

- 1) Para diseñar una red de condensadores de desacoplo de alimentación es necesario que su impedancia total equivalente se sitúe por debajo de un determinado valor calculado a partir del máximo rizado permitido y el pico máximo de corriente de consumo. Si el condensador de mayor valor (electrolítico o de tántalo) presenta una SRF baja (10 MHz): ¿Cómo es posible mitigar el carácter inductivo de su impedancia más allá de su SRF?
- a) Empleando una resistencia serie con el condensador de mayor valor
 - b) **Empleando un cierto número de condensadores agrupados en paralelo de SRF mayor (y posiblemente de menor valor nominal).**
 - c) Empleando choques de RF en serie con la fuente principal.
- 2) Respecto de la tensión VDS de saturación en los dispositivos MOSFET indicar cuál de estas afirmaciones es VERDADERA;
- a) **Está relacionada con la velocidad máxima de las cargas en el canal y provoca un efecto de reducción de la ganancia cuando la tensión de polarización VDS supera dicho límite.**
 - b) Su valor aumenta en las tecnologías de canal más corto y ya no presenta ningún problema por debajo de $L=0.35\mu m$.
 - c) Este efecto sólo aparece en las uniones PN y no en el canal de los dispositivos MOSFET.
- 3) En un circuito tanque paralelo R/L/C resonante con Q elevado, respecto el valor de las corrientes por el condensador y la bobina, podemos afirmar que:
- a) La corriente por la bobina es Q^* (corriente por el condensador).
 - b) **Son muy similares en módulo y aproximadamente iguales a Q^* (corriente por la resistencia)**
 - c) Son prácticamente nulas pues en resonancia la impedancia del conjunto es muy elevada.
- 4) En términos de ruido se puede afirmar que:
- a) **Un MOSFET podría potencialmente presentar ventajas frente a un BJT cuando las impedancias de fuente son elevadas.**
 - b) Un MOSFET tendrá mejor comportamiento con impedancias de fuente reducidas
 - c) Un BJT tendrá mejor comportamiento con impedancias de fuente elevadas
- 5) El término Ganancia de Transductor (GT) se calcula como la relación entre:
- a) **La potencia de salida en las condiciones del diseño y la potencia de entrada para condiciones de máxima transferencia**
 - b) La potencia de salida en las condiciones del diseño y la potencia de entrada para condiciones de máxima transferencia
 - c) La potencia de entrada en las condiciones del diseño y la potencia de salida para condiciones de máxima transferencia
- 6) Indicar cuál de estas afirmaciones es correcta:
- a) **Los círculos de Ganancia y NF constante permiten representar las especificaciones de Ganancia y NF sobre una carta Smith para establecer los valores de S (coef. reflexión de la fuente) que cumplen las condiciones requeridas para el diseño**
 - b) Los círculos de Ganancia y NF constante permiten fijar las condiciones de diseño para Máxima Transferencia de Potencia tanto en la entrada como en la salida del amplificador
 - c) Los círculos de Ganancia y NF constante permiten representar las especificaciones de Ganancia y NF sobre una carta Smith para establecer los valores de impedancia de carga a la salida que cumplen las condiciones requeridas para el diseño

- 7) Se pretende obtener el NF de un amplificador en condiciones ideales de adaptación para máxima transferencia de potencia (MTP). Se conocen previamente los datos de ruido total a la salida ($V_{2nout(tctat)}$) y ganancia en tensión (A2), La expresión correcta a emplear es:
- $NF = \frac{V_{(nout(total))}^2}{A^2 \cdot 4kTR}$
 - $NF = \frac{V_{(nout(total))}^2}{A^2 \cdot kTR}$
 - $NF = \frac{V_{(nout(total))}^2}{A \cdot kTR + 2}$
- 8) En el proceso de diseño de un amplificador de bajo ruido LNA a nivel de ASIC (microelectrónica) la adaptación de la impedancia en las líneas internas puede relajarse debido a que:
- Las dimensiones del dispositivo son muy inferiores a la longitud de onda de las frecuencias de trabajo.
 - El ruido es muy inferior ya que las capacidades parásitas son menores.
 - Pueden emplearse técnicas de degeneración inductiva de surtidor para estabilizar los amplificadores.
- 9) Se denomina Corner Frequency (fe) a la frecuencia a la cual:
- El ruido popcorn representa el 50% del ruido del ruido total.
 - El peso del ruido flicker (en rms) se hace despreciable frente al peso del ruido térmico (en rms)
 - La densidad espectral de ruido térmico iguala a la densidad espectral de ruido flicker ($1/f$)
- 10) Con respecto a la técnica de diseño, "Degeneración Inductiva de Surtidor (Emisor)" indicar cuál de estas afirmaciones es FALSA:
- El uso de esta técnica permite incrementar la ganancia con respecto a una configuración en surtidor (emisor) común estándar
 - El valor de las inductancias obtenidas puede ser muy reducido y en ocasiones se emplean los hilos de bonding del encapsulado para implementarlas.
 - Permite fijar la impedancia de entrada (S11) para establecer las condiciones de diseño deseadas (Máxima Transferencia de Potencia, mínimo ruido etc.) y al mismo tiempo también es posible estabilizar el diseño (realimentación)
- 11) En un LNA se busca principalmente un bajo ruido con una ganancia mínima de señal, por lo tanto:
- Suele emplearse una aproximación de diseño para máxima transferencia de potencia (MTP) en la entrada del dispositivo
 - La polarización del dispositivo activo debe realizarse en el punto de trabajo donde ofrezca una ganancia óptima.
 - Las impedancias vistas por el amplificador en la entrada y en la salida deberían minimizar el factor de ruido (NF) manteniendo un mínimo de ganancia.

12) Indicar cuál de las siguientes afirmaciones es CORRECTA:

- a) El diseño de la red de salida de un amplificador tiene un efecto crucial en su NF
- b) El diseño de la red de entrada de un amplificador para conseguir Máxima Transferencia de Potencia implica un NF óptimo.
- c) La red de entrada diseñada, de forma óptima para minimizar el NF no tiene por qué proporcionar una Máxima Transferencia de Potencia.

13) A la hora de diseñar un oscilador se tiene que cumplir:

- a) La ganancia de lazo debe ser exactamente igual a la unidad (0 dB) para una fase de 0 grados. El arranque del oscilador se producirá por la presencia de ruido térmico.
- b) La ganancia de lazo debe ser ligeramente mayor a la unidad para una fase de 0 grados. Cuando la oscilación arranque, la distorsión no lineal provocará una reducción efectiva de ganancia que mantendrá la oscilación.
- c) El ruido del amplificador debe ser elevado para permitir el arranque de la oscilación.

14) En relación con un oscilador de Colpitts indica cuál de las siguientes respuestas es CORRECTA:

- a) La suma de las reactancias de la red de realimentación debe ser nula a la frecuencia de resonancia.
- b) Sólo podemos afirmar que hay dos reactancias inductivas y una capacitiva en la red de realimentación.
- c) La suma de las reactancias de la red de realimentación tiene que ser menor que la unidad a la frecuencia de resonancia.

15) En un oscilador de resistencia negativa podemos afirmar que:

- a) Ninguna de las otras respuestas es correcta.
- b) La resistencia negativa es nula en el estado estacionario.
- c) El valor absoluto de la resistencia negativa aumenta conforme lo hace la potencia de salida del oscilador.

16) Si el coeficiente de reflexión a la entrada de un amplificador es mayor que la unidad podemos afirmar que:

- a) El amplificador es inestable y la resistencia de entrada puede ser negativa o positiva.
- b) El amplificador es incondicionalmente estable.
- c) El amplificador es inestable y la resistencia de entrada es negativa.

17) Los contornos load-pull de un amplificador de potencia se obtienen:

- a) Variando la impedancia de carga y manteniendo la potencia de salida al valor óptimo.
- b) Variando la resistencia de carga y manteniendo el criterio de máxima transferencia de potencia (MTP) a la salida.
- c) Variando la impedancia de carga y manteniendo la potencia de salida constante.

18) Cuál de las siguientes afirmaciones es CORRECTA: "En un amplificador de potencia..."

- a) La resistencia de carga óptima es normalmente igual a la resistencia de generador para conseguir máxima potencia de salida.
- b) La resistencia de carga óptima es normalmente menor a la resistencia de generador para conseguir máxima potencia de salida.
- c) La resistencia de carga óptima es normalmente mayor a la resistencia de generador para conseguir máxima potencia de salida.

19) El rango dinámico lineal:

- a) Está limitado por el noise floor y el punto de intercepción del producto de intermodulación de tercer orden
- b) Está limitado por la ganancia mínima y máxima
- c) **Está limitado por el noise floor y el punto de compresión a 1 dB**

20) En relación con el amplificador de potencia Clase A. podemos afirmar que:

- a) **No tiene excusión dentro de las regiones fuertemente no lineales**
- b) Es el mejor amplificador de potencia en términos de eficiencia.
- c) Es un amplificador estrictamente lineal.

21) El ángulo de conducción en un amplificador de clase AB:

- a) Es de 360º
- b) Es superior a uno de clase A e inferior a uno de clase B
- c) **Es inferior a uno de clase A y superior a uno de clase B**

22) El Human Body Model se emplea en sus distintas variantes para:

- a) Modelizar los efectos de la radiación intensa en un dispositivo amplificador
- b) **Modelizar un evento de descarga electrostática provocada por un humano típico cargado electrostáticamente al tocar un componente**
- c) Fijar las protecciones contra ESD mínimas para un circuito MOSFET en un ambiente industrial

23) En un proceso de fabricación CMOS típico (indicar la VERDADERA):

- a) El primer paso de fabricación corresponde con la deposición de los metales que se encuentran en las capas inferiores.
- b) Los transistores n-MOS y p-MOS son exactamente iguales en el tipo y forma de sus difusiones cambiando únicamente el dopado de óxido en la puerta.
- c) **La primera difusión de impurezas para modificar las propiedades del semiconductor corresponde con la creación de los pozos n-WELL y p-WELL.**

24) Los contornos de Load Pull en un amplificador de potencia:

- a) Siempre aparecen centrados en la línea que marca en la carta las impedancias resistivas puras.
- b) **Pueden aparecer en diferentes zonas de la carta dependiendo de los elementos parásitos del transistor.**
- c) Su trazado es independiente de la frecuencia.

25) La PAE en un amplificador de potencia:

- a) **Relaciona la diferencia entre la potencia de la señal de RF a la entrada y a la salida con la potencia DC**
- b) Relaciona la potencia de salida de la señal de RF con la potencia DC, sin tener en cuenta la potencia de la señal RF de entrada al amplificador.
- c) Relaciona la potencia de entrada de la señal de RF con la potencia DC.

26) En un amplificador clase C:

- a) Su eficiencia es mayor a medida que se reduce su ángulo de conducción, esto es debido a que la potencia de salida también se incrementa a medida que dicho ángulo se reduce.
- b) **Su eficiencia es mayor a medida que se reduce su ángulo de conducción, sin embargo, la reducción del ángulo de conducción también implica una reducción en su potencia de salida.**
- c) Su eficiencia es mayor porque está polarizado justo en el punto central de la recta de carga, donde las excusiones de la señal son máximas, por esta razón su eficiencia aumenta.

27) Si el módulo del parámetro S11 de un amplificador es mayor que la unidad, podemos afirmar que:

- a) La resistencia de entrada es negativa si cargamos el puerto 2 con una carga de 50 Ohm
- b) La resistencia de entrada es negativa si cargamos el puerto 2 con una carga reactiva pura.
- c) La resistencia de entrada es positiva con independencia de lo que coloquemos en el puerto 2.

28) El ruido de fase:

- a) Ninguna de las otras respuestas es correcta.
- b) Se mide en dBm a una determinada frecuencia de separación en frecuencia de la portadora
- c) Es debido a la fluctuación instantánea de la amplitud del oscilador.

29) El equivalente de ruido formado por un conjunto de dos fuentes de ruido tensión serie y corriente paralelo más el circuito sin contribuciones de ruido, tiene como objetivo:

- a) Proporcionar un modelo simple para cálculos de ruido, pero únicamente permite trabajar con una impedancia de fuente de 50 Ohms.
- b) Proporcionar un modelo simple para cálculos de ruido que permita trabajar con independencia del valor de la impedancia de fuente que se conecte a la entrada.
- c) Proporcionar un modelo simple para cálculos de ruido únicamente válido para el caso que la entrada se haya diseñado para MTP.

30) El uso de MOSFET en inversión ultra débil ($I_c < 0.1$)

- a) Es muy empleado en RF debido al ahorro de potencia estática
- b) Es prácticamente imposible de emplear en RF debido al área de puerta y la aparición de capacidades parásitas asociadas.
- c) Los MOSFET no pueden polarizarse en zona ultra débil ya que es propia de los BJT.

31) Sea R_{in} la resistencia de entrada de un dispositivo de resistencia negativa (amplificador cargado) y R_s la resistencia de fuente, podemos afirmar que:

- a) En el estado estacionario $|R_s| > |R_{in}|$
- b) Ninguna de las otras respuestas es correcta
- c) Durante el arranque del oscilador $|R_s| = |R_{in}|$.

32) Los mejores condensadores discretos para aplicaciones de RF son:

- a) Los condensadores de tántalo por su excelente respuesta RF.
- b) Los condensadores de poli propeno debido a sus reducidos parásitos.
- c) Los condensadores cerámicos de tipo COG por su alta Q y alta SRF.

33) En un amplificador LNA, la configuración cascodo permite potencialmente:

- a) Disminuir en un factor 4 el ruido respecto de una configuración surtidor común 5equivalente.
- b) Aumentar la linealidad, disminuyendo el efecto sobre la distorsión de las variaciones en la tensión de drenador.
- c) Reducir el área de silicio y el consumo del amplificador respecto a un surtidor común.

34) Si el ruido de fase de un oscilador a 100 Hz es igual a -80 dBc/Hz significa que:

- a) La densidad de potencia de ruido a 100Hz de la portadora es 80 dB menos que la potencia de salida.
- b) La SNR en una banda de 100 Hz es 80 dB
- c) La Deltaf del ruido FM para 100Hz es 80 dB menor que la portadora
- d) El índice de modulación del ruido en AM a una frecuencia de 100 Hz es 10^{-8}

35) En un oscilador, el factor de calidad del circuito resonante es determinante en:

- a) El ancho de banda del oscilador.
- b) La ganancia mínima del circuito activo que permite la oscilación.
- c) El margen de frecuencia de sintonía.
- d) La potencia disponible en la puerta de salida.

36) ¿Cuándo consideramos que un oscilador es de alto factor de calidad?

- a) Cuando el rendimiento en potencia es superior al 90%.
- b) Cuando la frecuencia Flicker es inferior a $f_0/2Q$.
- c) Cuando $f_0/2Q$ es inferior a la frecuencia Flicker.
- d) Cuando el factor de calidad es superior a 1000.

37) En un amplificador de 26dB de ganancia y $P_{1dB} = 12\text{dBm}$,

- a) 12dBm es la potencia más alta que puede dar a su salida.
- b) La potencia de salida de 12dBm se consigue para una potencia de entrada de $50\mu\text{W}$.
- c) Los productos de intermodulación aparecen para niveles de señal a la salida de 12dBm.
- d) La ganancia se reduce en 1dB cuando la potencia de entrada es 15.8mW.

38) El pulling o, estabilidad de carga en un oscilador, mide

- a) Las variaciones de potencia al variar la corriente de alimentación.
- b) Las variaciones de potencia al variar la impedancia de carga en RF.
- c) La variacion de frecuencia al variar la impedancia de carga en RF.
- d) Las variaciones de frecuencia con la tensión de alimentación.

39) Si el ruido de fase de un oscilador a 1kHz es igual a -110dBc/Hz, significa que:

- a) La potencia de ruido en banda de 1 Hz a 1kHz de la portadora es 110dB menor que la potencia de salida.
- b) La potencia de ruido en una banda de 1kHz es -110dB menor que la potencia total de salida.
- c) La figura del ruido de FM, para una $fm=1$ kHz es 110 dB menor que la frecuencia portadora.
- d) El índice de modulación del ruido de AM con una frecuencia de modulación de 1kHz es de 10^9

40) Un amplificador de potencia clase B tiene como ventaja respecto del clase C,

- a) Tener un rendimiento próximo a la unidad para cualquier señal de entrada.
- b) Tener una respuesta lineal con la amplitud de entrada.
- c) Que no disipa apenas potencia en el dispositivo o amplificador.
- d) Es el más adecuado para amplificar señales moduladas en FM.

41) Un oscilador es un sistema que:

- a) Amplifica, con un nivel de ganancia variable, una señal periódica de frecuencia f.
- b) Genera una señal periódica de frecuencia fa partir de la señal periódica procedente de un cristal de cuarzo.
- c) **Genera una señal periódica de frecuencia f en ausencia de señal periódica a su entrada.**
- d) Filtra el ruido de la señal periódica de frecuencia presente a su entrada en el ancho de banda equivalente de su circuito resonante

42) En un amplificador de potencia clase B, la potencia disipada

- a) **Mínima en ausencia de señal RF**
- b) Máxima para un rendimiento del 80%
- c) Mínima para un rendimiento del 50%
- d) Mínima para el máximo rendimiento

43) Un amplificador clase A. tiene un rendimiento del 50% con una potencia de salida de 5w.

¿Qué potencia disipa cuando no hay señal a su entrada?

- a) 0 W
- b) 5 W
- c) **10 W**
- d) 20 W

44) El ruido de fase de un oscilador es una forma de medir:

- a) Un ruido que puede filtrarse fácilmente.
- b) Una modulación aleatoria que genera un rizado de amplitud
- c) Las bandas laterales del espectro en una modulación DBL
- d) **Una modulación aleatoria de fase sobre la frecuencia de oscilación**

45) Para conseguir que una red oscile al realimentar la salida, la ganancia en lazo abierto debe ser:

- a) **De amplitud unidad o superior y de fase nula.**
- b) De amplitud unidad o superior y de fase -180° .
- c) De amplitud infinita o muy grande y fase cualquiera.
- d) Muy grande y con derivada negativa respecto a la frecuencia

46) El amplificador de clase D disipa muy poca potencia porque:

- a) El circuito resonante serie no disipa energía porque tiene un alto factor de calidad.
- b) La tensión entre drenador y fuente del transistor de polarización es siempre nula.
- c) **Los transistores sólo trabajan en saturación o en corte.**
- d) La resistencia de carga es muy alta y la corriente muy pequeña.

47) La saturación en amplificadores de RF es una forma de distorsión:

- a) Y se puede eliminar con un ecualizador de amplitud adecuado
- b) Es lineal y solo afecta a señales FM
- c) **Es no lineal y no puede compensarse con redes lineales**
- d) Que solo se produce en amplificadores de Clase E

48) Un amplificador de potencia clase A tiene como ventaja respecto del clase C:

- a) Tener un rendimiento próximo a la unidad para cualquier señal de entrada
- b) **Tener una respuesta lineal con la amplitud de entrada**
- c) Que es el más adecuado para amplificar señales moduladas en FM
- d) Que no disipa apenas potencia en el dispositivo amplificador.

