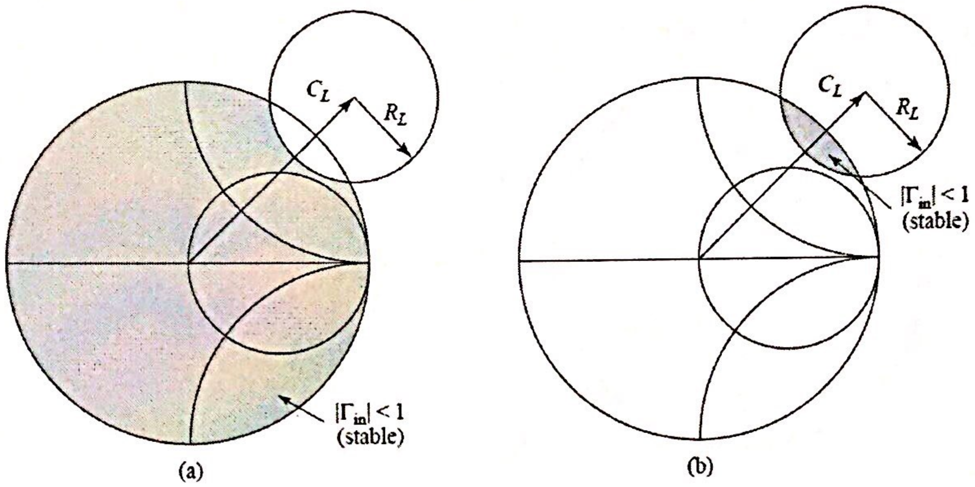
* To have the system Completely stable, we could use: (resistive Loading)

Between the transistor & the output matching network.



Procedure

- Given the scattering parameters of the transistor, we can plot the input and output stability circles to define, where $|\Gamma_{in}| = 1$ and $|\Gamma_{out}| = 1$.
- On one side of the input stability circle we will have $|\Gamma_{out}| < 1$, while on the other side we will have $|\Gamma_{out}| > 1$.
- Similarly, we will have $|\Gamma_{in}| < 1$ on one side of the output stability circle, and $|\Gamma_{in}| > 1$ on the other side.
- We need to determine which areas on the Smith chart represent the stable region, for which $|\Gamma_{in}| < 1$ and $|\Gamma_{out}| < 1$.



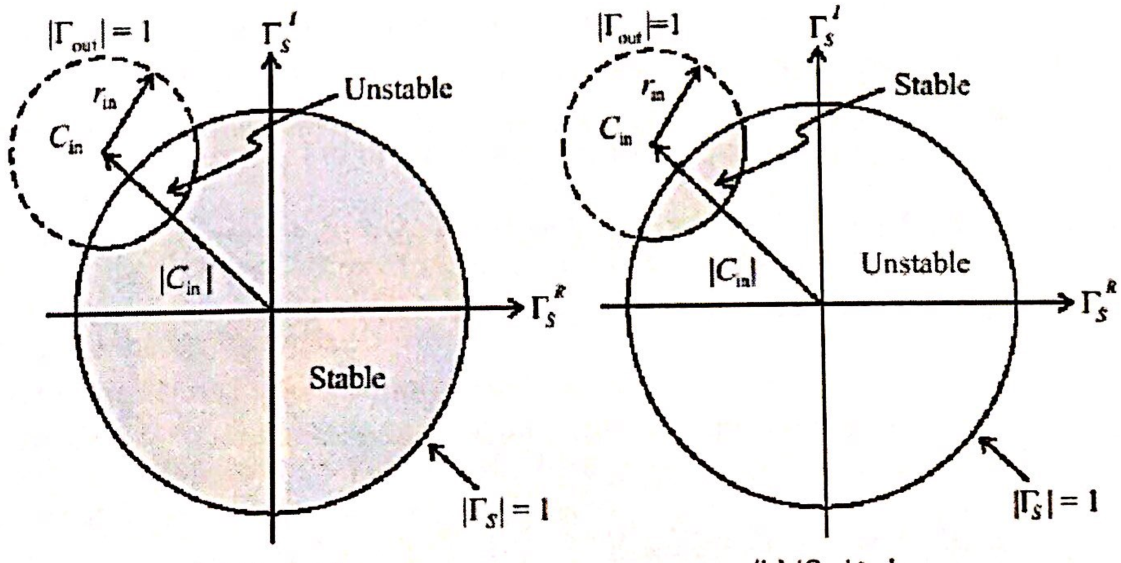
Output stability circles for a conditionally stable device. (a) $|S_{11}| < 1$. (b) $|S_{11}| > 1$.

if $|S_{11}| < 1 \Rightarrow$ stable outside the stability circle.

if $|S_{11}| > 1 \Rightarrow$ stable inside the stability circle.

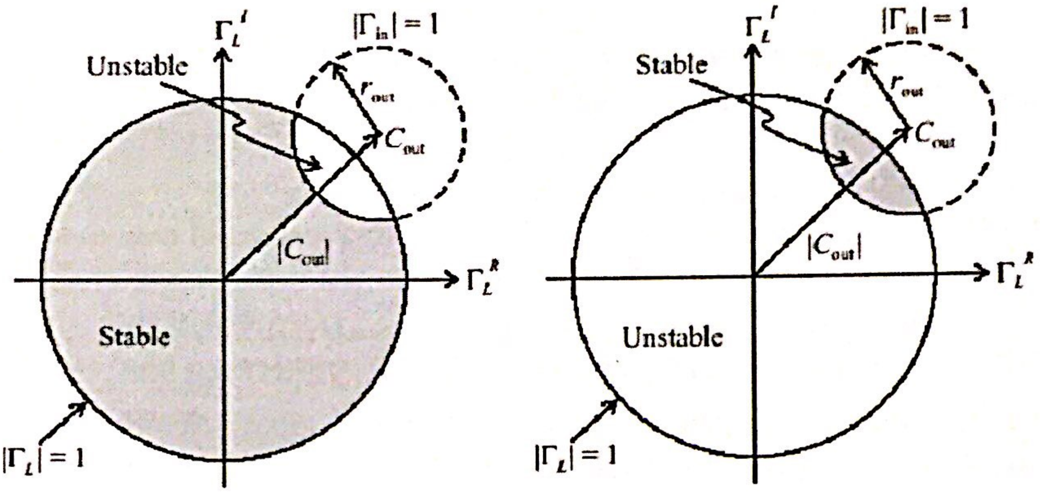
As shown Above in (a) & (b).

Input stability circles denoting stable and unstable regions:



(a) $|S_{22}| < 1$ (b) $|S_{22}| > 1$

Output stability circles denoting stable and unstable regions:



(a) Shaded region is stable, since $|S_{11}| < 1$

(a) Stable region excludes the origin, $\Gamma_L = 0$, since $|S_{11}| > 1$

$|S_{11}| > 1$ $|S_{11}| < 1$

في حال كانت الدائرة كلها بالخارج أو كلها بالداخل منسمي هاي الحالة:

Unconditional Stable.

- If we set $Z_L = Z_0$, then $\Gamma_L = 0$, then $|\Gamma_{in}| = |S_{11}|$.
- Now if $|S_{11}| < 1$, then $|\Gamma_{in}| < 1$, so $\Gamma_L = 0$ must be in a stable region.
- This means that the center of the Smith chart ($\Gamma_L = 0$) is in the stable region, so all of the Smith chart ($|\Gamma_L| < 1$) that is exterior to the stability circle defines the stable range for Γ_L.
- Alternatively, if we set $Z_L = Z_0$ but have $|S_{11}| > 1$, then $|\Gamma_{in}| > 1$ for $\Gamma_L = 0$, and the center of the Smith chart must be in an unstable region.
- In this case the stable region is the inside region of the stability circle that intersects the Smith chart.
- Similar results apply to the input stability circle.
- If the device is unconditionally stable, the stability circles must be completely outside (or totally enclose) the Smith chart.
- We can state this result mathematically as

$$||C_L| - R_L| > 1 \quad \text{for} \quad |S_{11}| < 1 \qquad\qquad ||C_S| - R_S| > 1 \quad \text{for} \quad |S_{22}| < 1$$

- If $|S_{11}| > 1$ or $|S_{22}| > 1$, the amplifier cannot be unconditionally stable because we can always have a source or load impedance of Z_0 leading to $\Gamma_S = 0$ or $\Gamma_L = 0$, thus causing $|\Gamma_{in}| > 1$ or $|\Gamma_{out}| > 1$.
- If the device is only conditionally stable, operating points for Γ_S and Γ_L must be chosen in stable regions, and it is good practice to check stability at several frequencies over the range where the device operates.
- Also note that the **scattering parameters** of a transistor depend on the bias conditions, and so **stability** will also depend on bias conditions.
- If it is possible to accept a design with less than maximum gain, a transistor can usually be made to be unconditionally stable by using resistive loading.

* The following effects the S parameters:

1. Load Condition.
2. Frequancy;
3. Biasing;

Tests for Unconditional Stability

الدائرة هون كلها بالخارج أو كلها بالداخل .

- The stability circles discussed above can be used to determine regions for Γ_S and Γ_L where the amplifier circuit will be conditionally stable, but simpler tests can be used to determine unconditional stability.
- One of these is the $K - \Delta$ test, where it can be shown that a device will be unconditionally stable if Rollet's condition, defined as;

write K & |Δ| on formula sheet.

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}S_{21}|} > 1 \quad \text{where} \quad |\Delta| = |S_{11}S_{22} - S_{12}S_{21}| < 1$$

These two conditions are necessary and sufficient for unconditional stability, and are easily evaluated.

- If the device scattering parameters do not satisfy the K − Δ test, the device is not unconditionally stable, and stability circles must be used to determine if there are values of Γ_S and Γ_L for which the device will be conditionally stable.
- We must have $|S_{11}| < 1$ and $|S_{22}| < 1$ if the device is to be unconditionally stable.

- While the K − Δ test is a mathematically rigorous condition for unconditional stability, it cannot be used to compare the relative stability of two or more devices because it involves constraints on two separate parameters.
- Recently, however, a new criterion has been proposed that combines the scattering parameters in a test involving only a single parameter (μ) defined as;

μ- Test :

$$\mu = \frac{1 - |S_{11}|^2}{|S_{22} - \Delta S_{11}^*| + |S_{12}S_{21}|} > 1$$

- Thus, if $\mu > 1$, the device is unconditionally stable.
- In addition, it can be said that larger values of μ imply greater stability.

Example: Transistor Stability

- The Triquint T1G6000528 GaN HEMT has the following scattering parameters at 1.9 GHz ($Z_0 = 50\ \Omega$):

$S_{11} = 0.869\angle{-159°}$ $\quad$ $S_{12} = 0.031\angle{-9°}$ $\quad$ $S_{21} = 4.250\angle{61°}$ $\quad$ $S_{22} = 0.507\angle{-117°}$

Determine the stability of this transistor by using the K − Δ test and the μ-test, and plot the stability circles on a Smith chart.

- Solution:

$$|\Delta| = |S_{11}S_{22} - S_{12}S_{21}| = 0.336$$

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}S_{21}|} = 0.383$$

$|\Delta| < 1$ but $K < 1$, so the unconditional stability criteria are not satisfied, and the device is potentially unstable.
The stability of this device can also be evaluated using the μ-test, which gives **μ = 0.678**, again indicating potential instability.

- The centers and radii of the stability circles are:

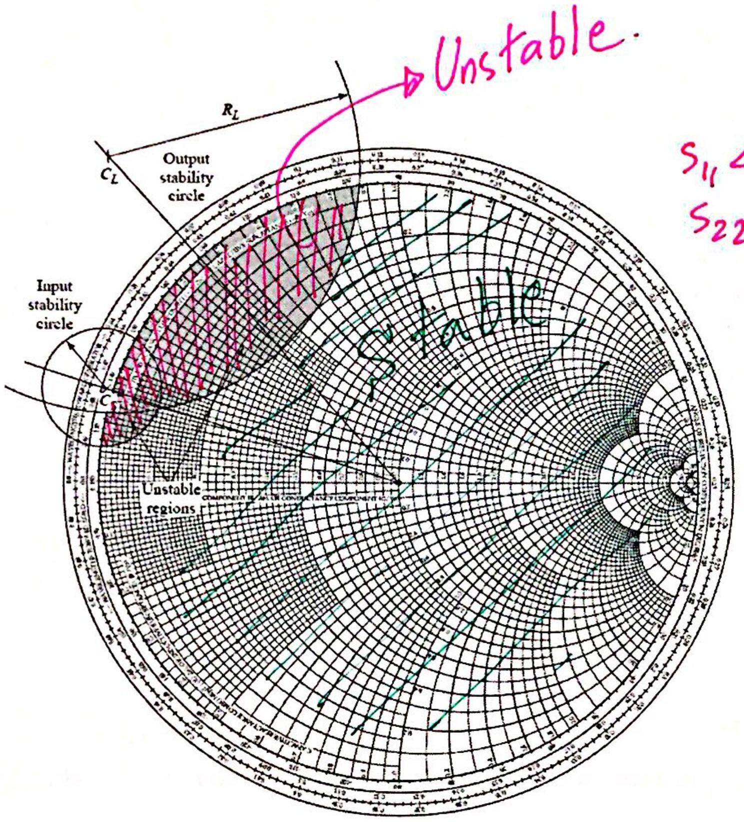
$$C_L = \frac{(S_{22} - \Delta S_{11}^*)^*}{|S_{22}|^2 - |\Delta|^2} = 1.59\angle{132°}$$

$$R_L = \frac{|S_{12}S_{21}|}{|S_{22}|^2 - |\Delta|^2} = 0.915$$

$$C_S = \frac{(S_{11} - \Delta S_{22}^*)^*}{|S_{11}|^2 - |\Delta|^2} = 1.09\angle{162°}$$

$$R_S = \frac{|S_{12}S_{21}|}{|S_{11}|^2 - |\Delta|^2} = 0.205$$

the central part of the Smith chart represents the stable operating region for Γ_S and Γ_L.
The unstable regions are shaded.