- 49) Si la potencia de salida de un amplificador está 20dB por debajo del OIP3:
- a) Los productos de intermodulación están 40dB por debajo del punto de cruce
 - b) Los productos de intermodulación tienen una potencia de -60 dBm.
 - c) No hay productos de intermodulación porque no ha saturado todavía
 - d) **Los productos de intermodulación están 40 dB por debajo de la potencia de salida**
- 50) Si dispone de dos amplificadores de bajo ruido con $F = 2\text{dB}$ y ganancias 10dB y 15dB ¿Cuál colocarías primero?
- a) El de menor ganancia amplifica menos el ruido y consigue menos NF total
 - b) Da lo mismo cual de los dos porque tienen NF igual
 - c) **Poniendo primero el de mayor ganancia**
 - d) El ruido depende del ancho de banda de cada uno, no se puede saber
- 51) El diseño realizado para maximizar la potencia a la salida del amplificador:
- a) **Proporciona ganancias de potencia mayores que el diseño realizado para máxima transferencia de potencia para potencias de entrada iguales o superiores al punto de compresión a 1dB**
 - b) Ninguna de las otras dos respuestas es correcta. El diseño que maximiza la potencia a la salida del amplificador siempre proporciona ganancias de potencia más pequeñas que el diseño para máxima transferencia de potencia incluso cuando se trabaja con potencias elevadas de la señal de entrada
 - c) Proporciona la máxima ganancia de potencia incluso para valores pequeños de potencia de entrada (pequeña señal)
- 52) Según el planteamiento de la resistencia negativa: la DEVICE LINE tiene una fuerte dependencia con:
- a) La frecuencia (efecto Q)
 - b) La amplitud (distorsión)
 - c) **La tensión de alimentación**
- 53) La resistencia óptima a la salida de un amplificador de potencia es aquella que:
- a) **Permite extraer del transistor máxima corriente y máxima tensión simultáneamente**
 - b) Permite extraer del transistor la máxima corriente, pero no la máxima tensión
 - c) Permite extraer del transistor la máxima tensión, pero no la máxima corriente
- 54) Indica la INCORRECTA:
- a) El ruido shot (rms) depende de la corriente de polarización del dispositivo con la raíz de la corriente de polarización
 - b) El ruido térmico en un MOSFET es superior al shot
 - c) **El ruido flicker (1/f) siempre es superior al ruido térmico a cualquier frecuencia**
- 55) En un circuito resonante serie RLC con Q elevado, respecto a las tensiones de condensador y bobina indica:
- a) La corriente por la bobina es $Q^*(\text{corriente condensador})$
 - b) Son prácticamente nulas pues en resonancia la impedancia del conjunto es muy elevada
 - c) **Son muy similares en módulo y aproximadamente iguales a Q^*V_{in} (V_{in} es la tensión de R)**

56) En una red de condensadores para desacoplo de alimentación

- a) El objetivo es que su impedancia equivalente tenga un valor máximo tal que los picos de consumo de corriente no produzcan rizado de tensión superior a un máximo especificado.
- b) El objetivo es que su impedancia equivalente tenga un valor superior a un mínimo establecido para actuar de filtro paso bajo junto con los parásitos de los pines de alimentación
- c) El objetivo de su diseño es filtrar las componentes de baja frecuencia para evitar el crosstalk en los canales de salida

57) En un proceso CMOS típico, indica la VERDADERA

- a) El primer proceso de difusión de impurezas se emplea para crear terminales DRAIN/SOURCE para los MOSFET pMOS
- b) La deposición de metales para interconexiones se realiza una vez terminado el procesado de los transistores nMOS y pMOS
- c) Los transistores nMOS y pMOS son iguales en tipo y forma de difusiones cambiando únicamente el dopado de la puerta.

58) Los transistores MESFET para aplicaciones RF:

- a) Tienen una estructura similar a los MOSFET, pero el aislante de puerta se sustituye por una unión metal semiconducto polarizada en inversa.
- b) Tienen una estructura similar a los MOSFET, pero el aislante de la puerta se sustituye por una resistencia de alto valor
- c) Son propios de aplicaciones de baja frecuencia y alta potencia

59) Sea un oscilador cuyo ruido de fase ha sido medido a una separación f_A y f_B de la frecuencia de oscilación, siendo $f_A > f_B$

- a) El ruido de fase es mayor en f_B
- b) El ruido de fase es mayor en f_A
- c) El ruido de fase es igual a ambas frecuencias

60) Las condiciones de oscilación en el estado estacionario se satisfacen cuando:

- a) La ganancia de lazo de realimentación es igual a la unidad y el desfase nulo
- b) La ganancia del lazo de realimentación es mayor que la unidad y el desfase 180°
- c) La ganancia del lazo de realimentación es menor que la unidad y el desfase nulo.

61) El ruido de fase:

- a) A frecuencias elevadas es predominantemente shot
- b) Es debido a la modulación del ruido $1/f$
- c) Tiene su origen en el ruido térmico del amplificador.

62) En un proceso de fabricación CMOS típico (indicar la VERDADERA):

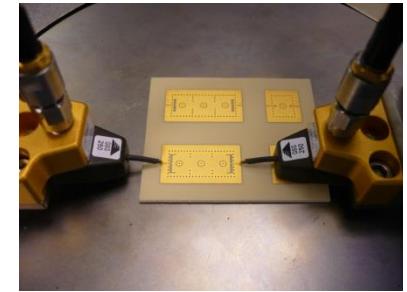
- a) Cada uno de los pasos de ataque o difusión del proceso de fabricación emplean técnicas fotolitográficas
- b) El desarrollo de las máscaras para los procesos fotolitográficos supone el menor de los costes en la manufactura de semiconductores
- c) La menor resolución alcanzable en la actualidad es de 350nm.

63) Teniendo disponibles para un mismo condensador cerámico los encapsulados SMD 0201 y 0805, indica cuál de ellos tendrá una SRF más elevada y por qué:

- a) Aquel de tamaño 0805 por estar adaptado a la impedancia de referencia.
- b) Aquel de tamaño 0021 por tener menores parásitos
- c) Aquel de tamaño 0201 por presentar menos resistividad en sus contactos.



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Chapter 5 – Oscillators (I)

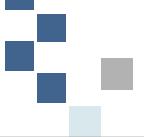
Feedback oscillators. Negative resistance oscillators.

High Frequency Electronic Circuits
Máster Universitario en Ingeniería de Telecomunicación (MUIT)



■ Outline

- Basic concepts
- Positive Feedback Oscillators
- Negative Resistance Oscillators
- Phase Noise



- Basic concepts



■ Introduction

- Oscillators are **one-terminal** circuits producing a periodical output upon power up
- They are used in numerous applications: signal sources, modulation and demodulation, frequency conversion,...,
- They can be generally divided into two types:
 - R-C or relaxation oscillators
 - L-C or harmonic oscillators
- Relaxation oscillators have two unstable states and the circuit switches back-and-forth between them usually producing square waves

- **Harmonic oscillators** can produce nearly sinusoidal outputs due to the use of resonant L-C selective circuits and can be seen from two different perspectives:
 - Positive feedback oscillators (Control Theory)
 - Negative resistance oscillators
- Typical resonant circuits are based on lumped elements and quartz crystal (< 500 MHz), transmission lines (<5 GHz) or dielectric resonators (from 2 up to 40 GHz).



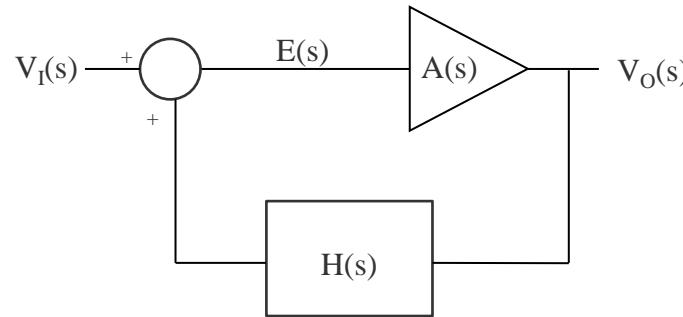
- Positive feedback oscillators

Positive Feedback Oscillators

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■ Principles of Operation

- This is the classical approach based on **control theory** concepts
- If we *assume* a high impedance at the input of both feedback network $H(s)$ and amplifier $A(s)$ a simple transfer function can be extracted



$$V_0(s) = A(s)V_I(s) + H(s)A(s)V_o(s)$$

$$\frac{V_o(s)}{V_I(s)} = \frac{A(s)}{1 - A(s)H(s)}$$

Loop gain

- We can get a non-zero output at $V_O(s)$ if loop gain $A(s)H(s) = 1$, then the circuit oscillates at $s=j\omega$ where this condition holds
- $V_I(s)$ is either thermal noise or the step response due to circuit power up

Positive Feedback Oscillators

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■ Principles of Operation

- The conditions for sustained oscillation are from the positive feedback perspective:

Barkhausen criterion

$$\Re(A(s)H(s)) = 1$$

$$\Im(A(s)H(s)) = 0 \longleftarrow$$

Total shift around
the loop is zero

- However, for oscillation to start-up the loop gain must be greater than unity due to gain reduction when the oscillator begins to saturate

$$|A(s)H(s)| > 1 \longleftarrow \text{Start-up condition}$$

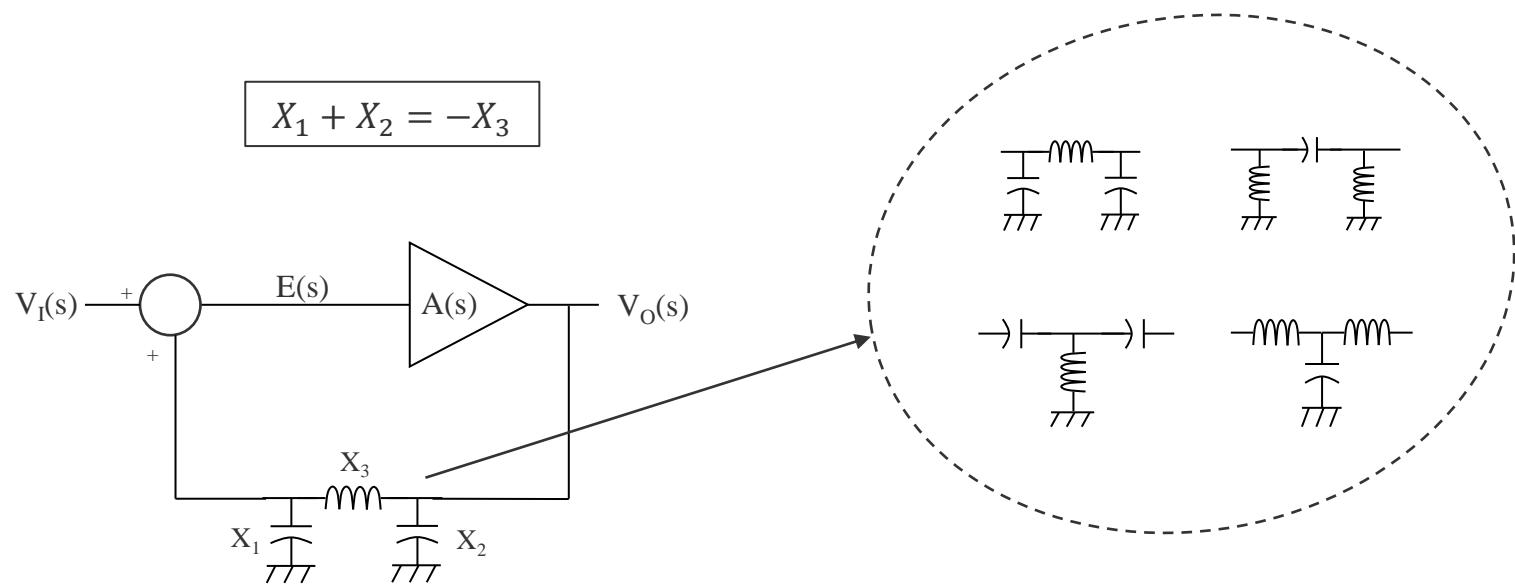
- The steady state condition is then reached when $A(s)H(s) = 1$
- For preventing multiple simultaneous oscillations, the Barkhausen criterion must be satisfied at **just one frequency**
- Usually the amplifier $A(s)$ is designed as a wideband device, while the oscillation frequency is selected by the feedback network $H(s)$

Positive Feedback Oscillators

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■ Principles of Operation

- In general, Π or T networks are used for implementing the feedback network $H(s)$
- It can be demonstrated that reactances X_1 , X_2 and X_3 must fulfill the following condition:

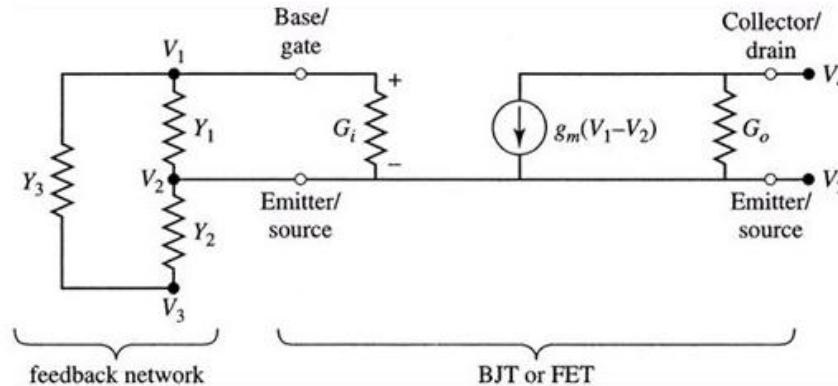


Positive Feedback Oscillators

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■ General Analysis for Transistor Oscillators

- A general analysis can be performed for transistor oscillator circuits



- We have assumed a unilateral transistor, real input/output admittances and a bridged-T feedback network based on reactive elements Y_1 , Y_2 and Y_3
- Common-emitter/source, common-base/gate and common-collector/drain configurations can be obtained by grounding V_2 , V_1 or V_4 respectively
- A feedback path can be introduced by connecting the corresponding input/output nodes

Positive Feedback Oscillators

■ General Analysis for Transistor Oscillators

- The indefinite admittance matrix can be computed by applying KCL to the four nodes of the circuit

$$\sum_{i=0}^k I_k = 0 \quad \xrightarrow{\hspace{1cm}} \quad \begin{bmatrix} (Y_1 + Y_3 + G_i) & -(Y_1 + G_i) & -Y_3 & 0 \\ -(Y_1 + G_i + g_m) & (Y_1 + Y_2 + G_i + G_o + g_m) & -Y_2 & -G_o \\ -Y_3 & -Y_2 & (Y_2 + Y_3) & 0 \\ g_m & -(G_o + g_m) & 0 & G_o \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \end{bmatrix} = 0$$

- From this 4-port admittance matrix, a simplified 2-port matrix can be obtained by considering the grounded node ($V_i = 0$) and the feedback path ($V_j = V_k$)
 - If a node is grounded the corresponding row and column can be *annihilated*
 - If two nodes are short-circuited the corresponding rows and columns can be *combined*
 - Thus, the order of the matrix can be reduced by two in a feedback transistor oscillator

Positive Feedback Oscillators

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- Example: BJT Common Emitter

- In this case we have $V_2 = 0$ and collector feedback $V_3 = V_4$
- For a bipolar transistor we can assume $G_o = 0$
- We can reduce the indefinite admittance matrix of the general equivalent circuit as

$$\begin{bmatrix}
 (Y_1 + Y_3 + G_i) & -(Y_1 + G_i) & -Y_3 & 0 \\
 (Y_1 + G_i + g_m) & (Y_1 + Y_2 + G_i + C_o + g_m) & Y_2 & C_o \\
 -Y_3 & -Y_2 & (Y_2 + Y_3) & 0 \\
 g_m & -(G_o + g_m) & 0 & G_o
 \end{bmatrix}
 \begin{bmatrix}
 V_1 \\ V_2 \\ V_3 \\ V_4
 \end{bmatrix} = 0$$

↓

$$\begin{bmatrix}
 (Y_1 + Y_3 + G_i) & -Y_3 \\
 (g_m - Y_3) & (Y_2 + Y_3)
 \end{bmatrix}
 \begin{bmatrix}
 V_1 \\ V
 \end{bmatrix} = 0$$

- If the feedback network is based on purely reactive elements then $Y_1 = jB_1$, $Y_2 = jB_2$ and $Y_3 = jB_3$



Positive Feedback Oscillators

■ Example: BJT Common Emitter

- The former equations must be satisfied for non-zero values of V_1 and $V = V_3 = V_4$, so the determinant of the matrix must be zero

$$\begin{vmatrix} G_i + j(B_1 + B_3) & -jB_3 \\ g_m - jB_3 & j(B_2 + B_3) \end{vmatrix} = 0$$

- By separating the real and imaginary parts of the determinant we can get the following two equations:

$$\frac{1}{B_1} + \frac{1}{B_2} + \frac{1}{B_3} = 0$$

Set the oscillation frequency

$$\frac{1}{B_3} + \left(1 + \frac{g_m}{G_i}\right) \frac{1}{B_2} = 0$$

Necessary condition for oscillation

- Similar approach can be applied to other transistors (e.g. FET with $G_i = 0$)

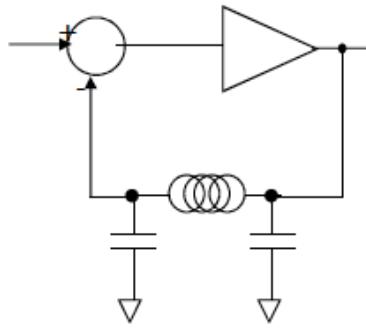
Positive Feedback Oscillators

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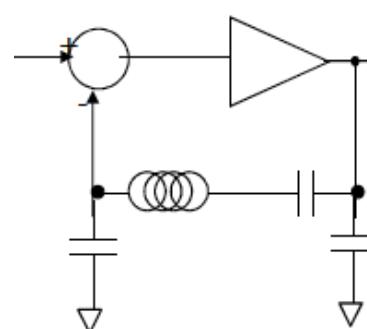
- Example: BJT Common Emitter

- Different oscillators depending on the feedback network implementation

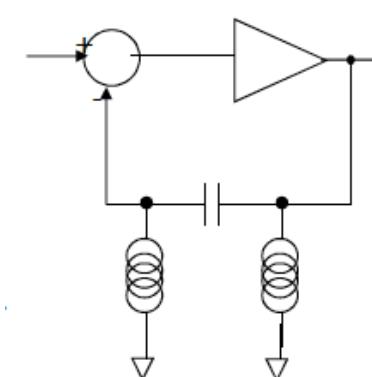
Colpitts



Clapp



Hartley



- For instance, in a Colpitts oscillator

$$\omega_0 = \sqrt{\frac{1}{L_3} \left(\frac{C_1 + C_2}{C_1 C_2} \right)}$$

$$\frac{C_2}{C_1} = \frac{g_m}{G_i}$$

Positive Feedback Oscillators

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■ Summary

- At higher frequencies, the loading effect of the amplifier and feedback network cannot be neglected and must be included in the actual loop gain equation.
- Obtaining loop gain can be difficult at very high frequencies.
- Multiple feedback paths can be present due to parasitic elements and couplings. It can be very difficult to distinguish between amplifier and feedback path in some cases.
- Eventually, the positive feedback approach is restricted to operating frequencies below 500 MHz.