$S_{11} < 1$
$S_{22} < 1$

Single-Stage Transistor Amplifier Design

- **Design for Maximum Gain (Conjugate Matching):**
- After the stability of the transistor has been determined and the stable regions for Γ_S and Γ_L have been located on the Smith chart, the input and output matching sections can be designed.
- Since G_0 is fixed for a given transistor, the overall transducer gain of the amplifier will be controlled by the gains, G_S and G_L, of the matching sections.
- Maximum gain will be realized when these sections provide a conjugate match between the amplifier source or load impedance and the transistor.
- Because most transistors exhibit a significant impedance mismatch (large $|S_{11}|$ and $|S_{22}|$), the resulting frequency response may be narrowband.

- Maximum power transfer from the input matching network to the transistor will occur when: $\Gamma_{in} = \Gamma_S^*$
- and that maximum power transfer from the transistor to the output matching network will occur when:

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

- With the assumption of lossless matching sections, these conditions will maximize the overall transducer gain given by:

$$G_{T_{max}} = \frac{1}{1-|\Gamma_S|^2}|S_{21}|^2\frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2}$$

In addition, with conjugate matching and lossless matching sections, the input and output ports of the amplifier will be matched to Z_0.

means NOT unilateral.

- In the general case with a bilateral ($S_{12} \neq 0$) transistor, Γ_{in} is affected by Γ_{out} and vice versa, so the input and output sections must be matched simultaneously. The necessary equations:

$$\Gamma_S^* = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1-S_{22}\Gamma_L} \qquad \Gamma_L^* = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1-S_{11}\Gamma_S}$$

Solve for Γ_S and Γ_L: $\Gamma_S = \frac{B_1 \pm \sqrt{B_1^2 - 4|C_1|^2}}{2C_1}$, $\Gamma_L = \frac{B_2 \pm \sqrt{B_2^2 - 4|C_2|^2}}{2C_2}$

write on formula sheet.

The variables B_1, C_1, B_2, C_2 are defined as:

$$\boxed{\begin{aligned} B_1 &= 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2 & C_1 &= S_{11} - \Delta S_{22}^* \\ B_2 &= 1 + |S_{22}|^2 - |S_{11}|^2 - |\Delta|^2 & C_2 &= S_{22} - \Delta S_{11}^* \end{aligned}}$$

- Solutions to Γ_S and Γ_L are only possible if the quantity within the square root is positive, and it can be shown that this is equivalent to requiring $K > 1$.
- Thus, unconditionally stable devices can always be conjugately matched for maximum gain, and potentially unstable devices can be conjugately matched if $K > 1$ and $|\Delta| < 1$.
- The results are much simpler for the unilateral case. When $S_{12} = 0$, then $\Gamma_S = S_{11}^*$ and $\Gamma_L = S_{22}^*$, and then maximum transducer gain reduces to;

$$\boxed{G_{TU_{max}} = \frac{1}{1-|S_{11}|^2}|S_{21}|^2\frac{1}{1-|S_{22}|^2}}$$

Unilateral.

- If the transistor is unconditionally stable, so that $K > 1$, the maximum transducer power gain can be simply rewritten as follows:

$$G_{T_{max}} = \frac{|S_{21}|}{|S_{12}|}(K - \sqrt{K^2 - 1})$$

- The maximum transducer power gain is also sometimes referred to as the **matched gain**.
- The maximum gain does not provide a meaningful result if the device is only conditionally stable since simultaneous conjugate matching of the source and load is not possible if $K < 1$.
- In this case a useful figure of merit is the **maximum stable gain**, defined as the maximum transducer power gain with $K = 1$. Thus,

$$G_{msg} = \frac{|S_{21}|}{|S_{12}|}$$

أقل قيمة لل stablity عند K=1

MSG ≡ "Max stable gain" occur @ K=1

Example: Conjugately Matched Amplifier Design

- Design an amplifier for maximum gain at 4 GHz using single-stub matching sections.
- Calculate and plot the input return loss and the gain from 3 to 5 GHz. The transistor is a GaAs MESFET with the following scattering parameters (Z_0 =50 Ω):

f(GHz)	S_{11}	S_{12}	S_{21}	S_{22}
3.0	$0.80\angle{-89°}$	$0.03\angle 56°$	$2.86\angle 99°$	$0.76\angle{-41°}$
4.0	$0.72\angle{-116°}$	$0.03\angle 57°$	$2.60\angle 76°$	$0.73\angle{-54°}$
5.0	$0.66\angle{-142°}$	$0.03\angle 62°$	$2.39\angle 54°$	$0.72\angle{-68°}$

Solution:

- Note that K > 1 and |Δ| < 1 at 4 and 5 GHz, so the transistor is unconditionally stable at these frequencies, but it is only conditionally stable at 3 GHz.

f(GHz)	K	Δ
3.0	0.77	0.592 → potentially stable.
4.0	1.19	0.487 → Unconditionally stable.
5.0	1.53	0.418

At 4 GHz:

For maximum gain:

$$\Gamma_S = \frac{B_1 \pm \sqrt{B_1^2 - 4|C_1|^2}}{2C_1} = 0.872\angle 123°$$

$$\Gamma_L = \frac{B_2 \pm \sqrt{B_2^2 - 4|C_2|^2}}{2C_2} = 0.876\angle 61°$$

The Effective gain:

$$G_S = \frac{1}{1 - |\Gamma_S|^2} = 4.17 = 6.20 \text{ dB},$$

$$G_0 = |S_{21}|^2 = 6.76 = 8.30 \text{ dB},$$

$$G_L = \frac{1 - |\Gamma_L|^2}{|1 - S_{22}\Gamma_L|^2} = 1.67 = 2.22 \text{ dB}.$$

The overall transducer gain is:

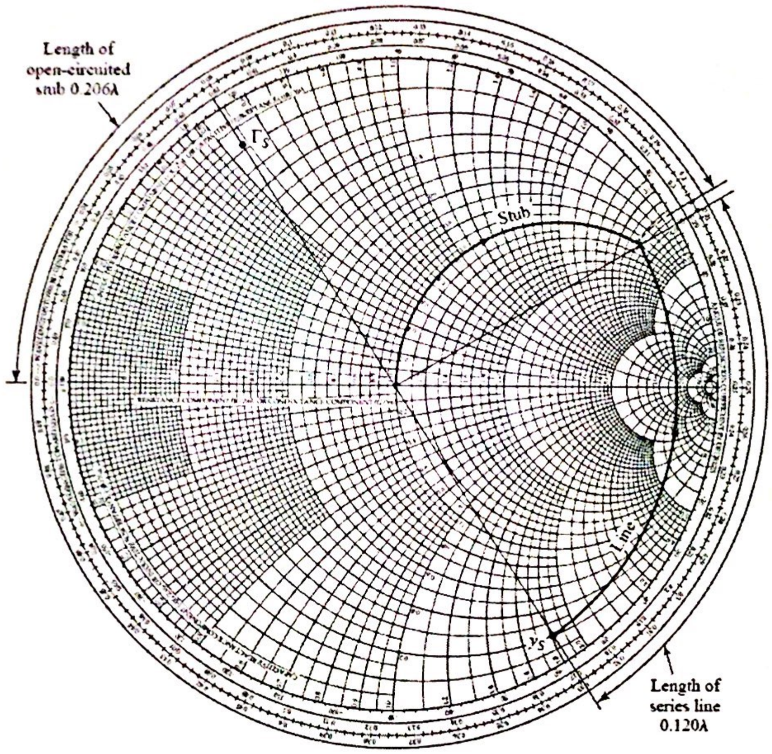
$$G_{T_{max}} = 6.20 + 8.30 + 2.22 = 16.7 \text{ dB}$$

or

$$G_{T_{max}} = 10 \log_{10}(4.17 \times 6.76 \times 1.67)$$

$$= 16.7 \text{ dB}.$$

Scanned by CamScanner

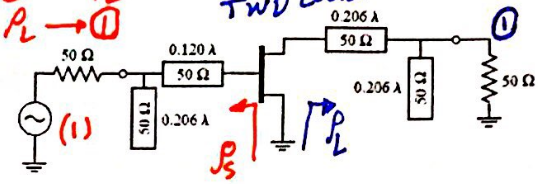


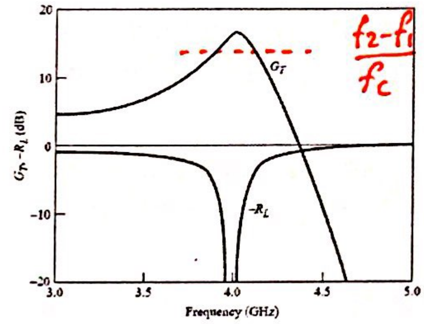
$\rho_s \to (1)$ TWD Gen.

$(1) \to \rho_s$ TWD Load.

$① \to \rho_L$ TWD Gen.

$\rho_L \to ①$ TWD Load.





$\frac{f_2 - f_1}{f_c} = 2.5\%$

The expected gain of 16.7 dB at 4 GHz, with a very good return loss. The bandwidth where the gain drops by 1 dB is about 2.5%.

- The matching networks can easily be determined using the Smith chart.
- For the input matching section, first plot Γ_S.
- The impedance, Z_S, represented by this reflection coefficient is the impedance seen looking into the matching section toward the source impedance, Z_0.
- Thus, the matching section must transform Z_0 to the impedance Z_S. There are several ways of doing this, but we will use an **open-circuited shunt stub** followed by a length of line.
- We convert to the normalized admittance y_s , and work backward (toward the load on the Smith chart) to find that a line of length 0.120λ will bring us to the 1 + jb circle.
- Then we see that the required stub admittance is +j3.5, for an open-circuited stub length of 0.206λ.
- A similar procedure gives a line length of 0.206λ and a stub length of 0.206λ for the output matching circuit.

Constant-Gain Circles and Design for Specific Gain

- In many cases it is preferable to design for less than the maximum obtainable gain, to improve bandwidth or to obtain a specific value of amplifier gain.
- This can be done by designing the input and output matching sections to have less than maximum gains; in other words, mismatches are purposely introduced to reduce the overall gain.
- The design procedure is facilitated by plotting constant-gain circles on the Smith chart to represent loci of Γ_S and Γ_L that give fixed values of gain (G_S and G_L).
- For many transistors $|S_{12}|$ is small enough to be ignored, and the device can be assumed to be unilateral. This greatly simplifies the design procedure.

- The error in the transducer gain caused by approximating $|S_{12}|$ as zero is given by the ratio G_T/G_{TU}; where U is defined as the **unilateral figure of merit,**

$$\frac{1}{(1+U)^2} < \frac{G_T}{G_{TU}} < \frac{1}{(1-U)^2}$$

$$U = \frac{|S_{12}||S_{21}||S_{11}||S_{22}|}{(1-|S_{11}|^2)(1-|S_{22}|^2)}$$

Note that U is real number.

Usually an error of a few tenths of a dB or less justifies the unilateral assumption

$$G_S = \frac{1-|\Gamma_S|^2}{|1-S_{11}\Gamma_S|^2}, \qquad G_L = \frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2}.$$

maximized when $\Gamma_S = S_{11}^*$ and $\Gamma_L = S_{22}^*$

$$G_{S_{max}} = \frac{1}{1-|S_{11}|^2}, \qquad G_{L_{max}} = \frac{1}{1-|S_{22}|^2}.$$

- Define normalized gain factors g_S and g_L as;

$$g_S = \frac{G_S}{G_{S_{max}}} = \frac{1-|\Gamma_S|^2}{|1-S_{11}\Gamma_S|^2}(1-|S_{11}|^2) \qquad g_L = \frac{G_L}{G_{L_{max}}} = \frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2}(1-|S_{22}|^2)$$

Then we have that $0 \le g_S \le 1$ and $0 \le g_L \le 1$

- The center and radius for the input and output constant gain circles:

$$C_S = \frac{g_S S_{11}^*}{1-(1-g_S)|S_{11}|^2}, \qquad C_L = \frac{g_L S_{22}^*}{1-(1-g_L)|S_{22}|^2},$$

$$R_S = \frac{\sqrt{1-g_S}\,(1-|S_{11}|^2)}{1-(1-g_S)|S_{11}|^2} \qquad R_L = \frac{\sqrt{1-g_L}\,(1-|S_{22}|^2)}{1-(1-g_L)|S_{22}|^2}$$

These for Constant gain circles.

DO NOT mix it up with the equations of the stable circle.

- The centers of each family of circles lie along straight lines given by the angle of S_{11}^* or S_{22}^*.
- Note that when g_S (or g_L) = 1 (maximum gain), the radius R_S (or R_L) = 0, and the center reduces to S_{11}^* (or S_{22}^*), as expected.
- In addition, it can be shown that the 0 dB gain circles ($G_S = 1$ or $G_L = 1$) will always pass through the center of the Smith chart.
- These results can be used to plot a family of circles of constant gain for the input and output sections.
- Then Γ_S and Γ_L can be chosen along these circles to provide the desired gains.
- The choices for Γ_S and Γ_L are not unique, but it makes sense to choose points close to the center of the Smith chart to minimize mismatch, and thus maximize bandwidth.
- Alternatively, the input network mismatch can be chosen to provide a low-noise design.

Example: Amplifier Design for Specified Gain

- Design an amplifier to have a gain of 11 dB at 4.0 GHz. Plot constant-gain circles for $G_S = 2$ and 3 dB, and $G_L = 0$ and 1 dB.
- Calculate and plot the input return loss and overall amplifier gain from 3 to 5 GHz. The transistor has the following scattering parameters ($Z_0 = 50\ \Omega$):

f(GHz)	S_{11}	S_{12}	S_{21}	S_{22}
3	$0.80\angle{-90°}$	0	$2.8\angle 100°$	$0.66\angle{-50°}$
4	$0.75\angle{-120°}$	0	$2.5\angle 80°$	$0.60\angle{-70°}$
5	$0.71\angle{-140°}$	0	$2.3\angle 60°$	$0.58\angle{-85°}$

. First we check for the stability:
we Do the K-Δ Test.

since unconditionally stable (No need to draw stability circle).

Solution:

- Since $S_{12} = 0$ and $|S_{11}| < 1$ and $|S_{22}| < 1$, the transistor is unilateral and unconditionally stable at the given frequencies.

$$G_{S_{max}} = \frac{1}{1-|S_{11}|^2} = 2.29 = 3.6\text{ dB}.$$

$$G_{L_{max}} = \frac{1}{1-|S_{22}|^2} = 1.56 = 1.9\text{ dB}.$$

$$G_0 = |S_{21}|^2 = 6.25 = 8.0\text{ dB}$$

$$G_{TU_{max}} = 3.6 + 1.9 + 8.0 = 13.5\text{ dB}$$

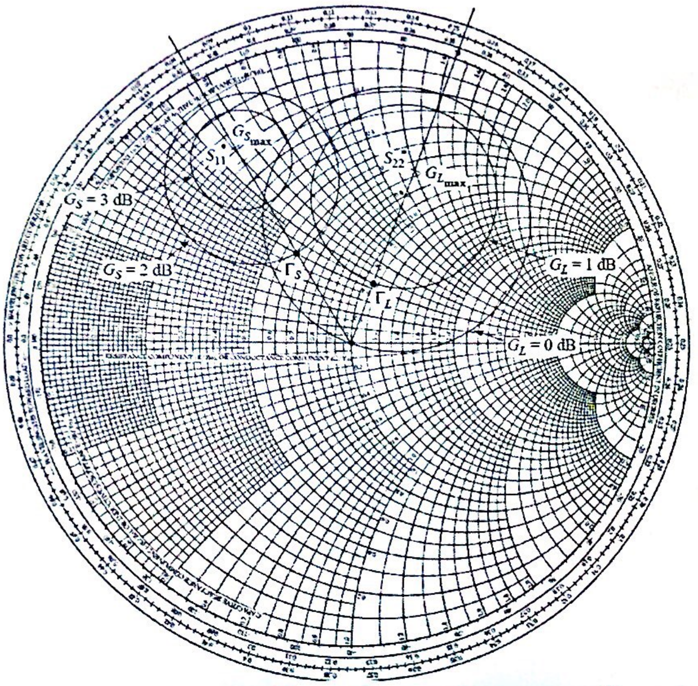
Constant-gain circles:

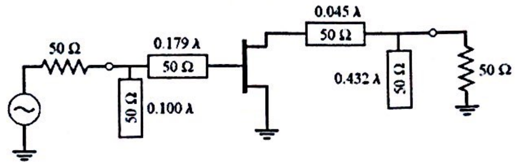
$G_S = 3$ dB $\quad g_S = 0.875 \quad C_S = 0.706\angle 120° \quad R_S = 0.166$
$G_S = 2$ dB $\quad g_S = 0.691 \quad C_S = 0.627\angle 120° \quad R_S = 0.294$
$G_L = 1$ dB $\quad g_L = 0.806 \quad C_L = 0.520\angle 70° \quad R_L = 0.303$
$G_L = 0$ dB $\quad g_L = 0.640 \quad C_L = 0.440\angle 70° \quad R_L = 0.440$

$g_s = \dfrac{G_s\ (\text{Not in dB})}{G_{s\,max}\ (\text{Not in dB})}$

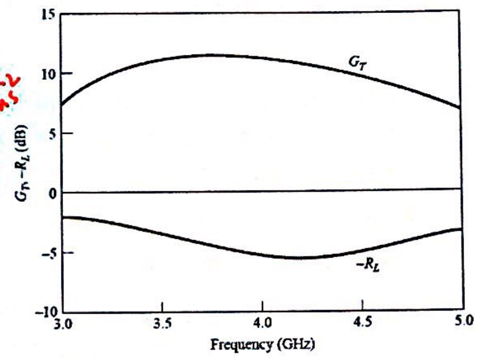
* all the center for source should have same phase shift

& all centers for load should have same phase shift.





$-R_L$ means S_{11}.



The desired gain of 11 dB is achieved at 4.0 GHz

we will take the intersected circles (G_S & G_L)

$G_S = 2$ intersect with $G_L = 1$ & $G_L = 0$ But since need 11 dB

we take $G_S = 2$ & $G_L = 1$ ($8 + 2 + 1$ gives 11 dB) — two solutions

$G_S = 3$ intersect with $G_L = 0$ ($8 + 3 + 0$ gives 11 dB).

* Always By default consider the case shunt-open

⇒ if it wasn't mentioned in the question any certain case.

- We choose $G_S = 2$ dB and $G_L = 1$ dB, for an overall amplifier gain of 11 dB. Then we select Γ_S and Γ_L along these circles as shown, to minimize the distance from the center of the chart (this places Γ_S and Γ_L along the radial lines at 120° and 70°, respectively).
- Thus, $\Gamma_S = 0.33\angle 120°$ and $\Gamma_L = 0.22\angle 70°$, and the matching networks can be designed using shunt stubs as in Example 12.3.
- The response was calculated using CAD software, with interpolation of the given scattering parameter data.
- The bandwidth over which the gain varies by ±1 dB or less is about 25%, which is considerably better than the bandwidth of the maximum gain design in the previous example.
- The return loss, however, is not very good, being only about 5 dB at the design frequency. This is due to the deliberate mismatch introduced into the matching sections to achieve the specified gain.

- For the two solutions, we choose the one that gives a point near to the center so that it give wider BW.

(LNA).

Low-Noise Amplifier Design

- Besides stability and gain, another important design consideration for a microwave amplifier is its noise figure.
- In receiver applications especially it is often required to have a preamplifier with as low a noise figure as possible since, the first stage of a receiver front end has the dominant effect on the noise performance of the overall system.
- Generally it is not possible to obtain both minimum noise figure and maximum gain for an amplifier, so some sort of compromise must be made.
- This can be done by using **constant-gain circles** and **circles of constant noise figure** to select a usable trade-off between noise figure and gain.