- Negative resistance oscillators

Negative Resistance Oscillators

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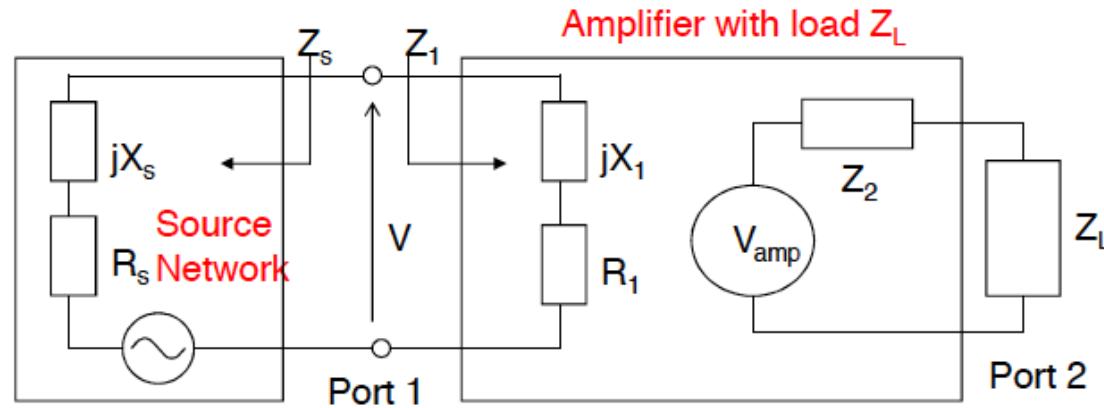
■ Introduction

- The oscillator can be seen as an amplifier producing an output when no input is present → *Unstable amplifier*
- For destabilizing an amplifier, the load Z_L or source Z_S impedance is chosen within the unstable regions resulting in $|\Gamma_1| > 1$ or $|\Gamma_2| > 1$ respectively
- A reflection coefficient magnitude greater than one corresponds to a negative resistance on that port
- For instance, if a load impedance has been chosen for $|\Gamma_1| > 1$ then the source impedance Z_S is designed to enforce $|\Gamma_1 \Gamma_s| > 1$
- In a fixed frequency oscillator, this condition only holds at one frequency

Negative Resistance Oscillators

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■ Negative Resistance Analysis



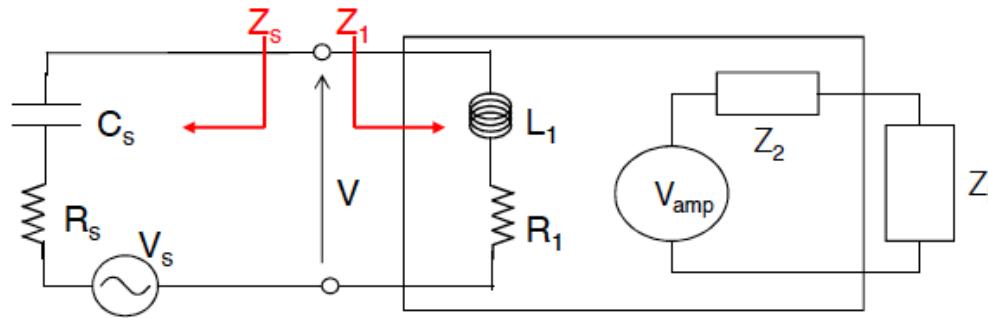
- Oscillation conditions can be studied in terms of terminal impedances
- Consider a series network for modelling source and input of the amplifier

$$V = \frac{R_1 + jX_1}{R_s + R_1 + j(X_s + X_1)} \cdot V_s$$

Negative Resistance Oscillators

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- Negative Resistance Analysis



- Let consider without loss of generality a series R-C network for the source and a series R-L network for the amplifier circuit
- In the Laplace domain the voltage $V(s)$ can be expressed as

$$V(s) = \frac{R_1 + sL_1}{R_s + R_1 + sL_1 + \frac{1}{sC_s}} \cdot V_s(s)$$

Negative Resistance Oscillators

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- Negative Resistance Analysis

- The former equation can be written in a canonical form as:

$$\frac{V(s)}{V_s(s)} = \frac{1}{L_1} \frac{s(R_1 + sL_1)}{s^2 + s(\frac{R_1 + R_s}{L_1}) + \frac{1}{L_s C_s}} = \frac{sC_s(R_1 + sL_1)\omega_n^2}{s^2 + 2\omega_n\delta s + \omega_n^2}$$

$$\delta = \frac{R_1 + R_s}{\sqrt[2]{L_1/C_s}} \quad \text{Damping Factor} \qquad \omega_n = \frac{1}{\sqrt[2]{L_1 C_s}} \quad \text{Natural Frequency}$$

- Corresponds to a 2nd order system with two poles given by

$$p_{1,2} = -\delta\omega_n \pm \omega_n\sqrt{\delta^2 - 1}$$

- Thus, if $(R_1 + R_s) < 0$ then the damping factor δ is negative. If the system is *under-damped* $|\delta| < 1$, the poles will be complex and located at the right-hand side of the complex plane
- A proper negative value of R_1 can be obtained by conveniently modifying the amplifier circuit: introducing local positive feedback or choosing a convenient Z_L

Negative Resistance Oscillators

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■ Negative Resistance Analysis

- A small perturbation will result in an oscillating signal that grows exponentially with frequency:

$$\omega_n \sqrt{\delta^2 - 1}$$

- When the signal amplitude increases, non-linear phenomena will occur limiting the gain of the device and reducing the magnitude of the negative resistance R_1
- At steady-state the damping factor is $\delta = 0$ and the poles are then located on the imaginary axis
- The steady-state oscillation frequency is the natural frequency of the system and corresponds to

$$\omega_n^2 = \frac{1}{L_1 C_s} \Rightarrow X_1 + X_s = 0$$

Negative Resistance Oscillators

■ Summary

- Conditions for the oscillation to start-up and reach steady-state

$$(R_1 + R_s)|_{\omega_0} < 0$$

Start-up

$$(R_1 + R_s)|_{\omega_0} = 0$$

Steady-state

$$(X_1 + X_s)|_{\omega_0} = 0$$

- The steady-state frequency of oscillation is fixed by the series network that is also called the resonator circuit. It has a direct impact on the performance of the oscillator (i.e. phase noise)
- Oscillation will always occur at both ports simultaneously

Negative Resistance Oscillators

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■ Design Procedure

Step 1) Design the amplifier circuit

Step 2) De-stabilize the amplifier by introducing local feedback

Step 3) Obtain the active circuit S-parameters at the desired frequency

Step 4) Choose Γ_L in the unstable region to fix $R_1 < 0$ with the aid of load stability circles

Step 5) Compute $Z_1 = R_1 + jX_1$ and find R_s and X_s in order to meet the start-up oscillation conditions. A rule of thumb is typically used for fixing R_s at one-third of the initial negative resistance

Step 6) Design proper impedance matching networks for Z_s and Z_L



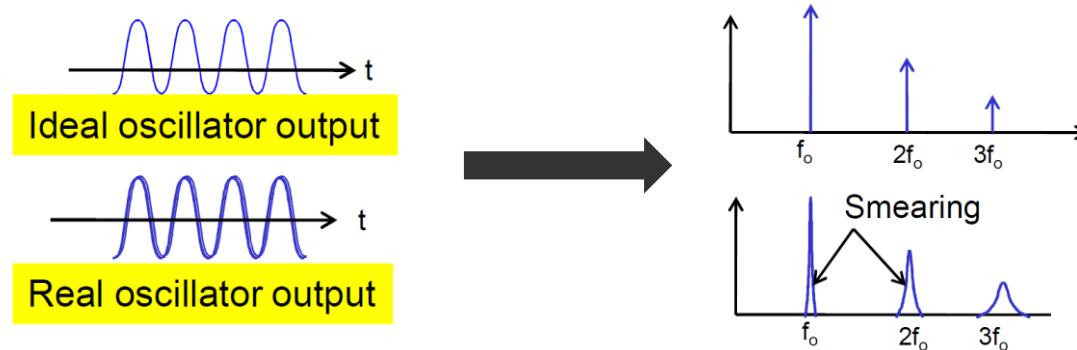
- Phase Noise

■ Introduction

- Oscillators should theoretically produce a series of harmonics in frequency domain
- However, in real oscillators the instantaneous frequency and magnitude of the oscillation are not constant

$$v_{out}(t) = (V_0 + V_n(t)) \cos(\omega t + \theta + \theta_n(t))$$

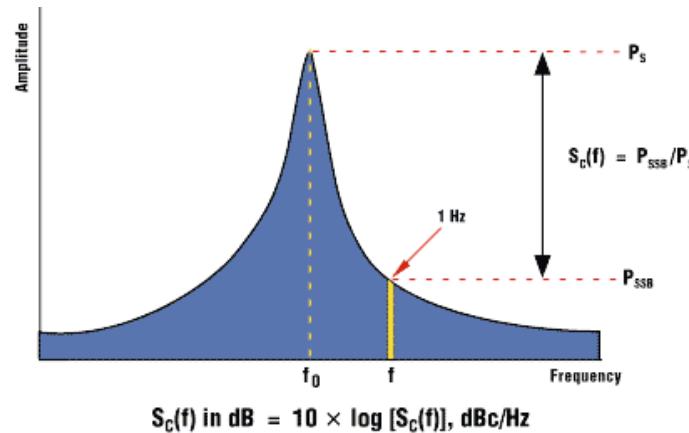
- These random fluctuations represent **noise** contributions, and translate to the frequency domain





■ Basic Concepts

- There is both magnitude and phase modulation of the otherwise harmonic output
- Magnitude fluctuation is kept small by the non-linear self-limiting process of the oscillator and it can be usually neglected.
- Instantaneous phase variation is the main contributor to frequency smearing and it is known as **Phase Noise**
- It is measured with respect to the signal level at a particular offset frequency
- The units are dBc/Hz @ f_{offset} and the power is measured at a 1 Hz resolution bandwidth





■ Basic Concepts

- Considering these small frequency fluctuations as a frequency modulation (ω_m) of the carrier, we would have

$$\theta(t) = \theta_p \sin \omega_m t$$

$$v_{out}(t) = V_0 (\cos \omega t \cos(\theta_p \sin \omega_m t) - \sin \omega t \sin(\theta_p \sin \omega_m t))$$

- For $\theta_p \ll 1$ the former expression can be written as

$$v_{out}(t) = V_0 \left[\cos \omega t - \frac{\theta_p}{2} [\cos(\omega + \omega_m)t - \cos(\omega - \omega_m)t] \right]$$

- It results in two sidebands located on either side of the carrier at frequency ω

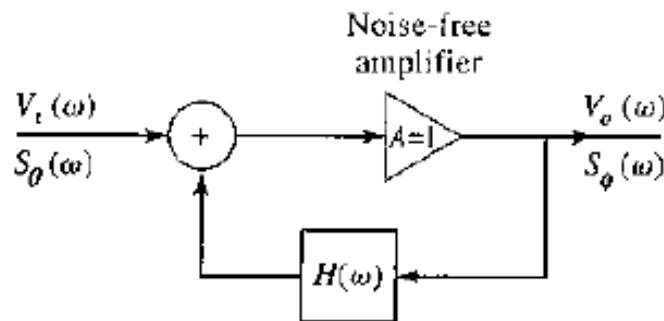


■ Leeson's Model

- For oscillators using high-Q resonant circuits in the feedback loop, the feedback transfer function can be represented as

$$H(\omega) = \frac{1}{1 + 2jQ\Delta\omega/\omega_0}$$

- Considering a noise-free amplifier with $A_v = 1$, the input and output spectral power densities are given by



$$\frac{S_\phi(\omega)}{S_\theta(\omega)} = \left(1 + \frac{\omega_0^2}{4Q^2\Delta\omega_0^2}\right)$$

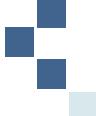
■ Leeson's Model

- Noise spectrum of a typical transistor amplifier is dominated by flicker noise ($1/f$) at low frequencies
- This flicker noise will modulate any applied signal (at f_0) due to amplifier nonlinearities, thus producing an input power spectral density for the oscillator given by

$$S_\theta(\omega) = \frac{kTF}{P_0} \left(1 + \frac{K\omega_\alpha}{\Delta\omega} \right)$$

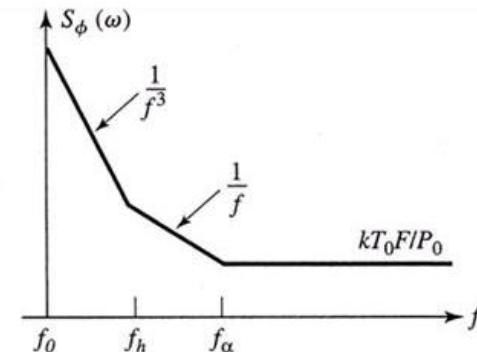
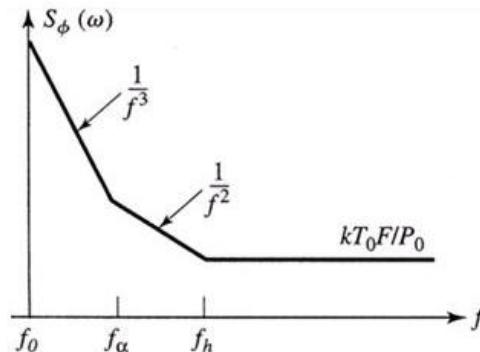
- Then, the power spectral density of the phase noise can be computed as

$$S_\phi(\omega) = \frac{kT_0F}{P_0} \left(\frac{K\omega_\alpha\omega_h^2}{\Delta\omega^3} + \frac{\omega_h^2}{\Delta\omega^2} + \frac{K\omega_\alpha}{\Delta\omega} + 1 \right)$$



■ Leeson's Model

- Noise power close to the carrier drops as $1/f^3$
- At frequencies between f_α and f_h
 - If the resonator has a low-Q (i.e. $f_h > f_\alpha$), the noise power drops as $1/f^2$
 - If the resonator has a high-Q (i.e. $f_\alpha > f_h$), the noise power drops as $1/f$
- At higher frequencies the noise is predominantly thermal
- In all cases, the noise close to the carrier is proportional to $1/Q^2$

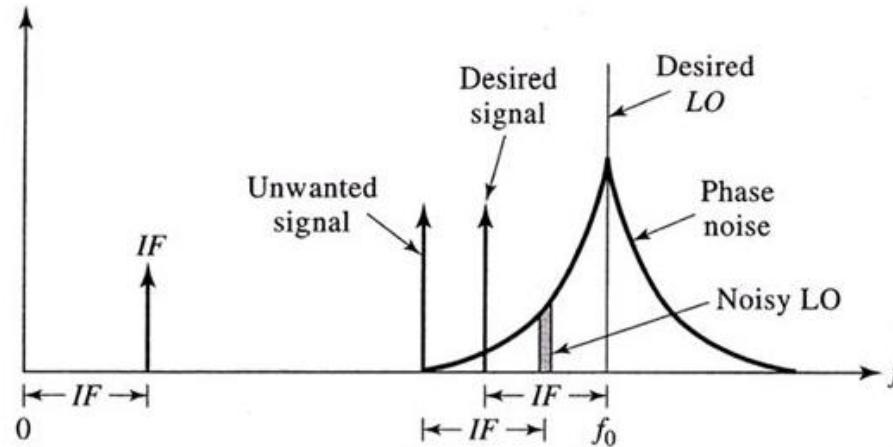




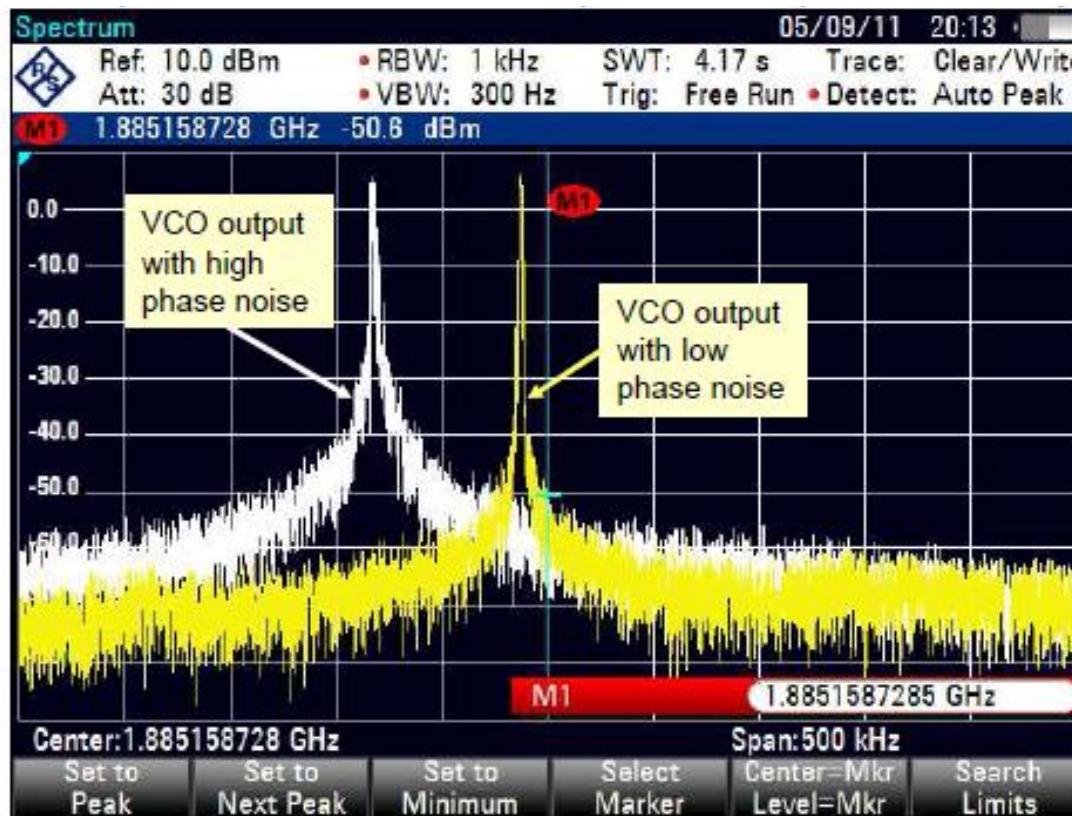
■ Leeson's Model

- Phase noise degrades receiver selectivity by causing down conversion of signals located nearby the desired frequency
- This problem is called reciprocal mixing
- The maximum allowable phase noise is in dBc/Hz at offset frequency f_m given by

$$P(f_m) = C(\text{dBm}) - S(\text{dB}) - I(\text{dBm}) - 10 \log(B)$$



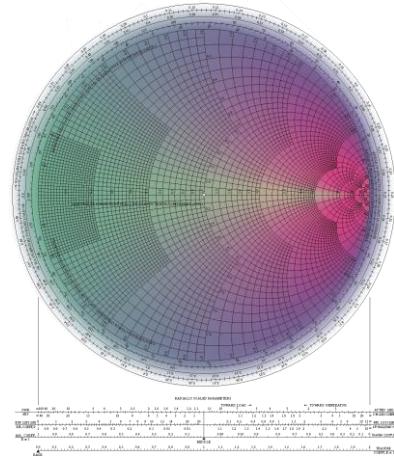
■ Phase Noise Example



High Frequency Electronic Circuits

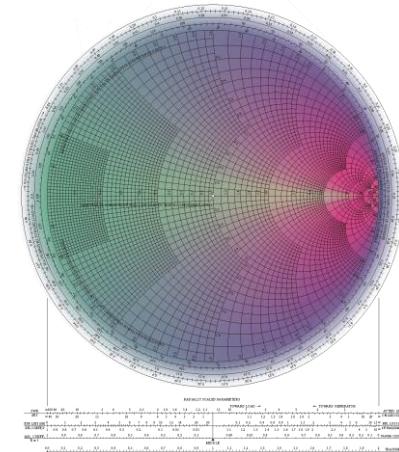
Noise & Low Noise Amplifiers (LNA)

Departamento de Ingeniería Electrónica
Universitat Politècnica de València (UPV)



Index

- Introduction
- Noise Types and Sources
- Noise in Electronic Components
- Generic Noise Equivalent Circuits
- System Level Noise Treatment
- Low Noise Amplifiers (LNAs)
 - Design specs
 - Subsystem Level issues
 - Device Level issues



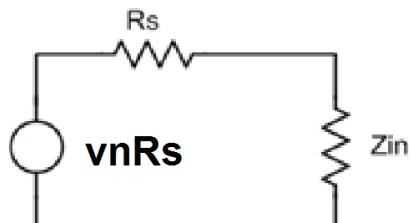
Introduction

- Introduction. Useful things to remember (I)

$$NF = \frac{SNR_{IN}}{SNR_{OUT}}$$

$$NF_{PASSIVE} = L_{(LOSSES)}$$

$$\text{FRIIS: } NF_T = NF_1 + \frac{NF_2 - 1}{A_1} + \dots + \frac{NF_m - 1}{A_1 A_2 \dots A_m}$$



Maximum Available Noise Power

Impedance Matched System

$$Zin = Rs$$



$$v_{onRs}^2 = 4kTRs \left(\frac{Zin}{Rs + Zin} \right)^2 = kTRs$$

$$P_{onRs} = \frac{kTRs}{Zin} = kT (-174 \text{ dBm/Hz}) \cdot B$$

Introduction

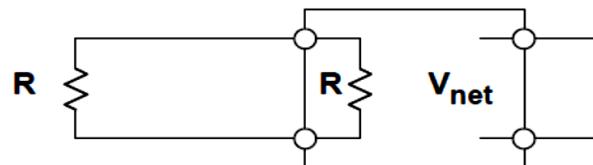
- Introduction. Useful things to remember (II)

Amplifier with source

Ideal impedance match conditions MTP

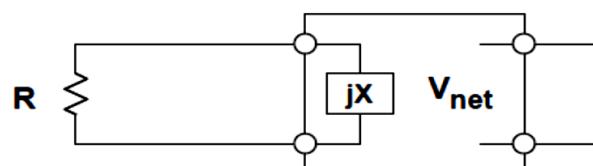
$$\text{NF} = \frac{P_{in}}{kT} \cdot \frac{kT \cdot G + P_{noAMP}}{P_{in} \cdot G} = 1 + \frac{P_{niAMP}^*}{kT} = 1 + \frac{T_{eqAMP}}{T_o}$$

$$\text{NF} = \frac{V_{n\text{OUT(total)}}^2}{A^2 kTR}$$



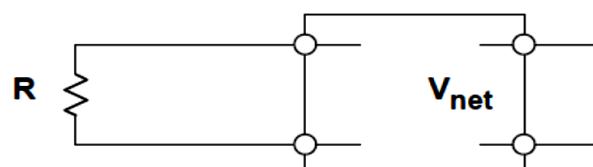
$$F = 2 + \frac{V_{net}^2}{A^2 kTR}$$

Matched Resistive
Termination



$$F = 1 + \frac{V_{net}^2}{A^2 kTR}$$

Matched Reactive
Termination
 $R = |X|$



$$F = 1 + \frac{V_{net}^2}{A^2 4kTR}$$

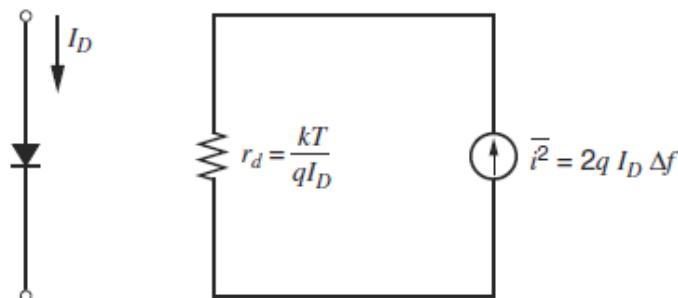
Unterminated

V_{net} = Voltage noise density of network excluding source and load terminations
 A = Open circuit voltage gain of network

Noise

- Noise Types. **SHOT NOISE**

- Related to PN junctions working under direct biasing (ON state).
- Charges require a certain energy to pass the junction voltage barrier.
- Since this is a statistical process (...) a certain variance is associated (NOISE)

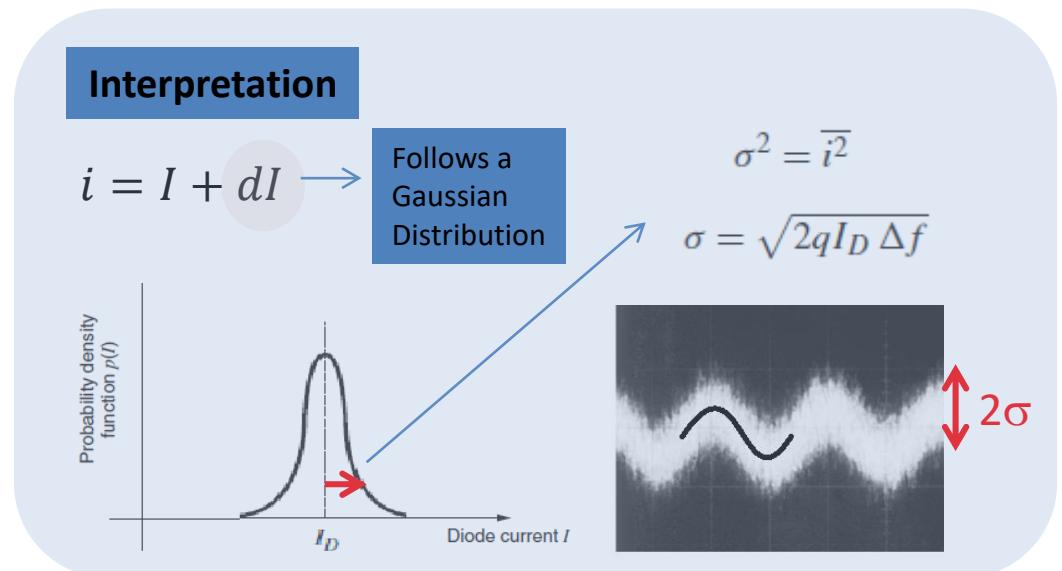


$\overline{i^2} = 2qI_D \cdot \Delta f$

Noise Current Spectral Density

White Noise

$\tau_{e,h} \ll (1/F_{\text{working}})$



Noise

- Noise Types. **THERMAL NOISE**

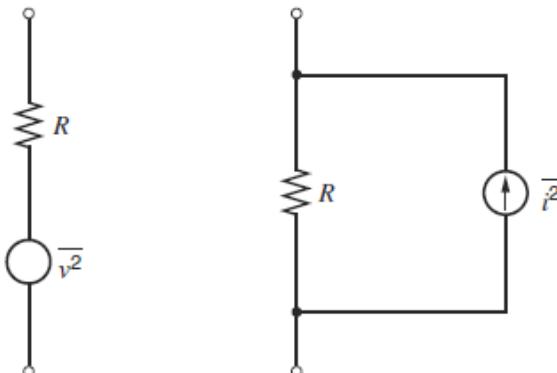
- Random charge movement inside a resistive element.
- Bias current independent.
- Temperature proportional.

**Voltage/ Current
Noise Spectral
Density**

$$\overline{v^2} = 4kTR \Delta f$$

$$\overline{i^2} = 4kT \frac{1}{R} \Delta f$$

White Noise



Useful Data for napkin calculations:

$$1\text{k}\Omega \rightarrow 4 \text{nV}/\sqrt{\text{Hz}}$$

Thermal Noise $1\text{k}\Omega \equiv$ Shot Noise 50uA

Noise

Other Noise types

- **Flicker Noise (1/f)**

- Active Devices and some passives (semicon. based)
- Charge Traps (Si crystal failures).
- Higher in Low Freq. (below MHz) though can have a strong effect if thermal noise is low.

Noise Current
Spectral Density

$$\overline{i^2} = K_1 \frac{I^a}{f^b} \Delta f$$

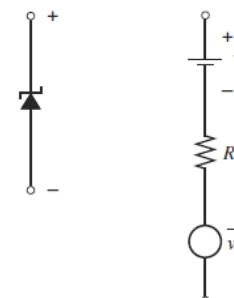
- **Popcorn Noise**

- Low Frequency.
- Comes from heavy (high Z) dopants (gold)

- **Avalanche (Dark Noise)**

- Associated to OFF state PN unions (zeners).
- BIAS circuitry critical.
- Constant p.s.d (white) but not Gaussian (Poisson)

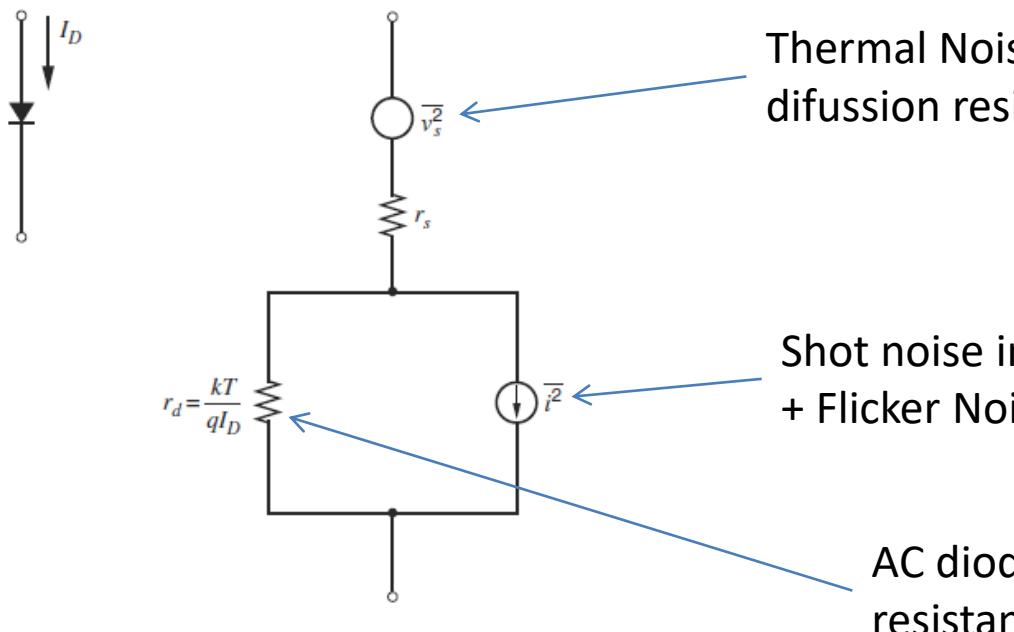
$$\overline{i^2} = K_2 \frac{I^c}{1 + \left(\frac{f}{f_c}\right)^2} \Delta f$$



100 nV/vHz

Noise

- Noise Equivalent Circuit. **DIODES**



Thermal Noise due to diffusion resistivity

$$\overline{v_s^2} = 4k T r_s \Delta f$$

Shot noise in PN union + Flicker Noise

$$\overline{i^2} = 2qI_D \Delta f + K \frac{I_D^a}{f} \Delta f$$

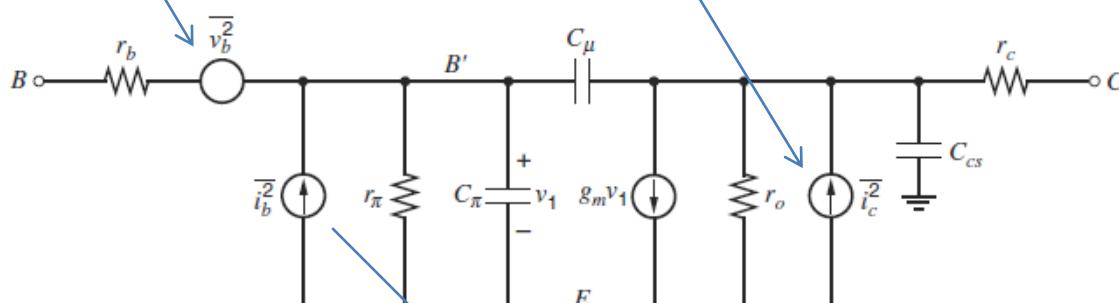
AC diode resistance

Noise

- Noise Equivalent Circuit. **Bipolar Transistor**

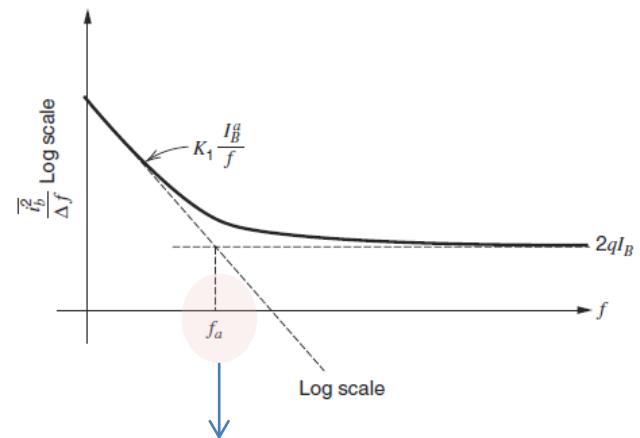
Base Resistor

$$\overline{v_b^2} = 4k T r_b \Delta f$$



Shot Noise in Base-
Collector PN Union

$$\overline{i_b^2} = \underbrace{2qI_B \Delta f}_{\text{Shot noise}} + \underbrace{K_1 \frac{I_B^a}{f} \Delta f}_{\text{Flicker noise}} + \underbrace{K_2 \frac{I_B^c}{1 + \left(\frac{f}{f_c}\right)^2} \Delta f}_{\text{Burst noise}}$$



Corner Frequency

Flicker noise in BJT is low
and f_a is low

Noise

- Noise Equivalent Circuit. **MOSFET**

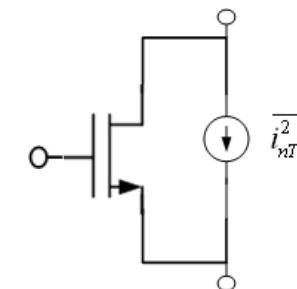
Thermal Noise

$$\overline{i_{nT}^2} = \frac{4kT}{R} \Delta f$$

MOSFET channel has a resistive nature

$$R \approx \frac{1}{n \cdot \Gamma} \cdot \frac{1}{g_m}$$

$n \rightarrow$ Slope Factor (1.3)



Thermal Noise in resistive element

$\Gamma \rightarrow$ Weak Inv.: 0.5
Strong Inv.: 0.67

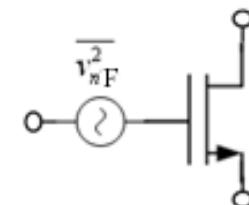
$$\overline{i_{nT}^2} = 4kT \cdot n \cdot \Gamma \cdot g_m \Delta f$$

Flicker Noise

Due to the charges interaction in the SiO_2 interphase

$$\overline{v_{nf}^2} = \frac{K}{C_{ox}WL} \frac{\Delta f}{f^{AF}}$$

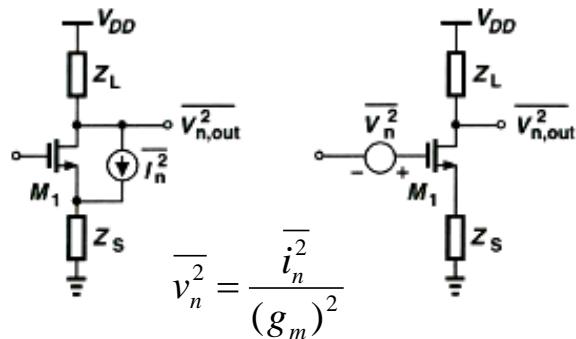
AF \rightarrow Freq. coefficient (~ 1)



Noise

- Noise Equivalent Circuit. **Equivalent Simplifications**

Gate related Total Noise (MOSFET)



$$v_{gs} = R_s(i_n - g_m v_{gs})$$

$$v_0 = -g_m v_{gs} R_D$$

$$v_0 = (g_m v_{gs} - i_n) R_D$$

$$v_n - g_m v_{gs} \cdot R_s = v_{gs}$$

$$v_0 = \frac{R_D}{1 + g_m R_s} i_n$$

$$v_0 = \frac{g_m R_D}{1 + g_m R_s} v_n$$

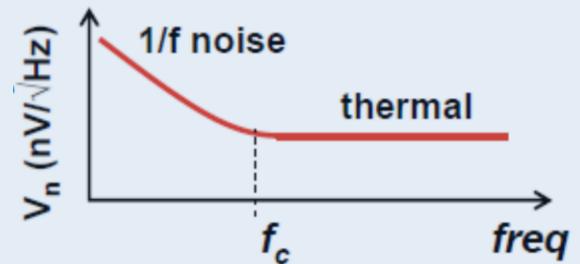
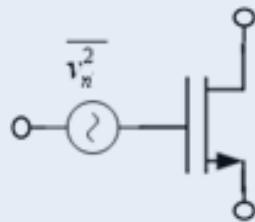
Both circuits are equivalent

THERMAL

$$\overline{i_{nT}^2} = 4kT.n.\Gamma.g_m\Delta f$$

$$\overline{v_{nT}^2} \approx \left(\frac{\overline{i_{nT}}}{g_m} \right)^2 = \frac{4kT.n\Gamma}{g_m} \Delta f$$

TOTAL referido a puerta

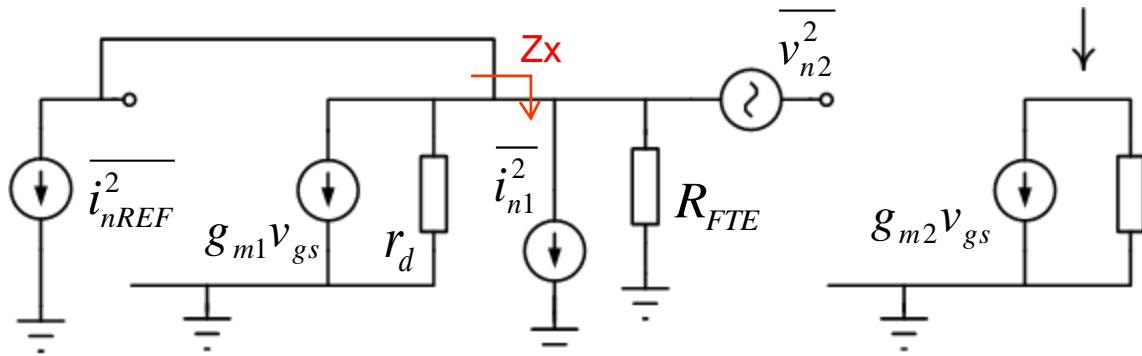
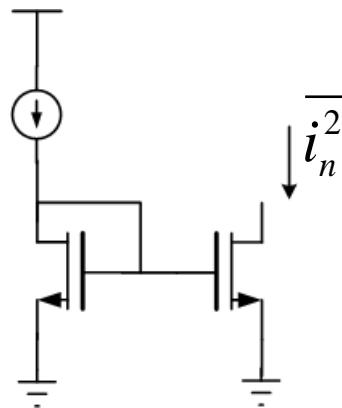


$$\overline{v_n^2} = \left(\frac{\overline{i_n}}{g_m} \right)^2 = \frac{4kT.n\Gamma}{g_m} \Delta f + \frac{K}{C_{ox}WL} \frac{\Delta f}{f^{AF}}$$

Noise

- Noise Calculations. Example 1**

- Non-correlated sources
- Reactances don't add noise
(only change its freq. shape)



- The goal is to obtain the transfer function from each noise source to the output.
- Remember we deal with quadratic magnitudes
- Each source will be analyzed disconnecting the other ones

$$Z_x \approx \frac{1}{g_{m1}}$$

$$\overline{i_{nREFo}^2} = \overline{i_{nREF}^2} \left(\frac{g_{m2}}{g_{m1}} \right)^2$$

$$\overline{i_{nEQo}^2} = \overline{i_{nEQo}^2} + \overline{i_{n1o}^2} + \overline{i_{n2o}^2}$$

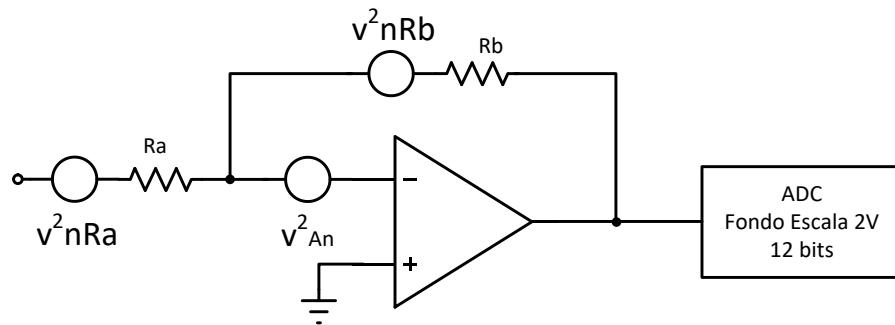
$$\overline{i_{n1o}^2} = \overline{i_{n1}^2} \left(\frac{g_{m2}}{g_{m1}} \right)^2$$

$$\overline{i_{nEQi}^2} = \frac{\overline{i_{nEQo}^2}}{\left(\frac{g_{m2}}{g_{m1}} \right)^2}$$

$$\overline{i_{n2o}^2} = \overline{v_{n2}^2} (g_{m2})^2$$

Noise

- Noise Calculations. Example 2**



What is the maximum equivalent noise level of the amplifier if a 12 effective bits resolution is required?