Noise Figure

- The noise figure of a two-port amplifier can be expressed as;

$$F = F_{\min} + \frac{R_N}{G_S}|Y_S - Y_{\text{opt}}|^2 \qquad Y_S = \frac{1}{Z_0}\frac{1-\Gamma_S}{1+\Gamma_S} \qquad Y_{\text{opt}} = \frac{1}{Z_0}\frac{1-\Gamma_{\text{opt}}}{1+\Gamma_{\text{opt}}}$$

F_{min}, Y_{opt} & R_N from the Data sheet.
take Γ_{opt} & use it to find Y_{opt}.

$Y_S = G_S + jB_S$ = source admittance presented to transistor.
Y_{opt} = optimum source admittance that results in minimum noise figure.
F_{min} = minimum noise figure of transistor, attained when $Y_S = Y_{opt}$
R_N = equivalent noise resistance of transistor.
G_S = real part of source admittance.

The quantities F_{min}, Γ_{opt}, and R_N are characteristics of the particular transistor being used, and are called the noise parameters of the device; they may be given by the manufacturer or measured.

derivation NOT required.

- We can express:

$$|Y_S - Y_{opt}|^2 = \frac{4}{Z_0^2}\frac{|\Gamma_S - \Gamma_{opt}|^2}{|1+\Gamma_S|^2|1+\Gamma_{opt}|^2}$$

$$G_S = \mathrm{Re}\{Y_S\} = \frac{1}{2Z_0}\left(\frac{1-\Gamma_S}{1+\Gamma_S} + \frac{1-\Gamma_S^*}{1+\Gamma_S^*}\right) = \frac{1}{Z_0}\frac{1-|\Gamma_S|^2}{|1+\Gamma_S|^2}$$

- So;

$$F = F_{min} + \frac{4R_N}{Z_0}\frac{|\Gamma_S - \Gamma_{opt}|^2}{(1-|\Gamma_S|^2)|1+\Gamma_{opt}|^2}$$

→ For a fixed noise figure F we can show that this result defines a circle in the Γ_S plane

→ if F was given mostly it will be required to find Γ_S.

→ if F is required then Γ_S should be given.

- Define the noise figure parameter, N, as;

$$N = \frac{|\Gamma_S - \Gamma_{opt}|^2}{1-|\Gamma_S|^2} = \frac{F - F_{min}}{4R_N/Z_0}|1+\Gamma_{opt}|^2$$

$F_{min} \equiv$ Minimum Noise Figure

- which is a constant for a given noise figure and set of noise parameters

This result defines circles of constant noise figure with:

$$C_F = \frac{\Gamma_{opt}}{N+1}$$

center of the Noise circle.

$$R_F = \frac{\sqrt{N(N+1-|\Gamma_{opt}|^2)}}{N+1}$$

Radius of the Noise circle.

Example: Low-Noise Amplifier Design

- A GaAs MESFET is biased for minimum noise figure, with the following scattering parameters and noise parameters at 4 GHz (Z_0 = 50 Ω):

$S_{12} \neq 0$ (Not unilateral).

$$S_{11} = 0.6\angle{-60°} \qquad S_{12} = 0.05\angle{26°} \qquad S_{21} = 1.9\angle{81°} \qquad S_{22} = 0.5\angle{-60°}$$

- F_{min} = 1.6 dB, Γ_{opt} = 0.62 ∟100°, and R_N = 20 Ω. For design purposes, assume the device is unilateral, and calculate the maximum error in G_T resulting from this assumption. Then design an amplifier having a 2.0 dB noise figure with the maximum gain that is compatible with this noise figure. this is F.
- Use open-circuited shunt stubs in the design of the matching sections.

* 4-steps:

$\boxed{1}$ CHECK for Stability (K-Δ Test)

$\boxed{2}$ CHECK for Unilateral.

$\boxed{3}$ Design Constant Gain Circle.

$\boxed{4}$ Do the Matching.

Solution:

- First calculate that K = 2.78 and Δ = 0.37, so the device is unconditionally stable even without the approximation of a unilateral device.
- Next, compute the unilateral figure of merit from:

$$U = \frac{|S_{12}S_{21}S_{11}S_{22}|}{(1-|S_{11}|^2)(1-|S_{22}|^2)} = 0.059 \qquad \frac{1}{(1+U)^2} < \frac{G_T}{G_{TU}} < \frac{1}{(1-U)^2} \qquad 0.891 < \frac{G_T}{G_{TU}} < 1.130$$

In dB ➔ $-0.50 < G_T - G_{TU} < 0.53$ dB

Thus, we should expect less than about ±0.5 dB error in gain

For 2 dB noise figure circle:

$$N = \frac{F - F_{min}}{4R_N/Z_0}|1+\Gamma_{opt}|^2 = \frac{1.58-1.445}{4(20/50)}|1+0.62\angle 100°|^2 = \underline{0.0986}$$

↳ mag. & phase. Becareful.

$$C_F = \frac{\Gamma_{opt}}{N+1} = 0.56\angle 100°$$

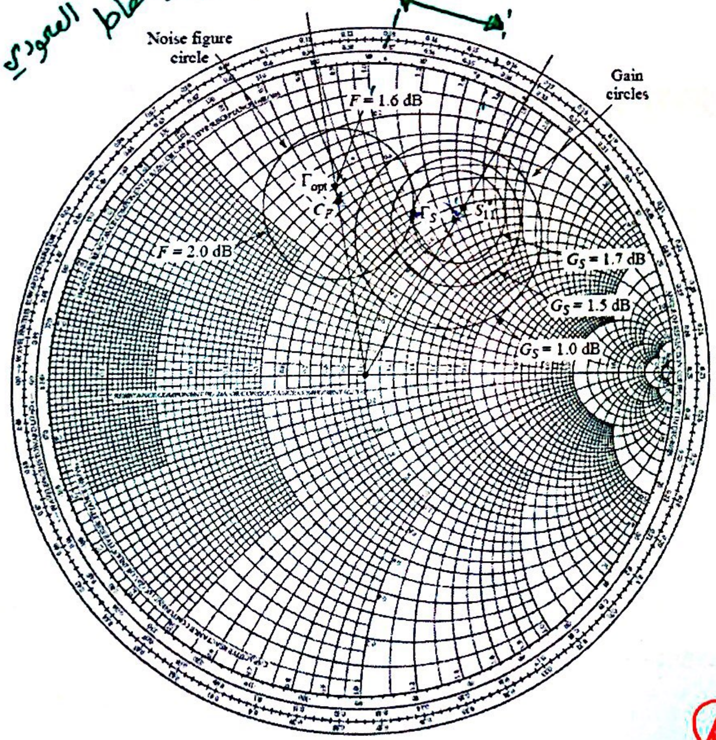
$$R_F = \frac{\sqrt{N(N+1-|\Gamma_{opt}|^2)}}{N+1} = 0.24$$

G_S (dB)	g_S	C_S	R_S
1.0	0.805	0.52∠60°	0.300
1.5	0.904	0.56∠60°	0.205
1.7	0.946	0.58∠60°	0.150

No need to use 3 of them one is enough since you know the angle=60°.

all of them have same phase shift.