DATA: BW = 100MHz // $4kT=1,65E-20$ []
 $R_b=12\text{K}\Omega$ // $R_a=1.2\text{K}\Omega$

$$\overline{v_{nRa_OUT}^2} = \left(\frac{R_b}{R_a} \right)^2 \cdot \overline{v_{nRa}^2}$$

$$\overline{v_{nA_OUT}^2} = \left(1 + \frac{R_b}{R_a} \right)^2 \cdot \overline{v_{nA}^2}$$

$$\overline{v_{nRb_OUT}^2} = \overline{v_{nRb}^2}$$

$$\overline{v_{nOUT}^2} = \overline{v_{nRa_OUT}^2} + \overline{v_{nA_OUT}^2} + \overline{v_{nRb_OUT}^2}$$

$$\left(\frac{2}{2^{12}} \right)^2 \quad 0.2 \text{ uV}^2 \quad 0.020 \text{ uV}^2$$

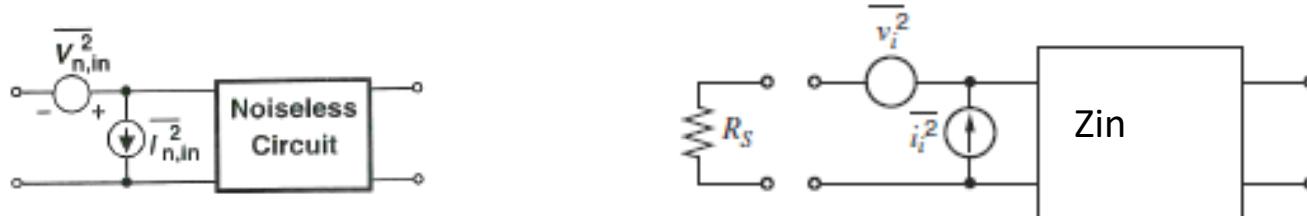
$$0.0184 \text{ uV}^2 (136 \text{ uV}_{\text{rms}})$$

$$\frac{0.0184 \text{ uV}^2}{11^2 \cdot 100 \text{ MHz}} = 1.23 \text{ } nV / \sqrt{\text{Hz}}$$

Noise

- Generic Noise Equivalents.

SOURCE IMPEDANCE EFFECT



$$NF = 1 + \frac{\left(\overline{v_n^2}\left(\frac{Z_{in}}{Z_{in} + R_s}\right)^2 + \overline{i_n^2}\left(\frac{Z_{in} \cdot R_s}{Z_{in} + R_s}\right)^2\right)}{4kTR_s \left(\frac{Z_{in}}{Z_{in} + R_s}\right)^2} = 1 + \frac{(\overline{v_n^2} + \overline{i_n^2}(R_s)^2)}{4kTR_s}$$

Valid for any impedance

$R_s \downarrow \downarrow$

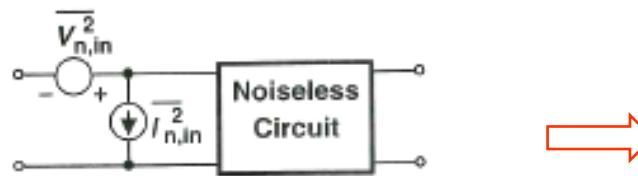
$$NF = \frac{\overline{v_{n OUT(total)}^2}}{A^2 4kTR_s}$$

Impedance Matching

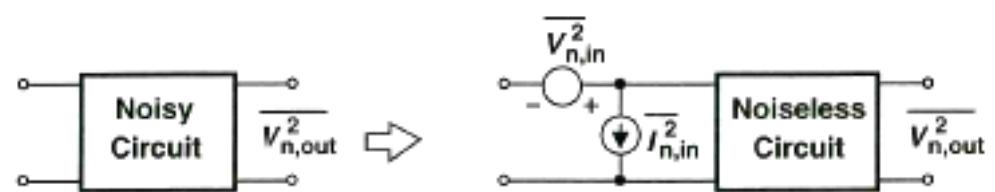
$$NF = \frac{\frac{4kTR_s}{4} + \left(\frac{\overline{v_n^2}}{4} + \frac{\overline{i_n^2}(R_s)^2}{4}\right)}{\frac{4kTR_s}{4}} = \frac{\overline{v_{n OUT(total)}^2}}{A^2 kTR_s}$$

Noise

- Generic Noise Equivalents.

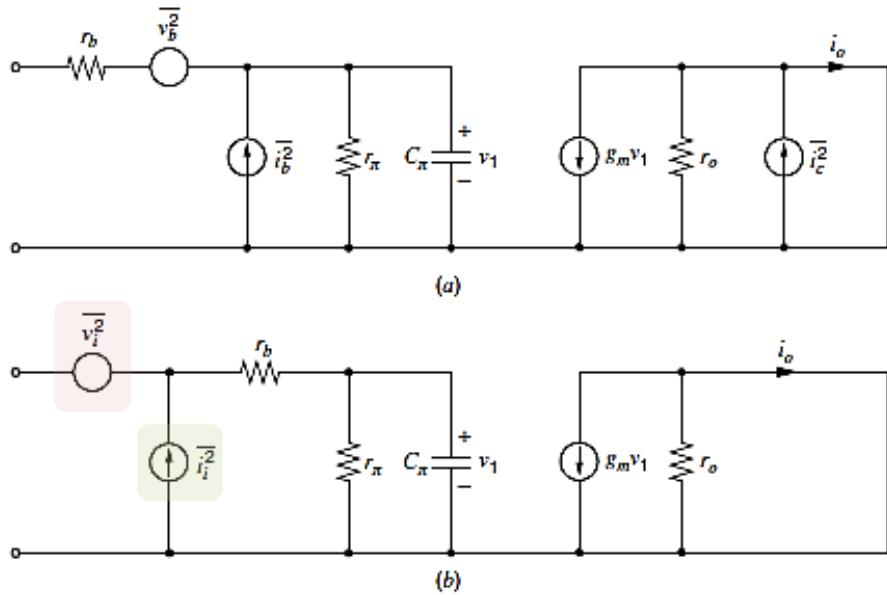


Generic Equivalent valid for
any source impedance



Noise

- Noise Equivalent. BJT



$$\beta(j\omega)i_i = i_c + \beta(j\omega)i_b$$

$$\overline{i_i^2} = \overline{i_b^2} + \frac{\overline{i_c^2}}{|\beta(j\omega)|^2}$$

$$\overline{i_i^2} = \frac{\overline{i_i^2}}{\Delta f} = 2q \left[I_B + K'_1 \frac{I_B^a}{f} + \frac{I_C}{|\beta(j\omega)|^2} \right]$$

Shot Noise - leq

$$r_b \ll r_\pi \quad g_m v_b + i_c = g_m v_i$$

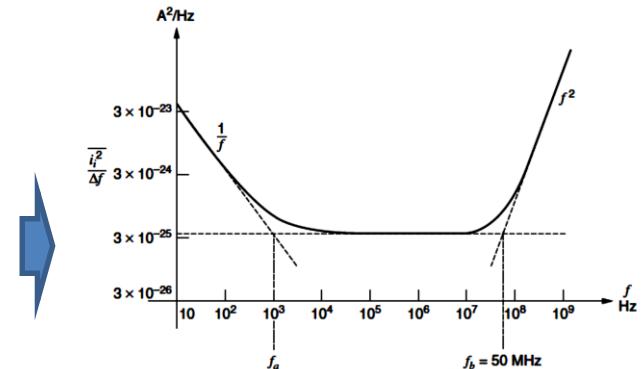
$$\overline{v_i^2} = \overline{v_b^2} + \frac{\overline{i_c^2}}{g_m^2}$$

$$\overline{v_i^2} = 4k T r_b \Delta f + \frac{2q I_C \Delta f}{g_m^2}$$

$$g_m = \frac{I_C}{kT/q}$$

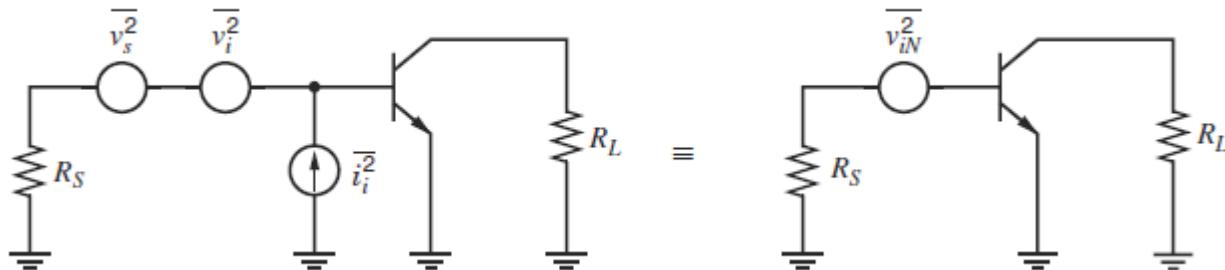
$$\frac{\overline{v_i^2}}{\Delta f} = 4k T \left(r_b + \frac{1}{2g_m} \right)$$

Noise Req



Noise

- Noise Equivalent. BJT (Calculations)



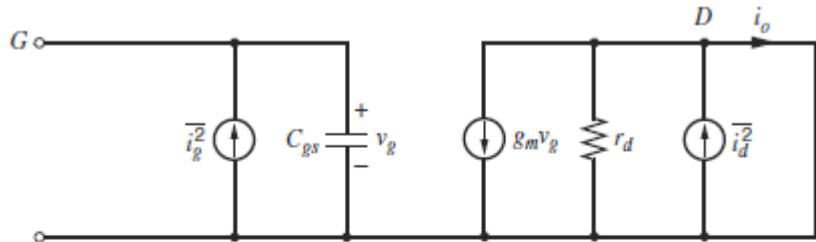
$$\frac{v_{iN} Zin}{Zin + R_S} = \frac{v_s Zin}{Zin + R_S} + \frac{v_i Zin}{Zin + R_S} + i_i \frac{R_S Zin}{Zin + R_S}$$

$$v_{iN} = v_s + v_i + i_i R_S$$

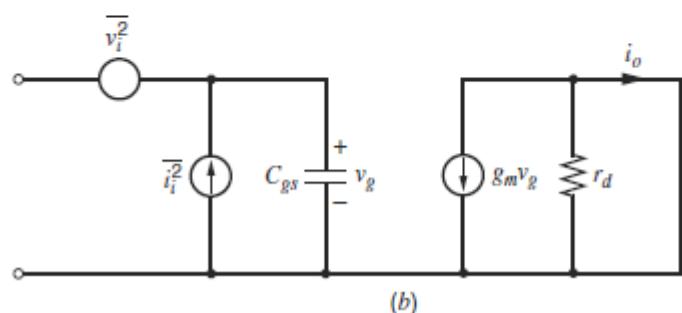
$$\overline{v_{iN}^2} = \overline{v_s^2} + \overline{v_i^2} + \overline{i_i^2} R_S^2 \quad \longrightarrow \quad \frac{\overline{v_{iN}^2}}{\Delta f} = 4k T R_S + 4k T \left(r_b + \frac{1}{2g_m} \right) + R_S^2 2q \left[I_B + \frac{I_C}{|\beta(jf)|^2} \right]$$

Noise

- Noise Equivalents. MOSFET



(a)



(b)

$$i_i \frac{g_m}{j\omega C_{gs}} = i_g \frac{g_m}{j\omega C_{gs}} + i_d$$



$$\overline{i_i^2} = \overline{i_g^2} + \frac{\omega^2 C_{gs}^2}{g_m^2} \overline{i_d^2} \quad \Rightarrow \quad \frac{\overline{i_i^2}}{\Delta f} = 2qI_G + \frac{\omega^2 C_{gs}^2}{g_m^2} \left(4kT \frac{2}{3} g_m + K \frac{I_D^a}{f} \right)$$

$$i_d = g_m v_i$$



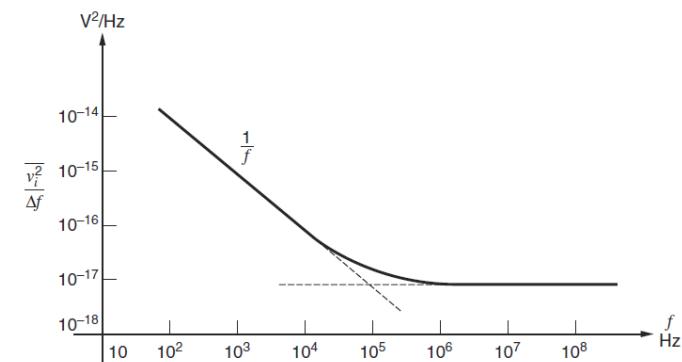
$$\overline{v_i^2} = \frac{\overline{i_d^2}}{g_m^2}$$



$$\frac{\overline{v_i^2}}{\Delta f} = 4kT \frac{2}{3} \frac{1}{g_m} + K \frac{I_D^a}{g_m^2 f}$$

$$\frac{\overline{v_i^2}}{\Delta f} = 4kT \left(\frac{1}{n\Gamma} \frac{1}{g_m} + K' \frac{I_D^a}{g_m^2 f} \right)$$

Noise Req



Noise

- Noise in transistors: **MOSFET vs BJT**

MOSFET

$$\frac{\overline{v_i^2}}{\Delta f} = 4k T \left(\frac{1}{n \Gamma} \frac{1}{g_m} + K' \frac{I_D^a}{g_m^2 f} \right)$$

>

BJT

$$\frac{\overline{v_i^2}}{\Delta f} = 4k T \left(r_b + \frac{1}{2g_m} \right) \quad g_{m\text{BJT}} > g_{m\text{MOSFET}}$$

0

$$\frac{\overline{i_i^2}}{\Delta f} = 2q I_G + \frac{\omega^2 C_{gs}^2}{g_m^2} \left(4k T \frac{2}{3} g_m + K \frac{I_D^a}{f} \right)$$

<

$$\frac{\overline{i_i^2}}{\Delta f} = 2q \left[I_B + K'_1 \frac{I_B^a}{f} + \frac{I_C}{|\beta(j\omega)|^2} \right]$$

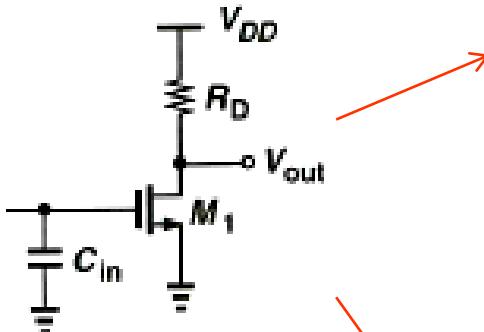
$$\beta(j\omega) = \frac{\beta_0}{1 + j \frac{\omega}{\omega_\beta}}$$

MOSFET is better with high source impedances (i^2)
BJT is better with lower source impedances (v^2) *

* This is a general conclusion. It might depend on specific design conditions

Noise

- Noise Equivalents. Amplifiers



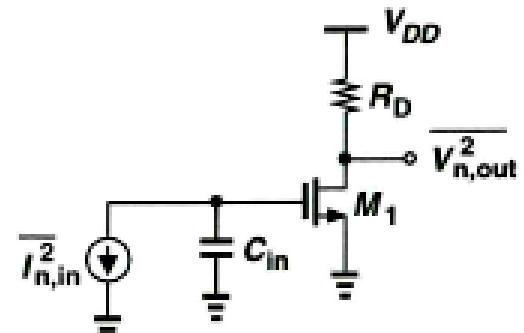
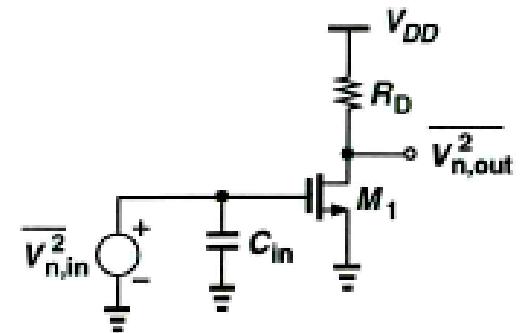
$$\overline{V_{n,out}^2} = \left(4kT \frac{2}{3} g_m + \frac{4kT}{R_D} \right) R_D^2$$

Equivalent Short-circuit
→ Equivalent noise voltage
at the input

$$\overline{V_{n,in}^2} = 4kT \frac{2}{3g_m} + \frac{4kT}{g_m^2 R_D}$$

Equivalent Open-circuit
→ Equivalent noise current
at the input

$$\overline{I_{n,in}^2} = (C_{in}\omega)^2 \frac{4kT}{g_m^2} \left(\frac{2}{3} g_m + \frac{1}{R_D} \right)$$

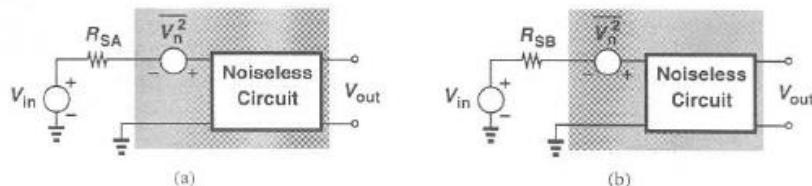


Noise

• System Level Noise Treatment

- In **homogeneous** systems, the NF and the Friis eq are enough for noise calculations. However in **heterogeneous** systems with different impedances...

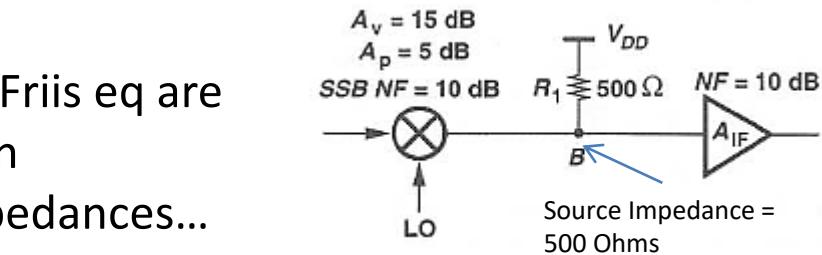
With negligible i^2_n components, translating NF(50Ω) to NF(500Ω):



$$NF_A = 1 + \frac{\overline{v_n^2}}{4kTR_{SA}}$$

$$NF_B = 1 + \frac{\overline{v_n^2}}{4kTR_{SB}}$$

$$(NF_A - 1)R_{SA} = (NF_B - 1)R_{SB}$$



With **non-negligible i^2_n** the noise equivalent is required:

$$(NF_A - 1)4kTR_{SA} - R_{SA}^2 \overline{I_n^2} = (NF_B - 1)4kTR_{SB} - R_{SB}^2 \overline{I_n^2},$$

(page 14)

NF ?

$$NF_{500} = NF_{mixer} + \frac{(NF_{Aif_50} - 1) 50 / 500}{(\overline{V_o^2} / 500) / (\overline{V_i^2} / 50)}$$

$$NF_{500} = NF_{mixer} + \frac{(NF_{Aif_50} - 1)}{A_{Vmixer}^2}$$

Noise

- System Level Noise Treatment

$$NF_E(dB) = NF_{6(500\Omega)}(dB) + L_5(dB) = 15 dB$$

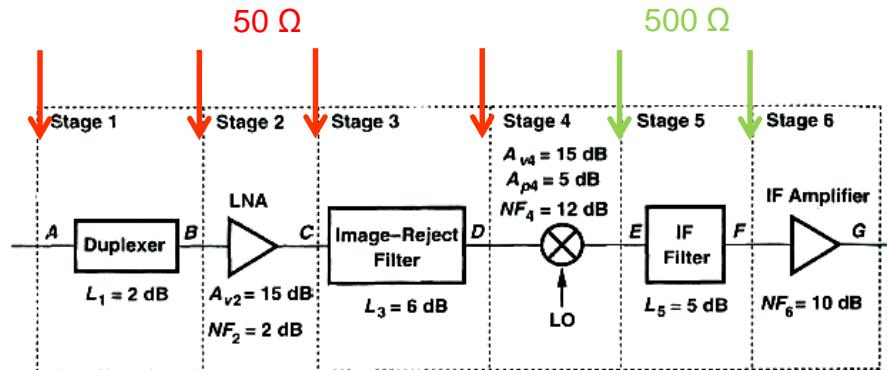
$$NF_D = NF_{4(500\Omega)} + \frac{NF_E - 1}{A_{P4}} = 25.53 \text{ (14.1 dB)}$$

$$NF_C(dB) = NF_D(dB) + L_3(dB) = 20.1 dB$$

$$NF_B = NF_{2(50\Omega)} + \frac{NF_C - 1}{(A_{V2})^2} = 4.78 \text{ (6.79 dB)}$$

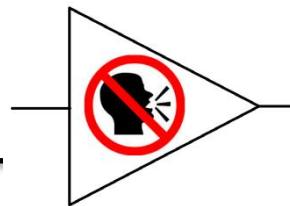
$$NF_A(dB) = NF_B(dB) + L_1(dB) = 8.79 dB$$

Regla: $R_{out1} \neq R_{NF2} \rightarrow$ Divide by A_V^2

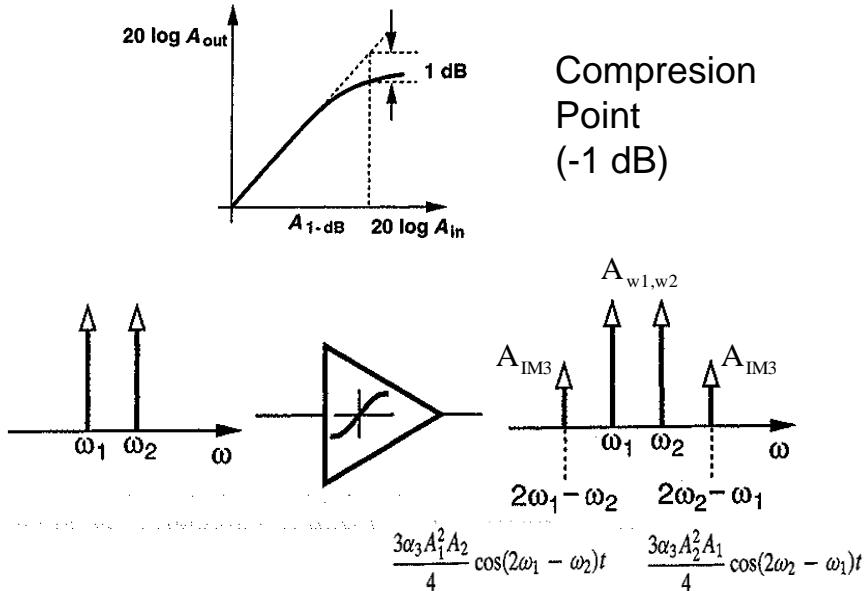


	A	Duplexer	B	LNA	C	Image-Reject Filter	D	Mixer	E	IF Filter	F	IF Amplifier
Stage Gain (dB)												
Voltage		-2		15			-6		15		-5	
Power		-2		15			-6		5			
Cumulative Voltage Gain (dB)		-2		13			7		22		17	
Stage NF (dB)		2		2			6		12		5	
Cumulative NF (dB)		8.79		6.79		20.1		14.1		15		10
Stage I_P3		+100 dBm		-12 dBm		+100 dBm		+5 dBm		1000 V _{rms}		700 mV _{rms}
Cumulative I_P3		-10.6 dBm		-12.6 dBm		+11 dBm		+5 dBm		22.1 V _{rms}		700 mV _{rms}

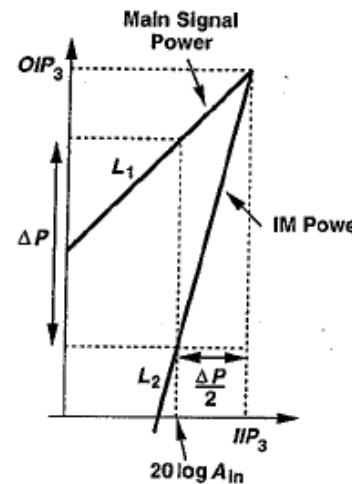
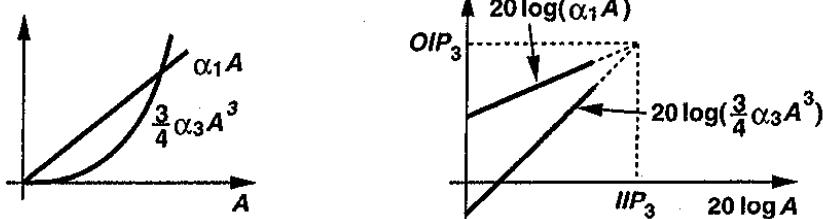
Distortion Measurements. SFDR



- Compression. IP3



A1 = A2



$$\alpha_1 A_{IP3} = \frac{3}{4} \alpha_3 A_{IP3}^3 \rightarrow A_{IP3} = \sqrt[3]{\frac{4 \alpha_1}{3 \alpha_3}}$$

$$\frac{A_{w1,w2}}{A_{IM3}} = \frac{\alpha_1 A_{in}}{(3/4) \alpha_3 A_{in}^3} = \frac{4}{3} \frac{\alpha_1}{\alpha_3} \frac{1}{A_{in}^2}$$

$$\frac{A_{w1,w2}}{A_{IM3}} = \frac{A_{IP3}^2}{A_{in}^2}$$

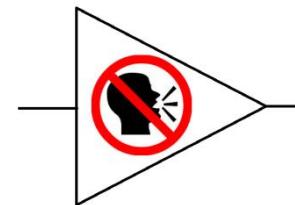
$$20 \log(A_{IP3}) = \frac{1}{2} (20 \log A_{w1,w2} - 20 \log A_{IM3}) + 20 \log A_{in}$$

$$P_{IIP3} = P_{IN} + \frac{P_{OUT} - P_{IM3out}}{2} = P_{IN} + \frac{P_{IN} - P_{IM3in}}{2} = \frac{3P_{IN} - P_{IM3in}}{2}$$

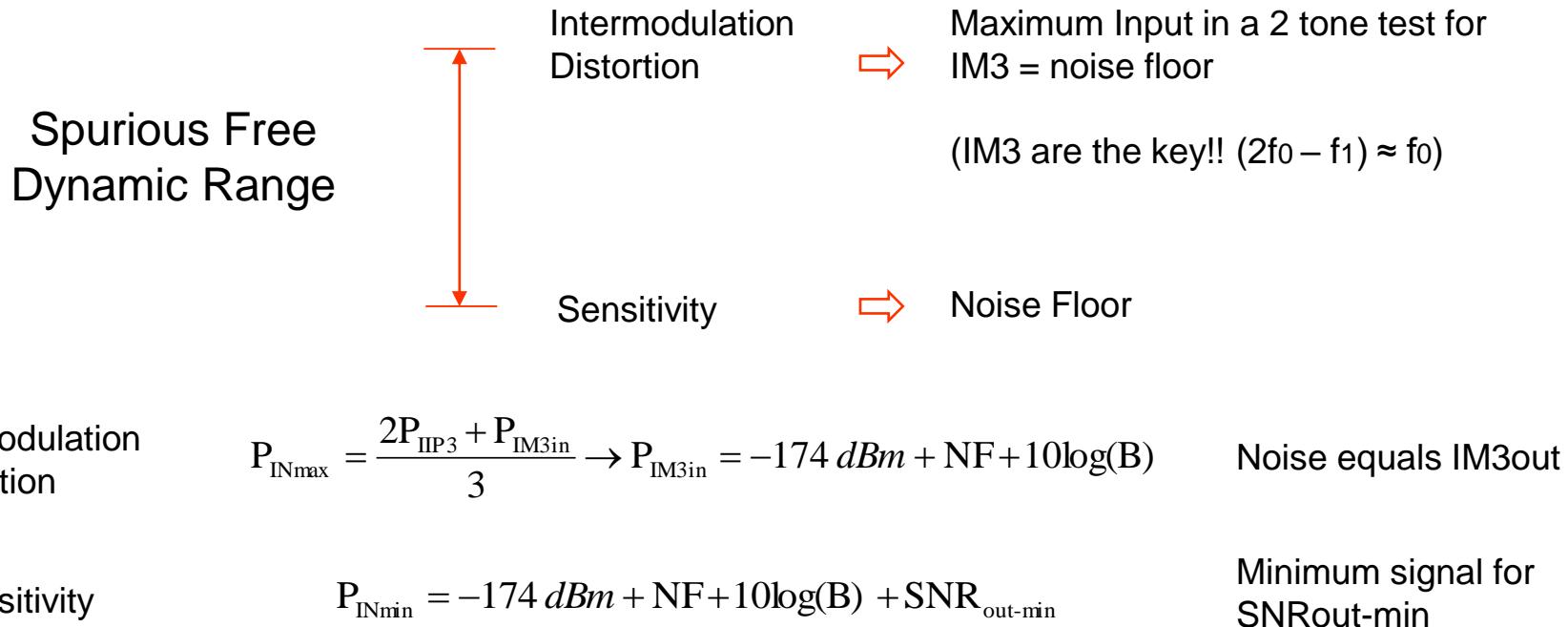
$$P_{IN} = \frac{2P_{IIP3} + P_{IM3in}}{3}$$

Input Equivalent Noise Power for IM3out (assuming a linear system)

Distortion Measurements. SFDR

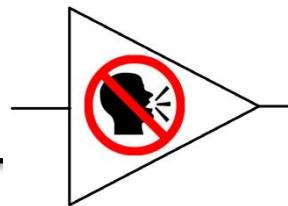


- Spurious Free Dynamic Range (SFDR)

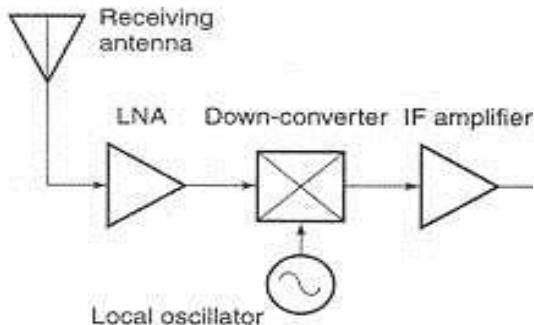


$$SFDR = \frac{2(P_{IIP3} + 174 \text{ dBm} - NF - 10\log(B))}{3} - SNR_{out-min}$$

Low Noise Amplifiers (LNA)

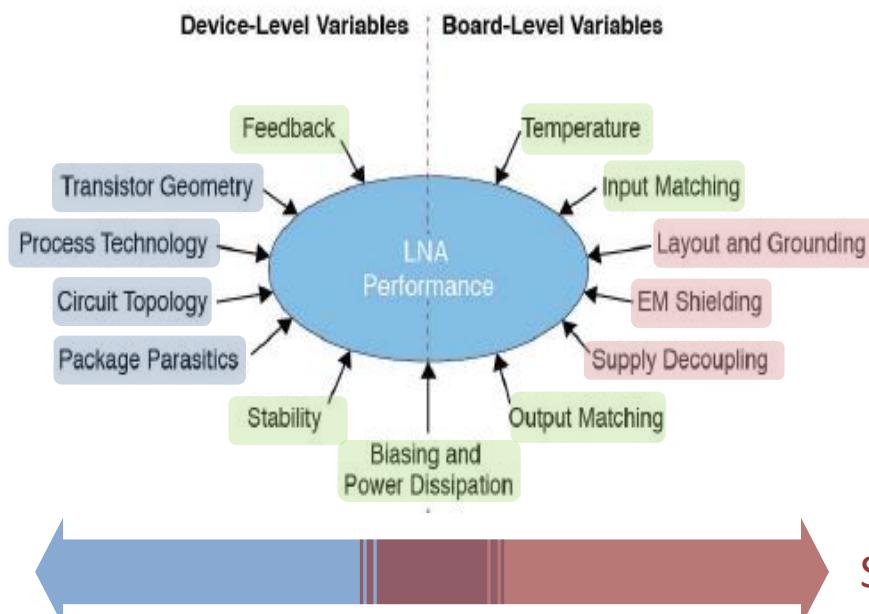


- Design Specifications for a LNA

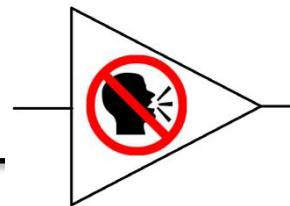


- **Noise Figure**
- **Gain**
- **Matching**
- **Bandwidth**
- **Linearity**
- **Dynamic Range**

Factors to be considered in a LNA design process

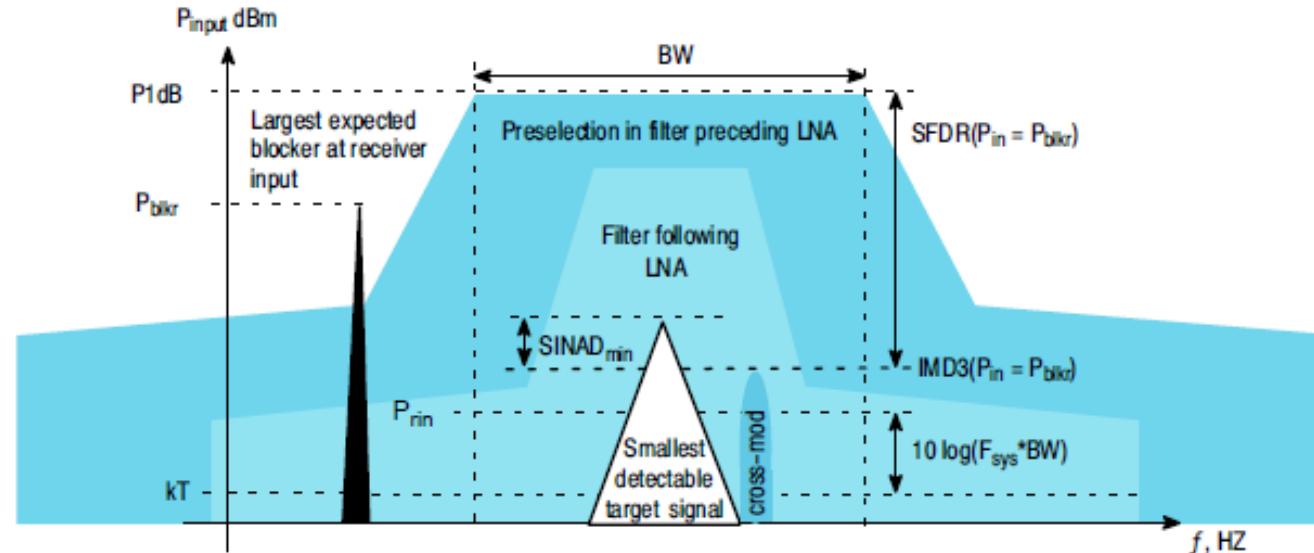


Low Noise Amplifiers (LNA)



- LNA Design Specifications

Receptor
Input



$$NF_{\text{sys}} = 10 \log(F_{\text{sys}}) = 10 \log\left(F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \dots + \frac{F_n - 1}{G_1 G_2 \dots G_{n-1}}\right), \text{ dB}$$

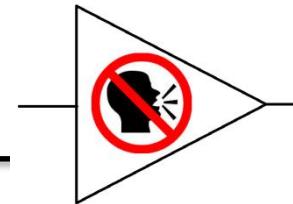
$$P_{\text{rin}} = kT + 10 \log(\text{BW}) + NF_{\text{sys}}, \text{ dBm}$$

$$[P1\text{dB}]_{\text{input}} = [P1\text{dB}]_{\text{output}} - G_{\text{sys}} + 1, \text{ dBm}$$

$$SFDR(P_{\text{in}} = P_{\text{blk}}) = [P1\text{dB}]_{\text{input}} - IMD3(P_{\text{in}} = P_{\text{blk}}), \text{ dB}$$

SFDR with the highest
interference signal in the
receptor

Low Noise Amplifiers (LNA)



- LNA Design Specifications

$$NF_{sys} = 10 \log(F_{sys}) = 10 \log\left(F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \dots + \frac{F_n - 1}{G_1 G_2 \dots G_{n-1}}\right), \text{dB}$$

$$P_{in} = kT + 10 \log(BW) + NF_{sys}, \text{dBm}$$

$$[P1dB]_{input} = [P1dB]_{output} - G_{sys} + 1, \text{dBm}$$

$$SFDR(P_{in} = P_{blk}) = [P1dB]_{input} - IMD3(P_{in} = P_{blk}), \text{dB}$$

Input Referred

$$\boxed{P_{IN} = \frac{2P_{IIP3} + P_{IM3in}}{3}}$$
$$P_{IM3in} = 3P_{IN} - 2P_{IIP3}$$

G₁ TRADE-OFF



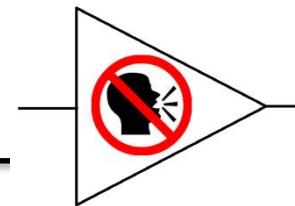
G₁ ↑↑ → NF ↓↓
→ P1dBinput ↓↓

LNA gain should be enough to control the noise chain without clipping the dynamic range

P_{in} (sensitivity ≡ noise floor) shows that for a given NF is better to reduce bandwidth to the lowest affordable for the application

In presence of interfering signals (desensitization), LNA linearity reduces IMD3 allowing a higher SFDR. **A 1 dB improvement in IIP3 translates into a 2 dB increase in the SFDR**

Low Noise Amplifiers (LNA)



- Design Constraints (**Subsystem Level**)

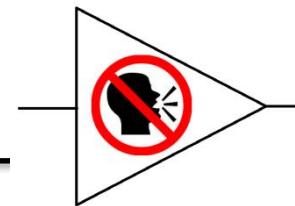
IMPEDANCE MATCHING

Noise (POV) → Impedance which minimizes NF (impedance translator?)
Power (POV) → Impedance Matching (MPT)

Usually a mixed approach strategy is used (impedance, noise and maximum available gain)