أوجد أقصر مسافة من خلال الاسقاط العمودي.



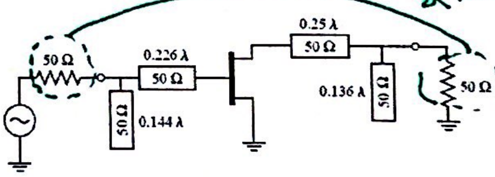
From the Smith chart the optimum solution is Γ_S= 0.53∟75°, yielding G_S= 1.7 dB and F = 2.0 dB.
For the output section we choose $\Gamma_L = S_{22}^* = 0.5∟60°$ for a maximum G_L of:

$$G_L = \frac{1}{1-|S_{22}|^2} = 1.33 = 1.25 \text{ dB}$$

$$G_0 = |S_{21}|^2 = 3.61 = 5.58 \text{ dB}$$

$$G_{TU} = G_S + G_0 + G_L = 8.53 \text{ dB}$$

always take 50 Ω for source & load.



drawing the circuit is required in the Exam.

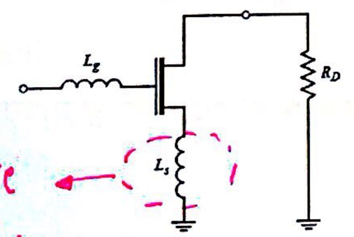
Low-Noise MOSFET Amplifier

- MOSFETs have a relatively low AC input resistance, making them difficult to impedance match.
- An external series resistance can be added to the gate, but this approach increases noise power and degrades efficiency.
- By using a series inductor at the source of a MOSFET, however, it is possible to create a resistive input impedance without adding noisy resistors.
- This technique is called **inductive source degeneration**; similar methods can be used with MESFETs and other transistors.

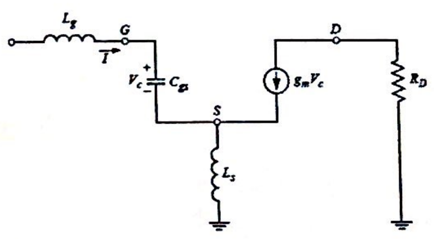
Amplifier Equivalent Circuit

- The model is simplified by assuming the transistor is unilateral, and that R_i, R_{ds}, and C_{ds} can be ignored.
- For an input current I at the gate of the transistor, the capacitor voltage is: $V_c = I/j\omega C_{gs}$

inductive loading :



Basic AC circuit



Equivalent circuit using a simplified unilateral FET model

resistive loading will increase the noise .

- The gate voltage, relative to ground, is then:

$$V = \frac{I}{j\omega C_{gs}} + j\omega L_s (I + g_m V_c) \; = I\left(\frac{1}{j\omega C_{gs}} + j\omega L_s + \frac{g_m L_s}{C_{gs}}\right)$$

- The input impedance at the gate is:

$$Z = \frac{V}{I} = \frac{g_m L_s}{C_{gs}} + j\left(\omega L_s - \frac{1}{\omega C_{gs}}\right)$$

- showing that the circuit has produced an input resistance of $g_m L_s / C_{gs}$. The series inductor (L_s) can be chosen to match the input resistance of the amplifier to a source impedance Z_0.
- The inductor at the gate (L_g) can then be chosen to cancel the residual input reactance, which is usually capacitive.
- The combination of the series matching inductor, the gate capacitance, and the effective input resistance forms a series RLC resonator. The Q of this resonator is:

$$Q = \frac{\omega L_g C_{gs}}{g_m L_s}$$

used when the hybrid model is given.

mostly we won't work on it in this course.

Example: Low-Noise MOSFET Amplifier Design

- An Infineon BF1005 n-channel MOSFET transistor having C_{gs} = 2.1 pF and g_m = 24 mS is used in a 900 MHz low-noise amplifier with inductive source degeneration. Determine the source and gate inductors, and estimate the bandwidth of the amplifier. Assume a source impedance of Z_0 = 50 Ω.
- Solution:
- By matching the input resistance to Z_0

$$L_s = \frac{Z_0 C_{gs}}{g_m} = \frac{(50)(2.1 \times 10^{-12})}{0.024} = 4.37 \text{ nH}$$

$$jX = j\left(\omega L_s - \frac{1}{\omega C_{gs}}\right) = -j59.5 \ \Omega$$

$$L_g = \frac{-X}{\omega} = \frac{59.5}{2\pi(900 \times 10^6)} = 10.5 \text{ nH}$$

$$Q = \frac{\omega L_g C_{gs}}{g_m L_s} = 1.2$$

so the bandwidth of the amplifier could be as high as 80%. This value is probably higher than what would be obtained in practice, due to the approximations that have been made in our analysis

هاي المعادلات تعويض مباشر ماني داعي نكتبهم
على sheet.

Power Amplifiers (1)

- Power amplifiers are used in the final stages of radar and radio transmitters to increase the radiated power level.
- Typical output powers may be on the order of 100–500 mW for mobile voice or data communications systems, or in the range of 1–100 W for radar or fixed point radio systems.
- Important considerations for RF and microwave power amplifiers are efficiency, gain, intermodulation distortion, and thermal effects.
- Single transistors can provide output powers of 10–100 W at UHF frequencies, while devices at higher frequencies are generally limited to output powers less than 10 W.
- Various power-combining techniques can be used in conjunction with multiple transistors if higher output powers are required.

Power Amplifiers (2)

- So far we have considered only small-signal amplifiers, where the input signal power is low enough that the transistor can be assumed to operate as a linear device.
- The scattering parameters of linear devices are well defined and do not depend on the input power level or output load impedance, a fact that greatly simplifies the design of fixed-gain and low-noise amplifiers.
- For high input powers (e.g., in the range of the 1 dB compression point or third-order intercept point), transistors do not behave linearly.
- In this case the impedances seen at the input and output of the transistor will depend on the input power level, and this greatly complicates the design of power amplifiers.

Characteristics of Power Amplifiers and Amplifier Classes

- The power amplifier is usually the primary consumer of DC power in most hand-held wireless devices, so amplifier efficiency is an important consideration.
- One measure of amplifier efficiency is the ratio of RF output power to DC input power:

$$\eta = \frac{P_{out}}{P_{DC}}$$

AC

DC

Three names.