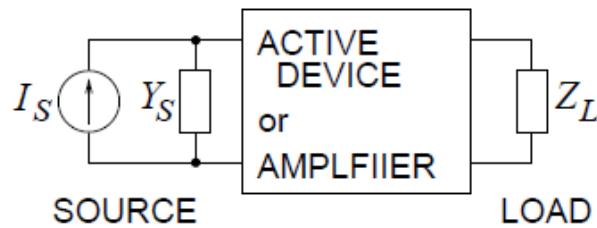
GOAL: Obtain an impedance matching with the lowest NF

Low Noise Amplifiers (LNA)



- Design Constraints (**Subsystem Level**)

IMPEDANCE MATCHING – NOISE - GAIN



In the traditional approach the admittance model and Y parameters are used

F_{\min} → Minimum Noise Factor

r_n → Noise Req normalized

y_{opt} → Admitancia obtenida para F_{\min}

g_s → real part of Y_s

$$F = F_{\min} + \frac{r_n}{g_s} |y_s - y_{opt}|^2$$
$$y_s = \frac{1 - \Gamma_s}{1 + \Gamma_s} \quad y_{opt} = \frac{1 - \Gamma_{opt}}{1 + \Gamma_{opt}}$$

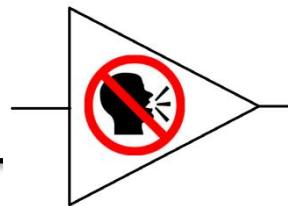
Usually Γ_{opt} doesn't provide impedance matching nor the highest gain

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

! f	NFmin	Gammaopt	rn/50
! GHz	dB	MAG ANG	-
1.900	0.60	0.15 39	0.15
2.400	0.63	0.15 46	0.14
3.500	0.73	0.13 57	0.13

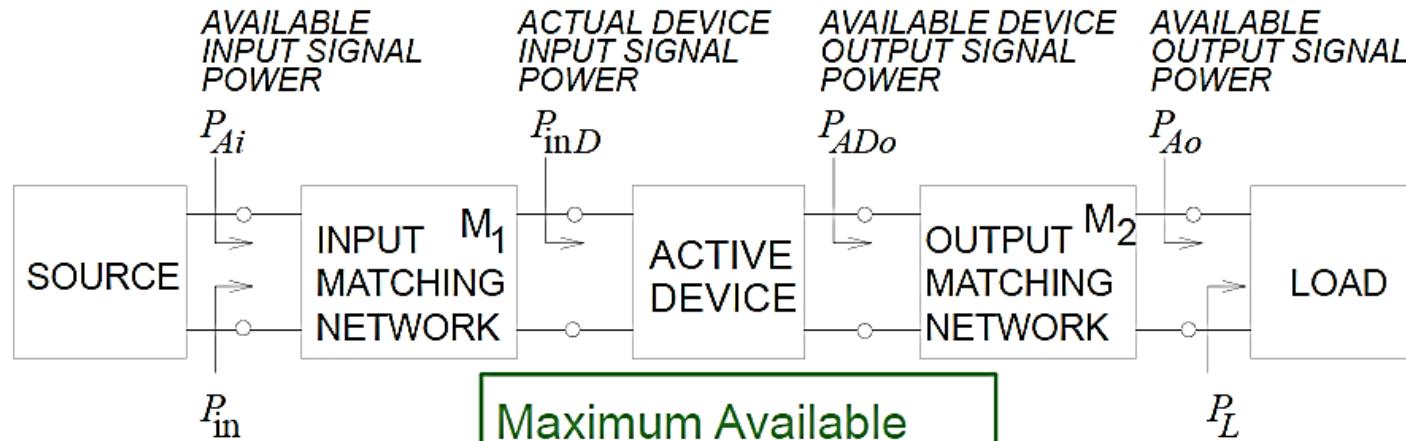
**Touchstone File
Noise Data**

Low Noise Amplifiers (LNA)



- Design Constraints (**Subsystem Level**)

IMPEDANCE MATCHING – NOISE - GAIN



Transducer Gain

$$G_T = \frac{P_L}{P_{Ai}}$$

Maximum Available Power Gain

$$G_{MA} = \left| \frac{S_{21}}{S_{12}} \right| \left(k - \sqrt{k^2 - 1} \right)$$

$$k = \left(\frac{1 - |S_{11}|^2 - |S_{22}|^2 + \Delta}{2|S_{12}||S_{21}|} \right)$$

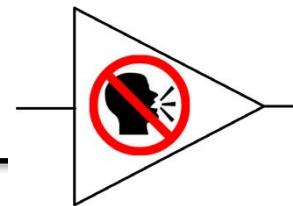
G_T with optimum M_1 and M_2

Maximum Stable Gain

$$G_{MS} = \left| \frac{S_{21}}{S_{12}} \right|$$

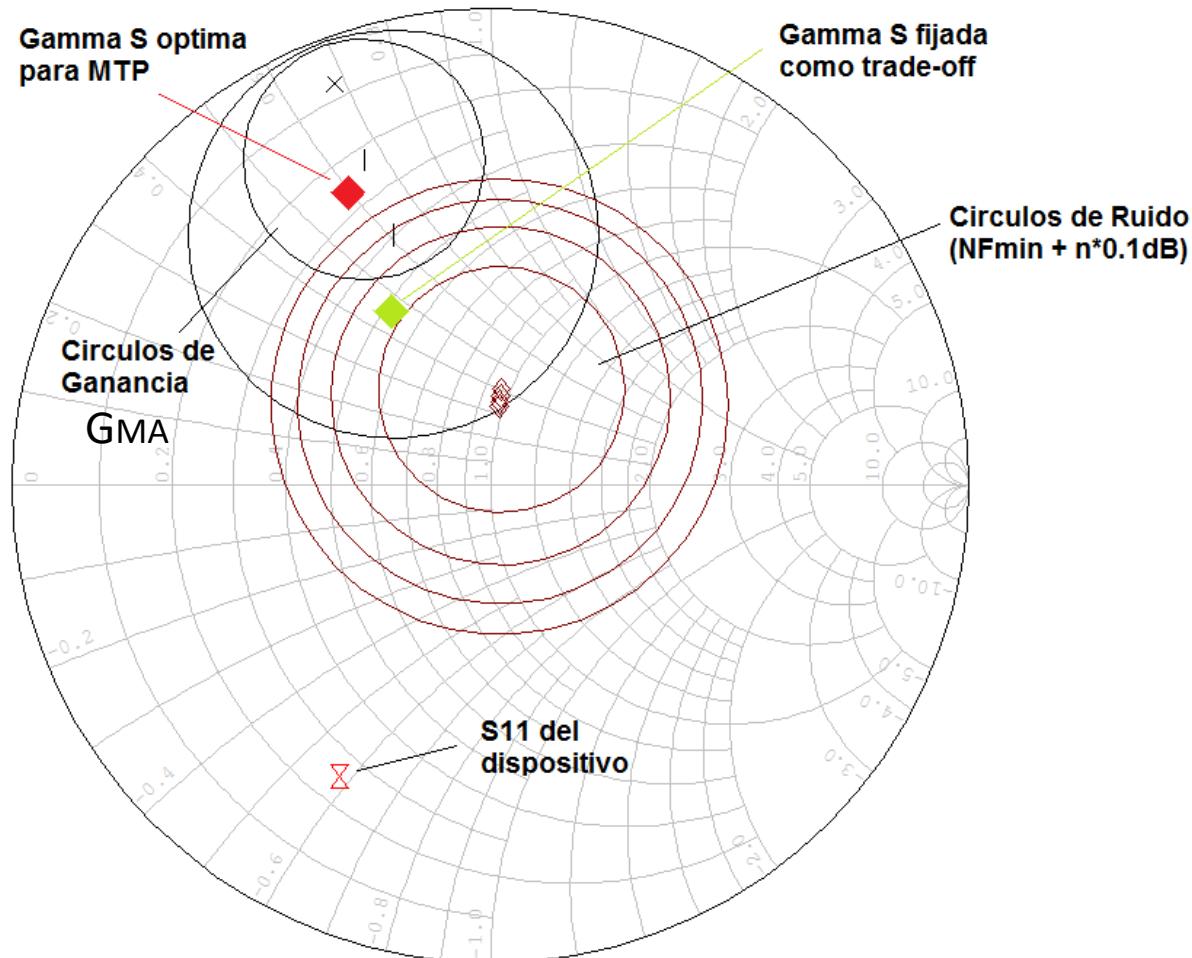
Like G_{MA} at edge of stability $k = 1$.

Low Noise Amplifiers (LNA)

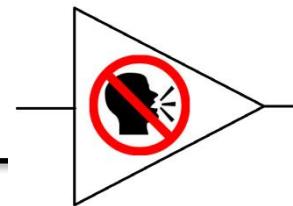


- Design Constraints (**Subsystem Level**)

IMPEDANCE MATCHING – NOISE – GAIN



Low Noise Amplifiers (LNA)



- Design Constraints (Subsystem Level)

STABILITY

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

$\rightarrow |\Gamma_{IN}| = \left| S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} \right| < 1$
 $\qquad\qquad\qquad |\Gamma_S| \leq 1 \quad |\Gamma_L| \leq 1$
 $\rightarrow |\Gamma_{OUT}| = \left| S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S} \right| < 1$

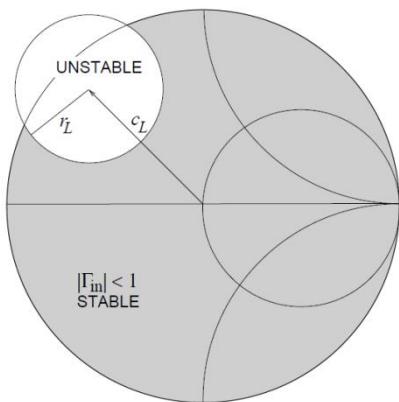
Oscillation limit for extreme cases ($\Gamma_L=1$ y $\Gamma_S=1$). STABILITY CIRCLES

center : $c_L = \frac{(S_{22} - \Delta S_{11}^*)^*}{|S_{22}|^2 - |\Delta|^2}$

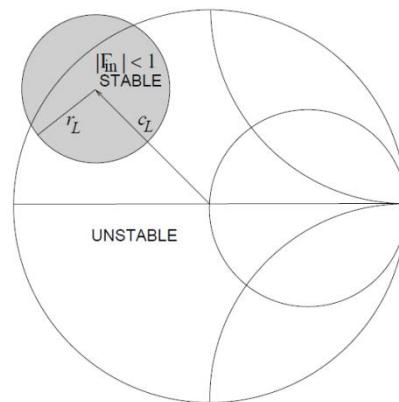
radius : $r_L = \left| \frac{S_{12}S_{21}}{|S_{22}|^2 - |\Delta|^2} \right|$

$|\Gamma_L - c| = r.$

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



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

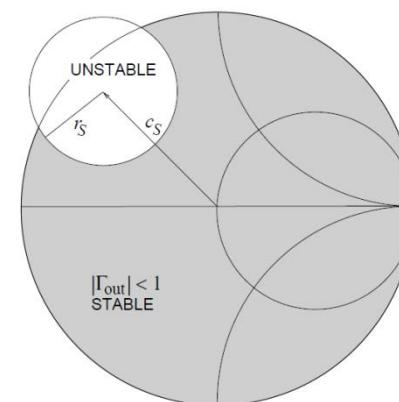


center : $c_S = \frac{(S_{11} - \Delta S_{22}^*)^*}{|S_{11}|^2 - |\Delta|^2}$

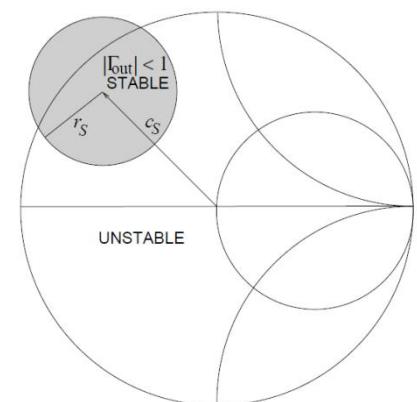
radius : $r_S = \left| \frac{S_{12}S_{21}}{|S_{11}|^2 - |\Delta|^2} \right|$

$|\Gamma_S - c| = r.$

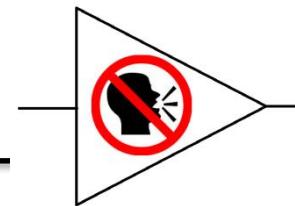
$$|S_{22}| < 1$$



$$|S_{22}| > 1$$



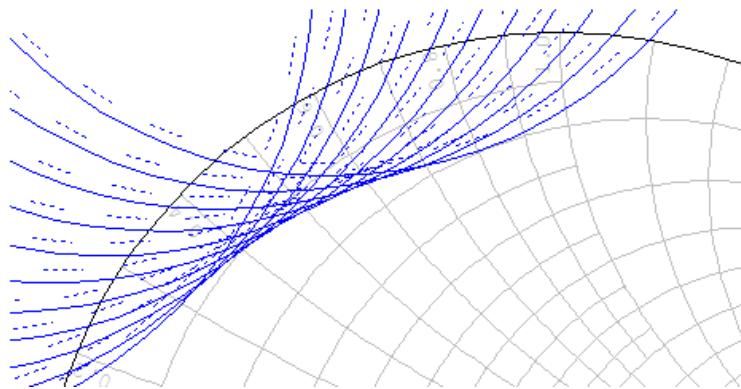
Low Noise Amplifiers (LNA)



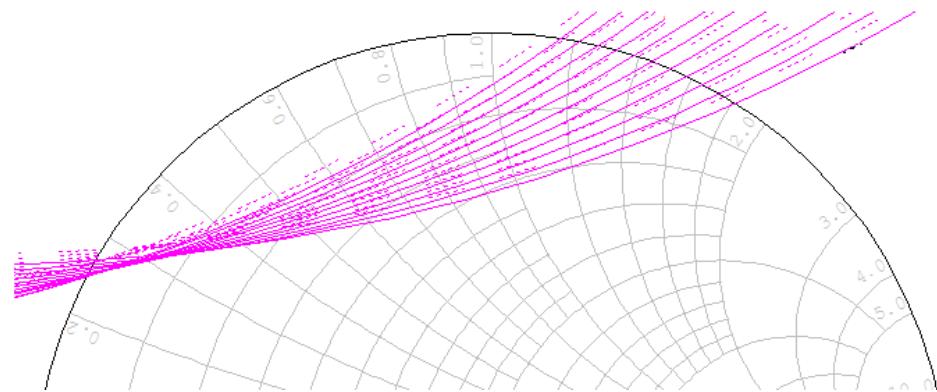
- Design Constraints (**Subsystem Level**)

STABILITY

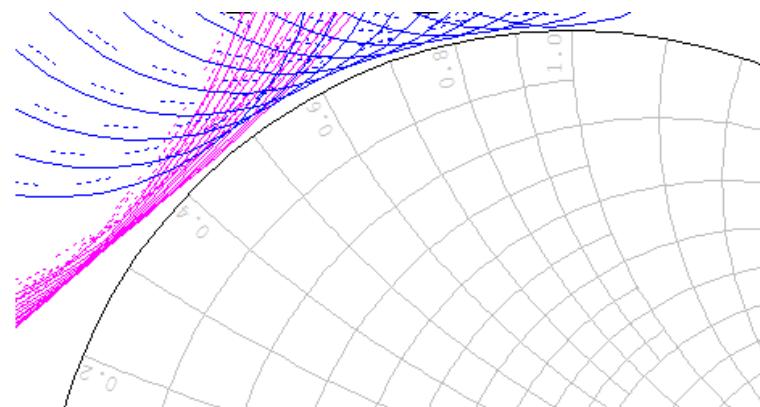
Input Circles $\rightarrow \Gamma_s$ stable zone



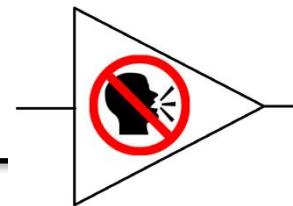
Output circles $\rightarrow \Gamma_L$ stable zone



But **BE CAREFUL**, the stability conditions must be fulfilled for **ALL** the frequencies (including those out of the band of interest)
Usually an absolutely stable design is preferred even if some gain is sacrificed (see Rollet (k) & Edwards-Sinsky (μ) criteria)

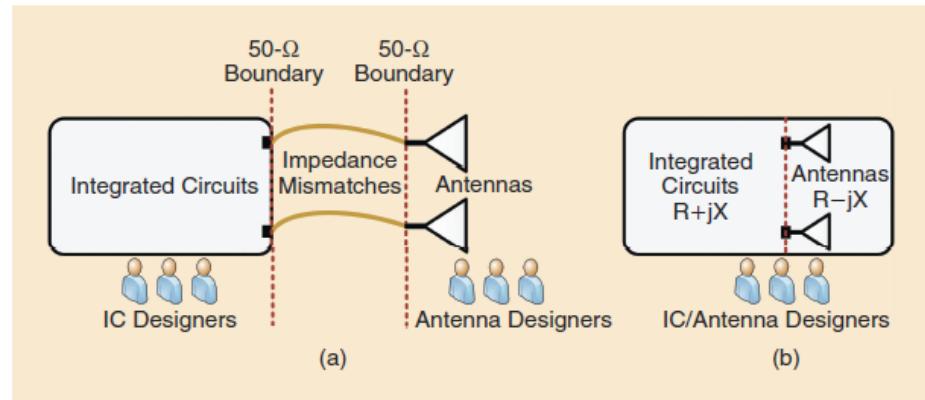
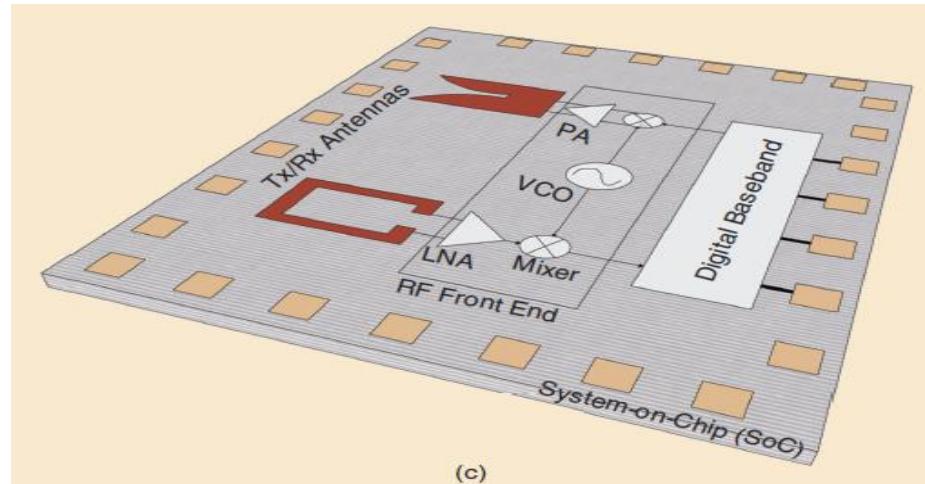
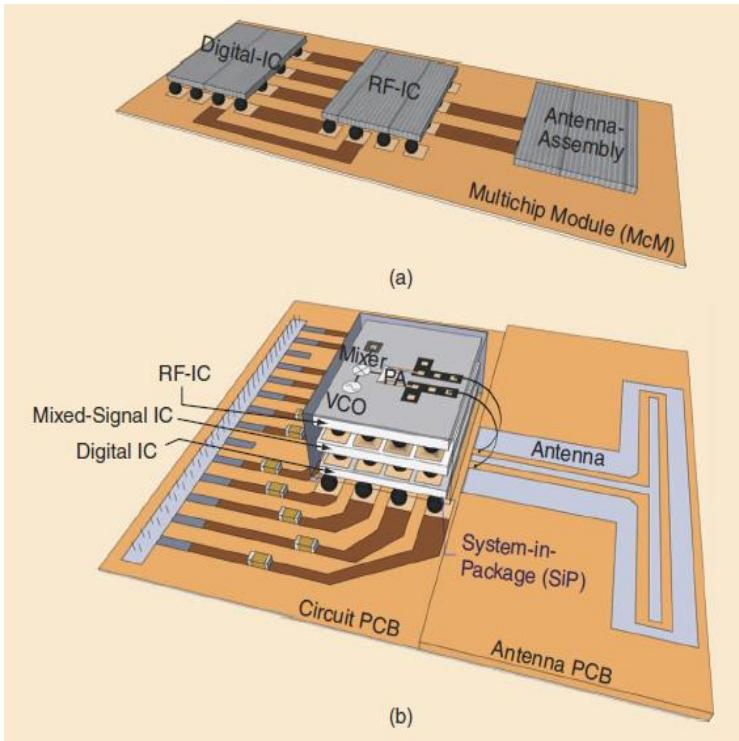


Low Noise Amplifiers (LNA)

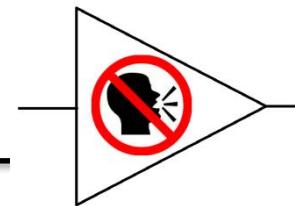


- Design Constraints (**Subsystem Level**)

IMPEDANCE MATCHING



Low Noise Amplifiers (LNA)

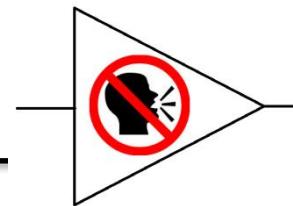


- Device Level Issues

IMPEDANCE MATCHING

- Antenna plus LNA Integration inside the same device allows a better optimization (LNA + Antenna optimization). **Ideal Case:** Antenna-On-Chip
- More design freedom. Impedance matching can be relaxed ($50\ \Omega$).
- Kirchhoff “friendly” up to many GigaHertz.
- Transistor level optimization (sizing etc) to enhance performance reducing external components.
- Feedback techniques can be employed in order to enhance stability and meet all the specs (see Inductive Source Degeneration)
- Parasitics are lower and better characterized. Sometimes can be even used as a part of the design (see bonding wire inductance).