- This quantity is sometimes referred to as drain efficiency (or collector efficiency).
- One drawback of this definition is that it does not account for the RF power delivered at the input to the amplifier.
- Since most power amplifiers have relatively low gains, the efficiency of tends to overrate the actual efficiency.

- A better measure that includes the effect of input power is the power added efficiency, defined as;

$$\eta_{PAE} = \text{PAE} = \frac{P_{out} - P_{in}}{P_{DC}} = \left(1 - \frac{1}{G}\right)\frac{P_{out}}{P_{DC}} = \left(1 - \frac{1}{G}\right)\eta$$

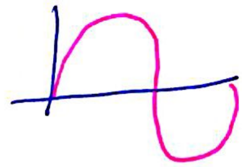
- where G is the power gain of the amplifier.
- Silicon bipolar junction transistor amplifiers in the cellular telephone band of 800–900 MHz band have power added efficiencies on the order of 80%, but efficiency drops quickly with increasing frequency.
- Power amplifiers are often designed to provide the best efficiency, even if this means that the resulting gain is less than the maximum possible.
- Another useful parameter for power amplifiers is the compressed gain (G_1) defined as the gain of the amplifier at the 1 dB compression point.
- Thus, if G_0 is the small-signal (linear) power gain, we have;

Maximum Gain. $|S_{21}|^2$

$$G_1(\text{dB}) = G_0(\text{dB}) - 1$$

$= |S_{21}|^2 - 1$

Complete period



- **Class A amplifiers** are inherently linear circuits, where the transistor is biased to conduct over the entire range of the input signal cycle. Because of this, class A amplifiers have a theoretical maximum efficiency of 50%. Most small-signal and low-noise amplifiers operate as class A circuits.
- In contrast, the transistor in a **class B amplifier** is biased to conduct only during one-half of the input signal cycle. Usually two complementary transistors are operated in a class B push-pull amplifier to provide amplification over the entire cycle. The theoretical efficiency of a class B amplifier is 78%.

- **Class C amplifiers** are operated with the transistor near cutoff for more than half of the input signal cycle, and generally use a resonant circuit in the output stage to recover the fundamental. Class C amplifiers can achieve efficiencies near 100% but can only be used with constant envelope modulations.
- **Higher classes**, such as class D, E, F, and S, use the transistor as a switch to pump a highly resonant tank circuit, and may achieve very high efficiencies.
- The majority of communications transmitters operating at UHF frequencies or above rely on class A, AB, or B power amplifiers because of the need for low distortion products.

see the new slide. ⇒ conduction angle ↓ ⇒ η ↑

Design of Class A Power Amplifiers

- In this section we will discuss the use of large-signal parameters for the design of class A amplifiers.
- Since class A amplifiers are ideally linear, it is sometimes possible to use small signal scattering parameters for design, but better results are usually obtained if large signal parameters are available. As with small-signal amplifier design, the first step is to check the stability of the device. Since instabilities begin at low signal levels, small-signal scattering parameters can be used for this purpose. Stability is especially important for power amplifiers, as high-power oscillations can easily damage active devices and related circuitry.
- The transistor should be chosen on the basis of frequency range and power output, ideally with about 20% more power capacity than is required by the design. Silicon bipolar transistors have higher power outputs than GaAs FETs at frequencies up to a few GHz, and are generally cheaper; GaN HBTs are becoming very popular for high-power applications at RF and low microwave frequencies. Good thermal contact of the transistor package to a heat sink is essential for any amplifier with more than a few tenths of a watt power output. Input matching networks may be designed for maximum power transfer (conjugate matching), while output matching networks are designed for maximum output power (as derived from *L P)*. The optimum values of source and load reflection coefficients are different from those obtained from small-signal scattering parameters via (12.40).
- Low-loss matching elements are important for good efficiency, particularly in the output stage, where currents are highest. Internally matched chip transistors are sometimes available and have the advantage of reducing the effect of parasitic package reactances, thus improving efficiency and bandwidth.

Example: Design of a Class A Power Amplifier

$Z_0 = 50\,\Omega$

- Design a power amplifier at 2.3 GHz using a Nitronex NPT25100 GaN HEMT transistor, with an output power of 10W. The scattering parameters of the transistor for $\underbrace{V_{DS} = 28\text{ V and } I_D = 600\text{ mA}}_{\text{Biasing conditions}}$ are as follows:

$S_{11} = 0.593\angle 178°$, $S_{12} = 0.009\angle -127°$, $S_{21} = 1.77\angle -106°$, and $S_{22} = 0.958\angle 175°$,

and the optimum large signal source and load impedances are $Z_{SP} = 10 - j3\ \Omega$ and $Z_{LP} = 2.5 - j2.3\ \Omega$. For an output power of 10 W, the power G gain is 16.4 dB and the drain efficiency is 26%. Design input and output impedance matching sections for the transistor, and find the required input power, the required DC drain current, and the power added efficiency.

Always: PAE < Drain Efficiency.

Solution:

$\boxed{dBm = 10 \log(P_{in\,mW})}$

firstly check for stability:

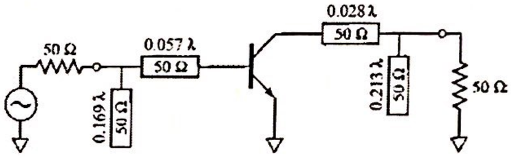
$|\Delta| = |S_{11}S_{22} - S_{12}S_{21}| = 0.579 < 1$

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}S_{21}|} = 2.08 > 1$$

$\left.\begin{array}{l}\Gamma_{SP} = 0.668\angle 187°,\\ \Gamma_{LP} = 0.905\angle -175°\end{array}\right\}$ from Large & small equivalent circuit.

$$\Gamma_S = \frac{B_1 \pm \sqrt{B_1^2 - 4|C_1|^2}}{2C_1} = 0.508\angle 166°$$

$$\Gamma_L = \frac{B_2 \pm \sqrt{B_2^2 - 4|C_2|^2}}{2C_2} = 0.954\angle -176°$$



$P_{in} = P_{out}(\text{dBm}) - G(\text{dB}) = 10\log(10{,}000) - 16.4 = 23.6\ \text{dBm} = 229\ \text{mW}$.

The DC input power can be found from the drain efficiency as $P_{DC} = P_{out}/\eta = 38.5$ W, so the DC drain current is $I_D = P_{DC}/V_{DS} = 1.37$ A.
The power added efficiency of the amplifier:

$$\eta_{PAE} = \frac{P_{out} - P_{in}}{P_{DC}} = \frac{10.0 - 0.229}{38.5} = 25\%$$

$P_{DC} = 38.5W$, $P_{out} = 10W$

The lost power due to the Biasing & matching;

- Note that these values are approximately equal to the large-signal values Γ_{SP} & Γ_{LP}, But not exactly, due to the fact that the scattering parameters used to calculate Γ_S & Γ_L do NOT apply for large power levels.
We should use the large signal reflection coefficients to Design the input & output matching networks.