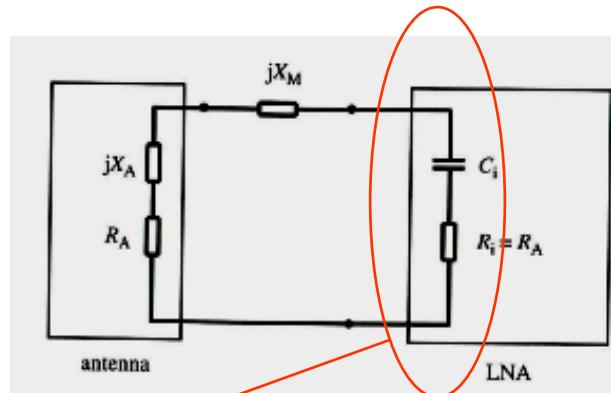
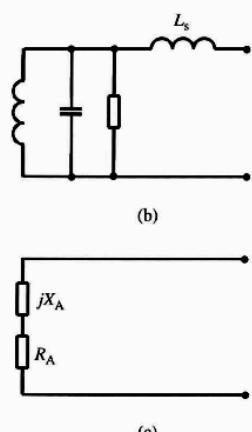
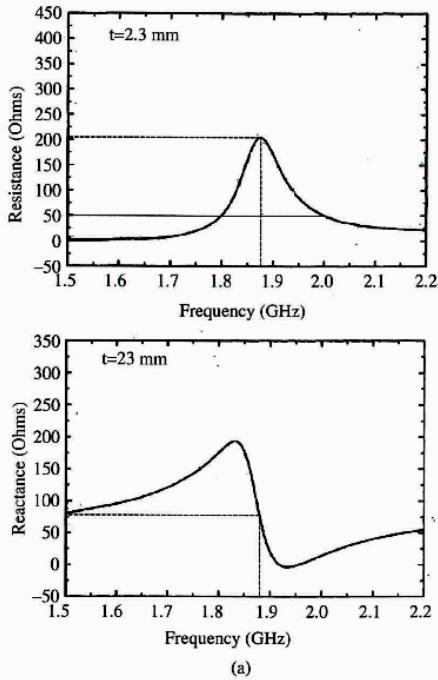
Low Noise Amplifiers (LNA)



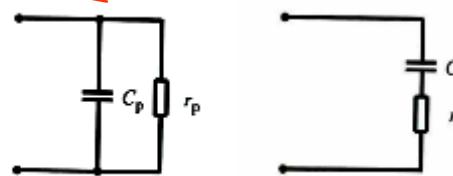
- Device Level Issues

IMPEDANCE MATCHING (Antennas).

F^{-1} Antenna Specs and matching



$$jX_A + jX_M + \frac{1}{j\omega_0 C_i} = 0$$

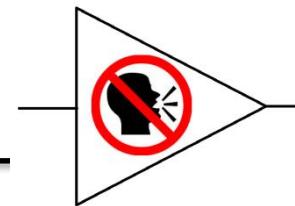


$$r_i = r_p \frac{1}{1 + (\omega \cdot C_p r_p)^2}$$

$$C_i = C_p \left(1 + \frac{1}{(\omega \cdot C_p r_p)^2} \right)$$

GOAL: Cancel reactance & match R_A / R_i

Low Noise Amplifiers (LNA)

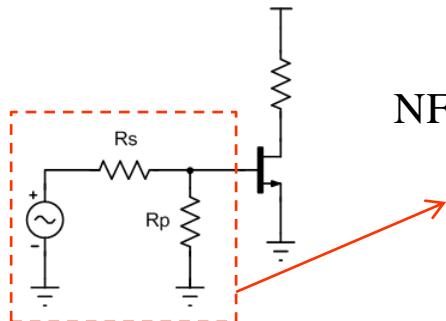


- Device Level Issues

Impedance Matching Design Options

TERMINAL RESISTOR

see page 4



$$NF = \frac{V_{nOUT}^2}{A^2 4kTR_S} = \frac{4kTR_S \left(\frac{R_P}{R_P + R_S} \right)^2 + 4kTR_P \left(\frac{R_S}{R_P + R_S} \right)^2}{\left(\frac{R_P}{R_P + R_S} \right)^2 4kTR_S} = 1 + \frac{R_S}{R_P}$$

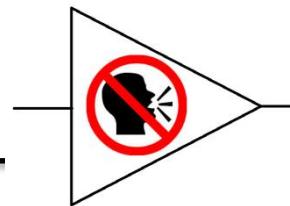
$$NF = 2 \text{ (3 dB)}$$

Resistor Divider only

FEEDBACK (voltage – current)

- ▼ Achieve stability conditions is VERY DIFFICULT in RF
- ▼ Circuit complexity increases NF

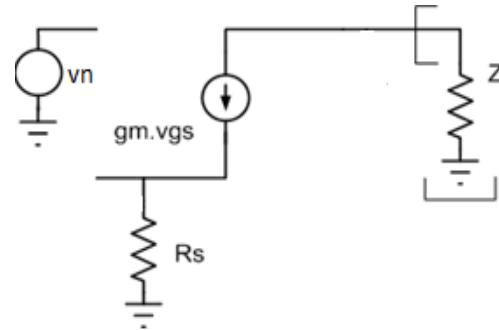
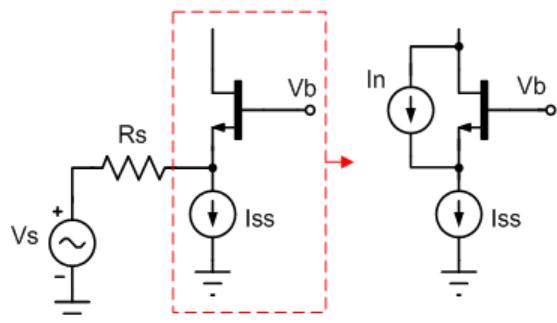
Low Noise Amplifiers (LNA)



- Device Level Issues

Impedance Matching Design Options

Common Gate Configuration



$$v_n^2 = \frac{4kT \cdot \gamma}{g_m}$$

$$\gamma \approx \frac{2}{3} \quad L \downarrow \rightarrow \gamma \uparrow$$

$$V_{nOUT_MOSFET} = g_m \cdot v_{gs} \cdot Z_L$$

$$\frac{V_{nOUT_MOSFET}}{v_n} = \frac{g_m Z_L}{1 + g_m \cdot R_s}$$

$$v_{gs} = v_n - g_m \cdot v_{gs} \cdot R_s$$

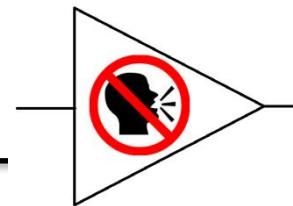
$$A = g_m Z_L$$

$$v_{gs} = \frac{v_n}{1 + g_m \cdot R_s}$$

$$\frac{V_{nOUT}^2}{A^2} = \left(\frac{1}{1 + g_m \cdot R_s} \right)^2 \frac{4kT \cdot \gamma}{g_m} + kT \cdot R_s \xrightarrow{\text{Adaptación_MTP}} kT \cdot R_s (1 + \gamma)$$

$$NF = \frac{V_{nOUT}^2}{A^2 kTR_s} = 1 + \gamma$$

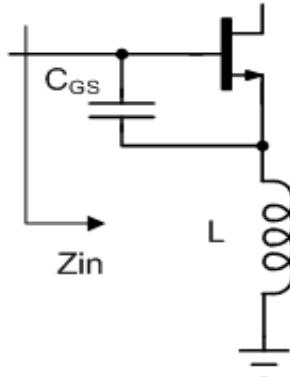
Low Noise Amplifiers (LNA)



- Device Level Issues

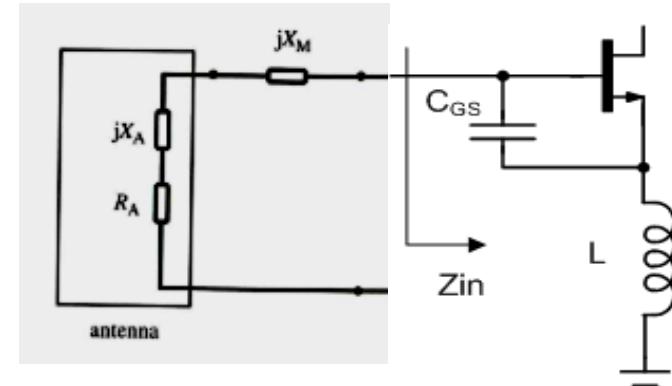
Impedance Matching Design Options

Inductive Source Degeneration



$$Z_{in} = \frac{1}{C_{GS} \cdot j\omega} + L \cdot j\omega + \frac{L}{C_{GS}} g_m$$

Cancel Reactance Match R

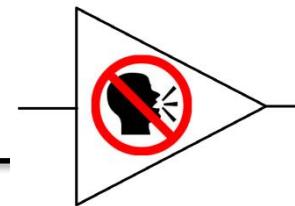


- Gain reduction
- Can be used for stabilization
- L is usually very low, same order or even lower than parasitics ground bond wire

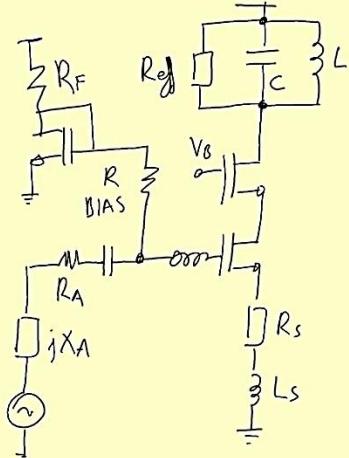
$$NF = 1 + \frac{\gamma}{g_m R_A}$$

Same as Common Source without matching (BEST??)

Low Noise Amplifiers (LNA)



Inductive Source Degeneration. EXAMPLE:



ANTENA F^{-1}

$$R_A = 210 \Omega \quad (1.8 + 64 \mu\text{F})$$

$$X_A = +75 \Omega$$

M1N - CAPACITOR
 $R_S \approx 0.1 \Omega$
 $(P_1 - P_2) - CAPACITOR$

$$R_B \approx 7.7 \Omega \quad (1 \mu\text{F}; 2.46 \mu\text{H})$$

$$Q = 54$$

CARAC. TÍPICAS

$$NF = 2 \text{ dB}$$

$$IIP_3 = -70 \text{ dBm}$$

$$G = 15 \text{ dB}$$

$$\text{RETURN LOSS} = -15 \text{ dB}$$

$$\text{ISOLATION} = 20 \text{ dB}$$

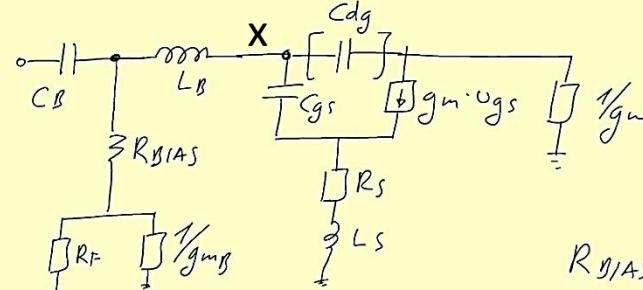
ESTABILIDAD > 1

$$C_P = \frac{1}{R_P + \frac{1}{j\omega C_S}}$$

$$C_S = C_P \left(1 + \frac{1}{(\omega C_B \cdot R_A)^2} \right)$$

$$R_S = \frac{R_P}{1 + (\omega C_P \cdot R_P)^2}$$

ADAPTACIÓN DE LA ENTRADA



$R_{BIAS} \text{ PP}$
 $[i_i^2 \text{ ff};$

AISLA EL CIRC. DE
 $BIAZ]$

$$Z_{in} \Rightarrow V_{gs} = i_i \frac{1}{C_{gs} \cdot S}$$

$$V_X = i_i \frac{1}{C_{gs} \cdot S} + \left(i_i + g_m \frac{i_i}{C_{gs} \cdot S} \right) \cdot (R_S + L_S \cdot S)$$

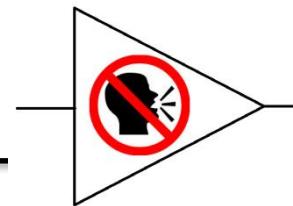
$$Z_{in} = \frac{1}{C_B \cdot S} + L_B \cdot S + \left[\frac{1}{C_{gs} \cdot S} + (R_S + L_S \cdot S) \cdot \left(1 + \frac{g_m}{C_{gs} \cdot S} \right) \right]$$

$$Z_{in} = R_A - j X_A \quad \leftarrow \text{ADAPTACIÓN}$$

$$\left[\frac{g_m L_S}{C_{gs}} + R_S \right] = R_A \{ Z_{in} \} = R_A$$

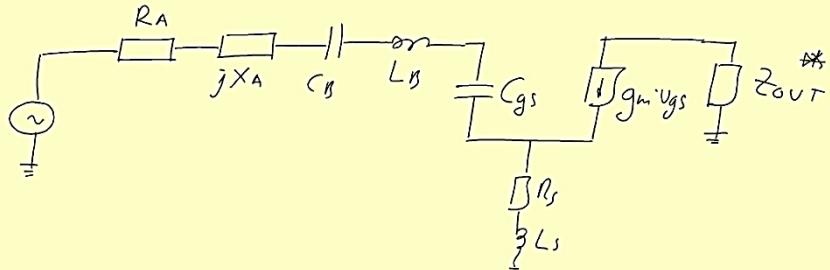
$$\frac{1}{C_B \cdot S} + \frac{1}{C_{gs} \cdot S} + \frac{g_m \cdot R_S}{C_{gs} \cdot S} + (L_B + L_S) \cdot S = I_m \{ Z_{in} \} = -j X_A$$

Low Noise Amplifiers (LNA)



Inductive Source Degeneration. EXAMPLE:

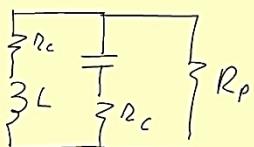
SUPONIENDO ADAPTACIÓN → GANANCIA



$$Z_{out}^* = Z_{TANK}$$

$$\text{f RESONANCIA (1,876 Hz)} \Rightarrow \frac{\frac{L_s}{C_s}}{L_s + \frac{1}{C_s}} = \infty \rightarrow \omega_0 = \frac{1}{\sqrt{LC}}$$

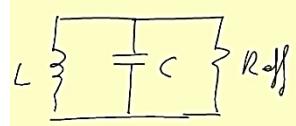
$$(Q_{eff,TANK})^{-1} = \frac{1}{Q_p} + \frac{1}{Q_L} + \frac{1}{Q_c} \Rightarrow Q_{eff,TANK} = Q_p // Q_L // Q_c$$



$$Q_p = \omega_0 C R_p = \frac{R_p}{\omega_0 L}$$

$$Q_L = \frac{\omega_0 L}{2L} \quad Q_c = \frac{1}{\omega_0 C \cdot 2C}$$

$$Q_{eff,TANK} = \frac{R_{eff}}{\omega_0 L} = \omega_0 C \cdot R_{eff}$$



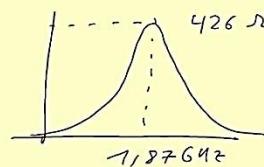
$$R_{eff} = R_p // R_L // R_C \xrightarrow{\text{Parallel}} R_{eff} \\ 2L \text{ en Paralelo}$$

CONSIDERAMOS TANQUE SIN R_p

$$\hookrightarrow Q_{eff} = Q_L // Q_c$$

$$Q_L = 10 \quad (\text{DATASHEET}) \quad Q_c = 436 \quad (\text{DATASHEET})$$

$$R_{eff} = \frac{Q_{eff}}{\omega_0 C} = \frac{9,77}{2\pi \cdot 1,87 \cdot 10^9 \cdot 1,95 \mu F} = 426 \parallel$$



(SALIDA CON $Z = 426 \Omega$)

VOLVIENDO A LA GANANCIA (SIMP. DEL CASCO DO)

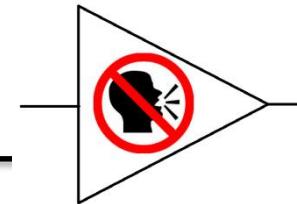
$$i_A = \frac{V_A}{Z_{in}^*} \rightarrow i_A \cdot \frac{1}{C_{gs} S} \cdot g_{m1} = i_d$$

$$\frac{i_d}{V_A} = \frac{g_{m1}}{(Z_{in}^*) \cdot C_{gs} \cdot S} = \frac{g_{m1}}{2R_A \cdot (C_{gs} \cdot S)}$$

$$\text{GANANCIA TOTAL} \sim \frac{g_{m1}}{2R_A \cdot C_{gs} \cdot S} \cdot (Z_{out}^*)$$

ES CASCO DO, POR LO QUE
ES APROXIMADA.

Low Noise Amplifiers (LNA)



Inductive Source Degeneration. EXAMPLE:

$$R_A = 2 \cdot 10 \rightarrow \frac{g_m}{C_{gs}} = \frac{R_A}{L_s} \quad / \quad C_{gs} = W L \frac{2}{3} C_{ox} + W C_{gs0}$$

$$C_{ox} = 4,54 \text{ fF}/\mu\text{m}^2$$

$$C_{gs0N} = 0,12 \text{ fF}/\mu\text{m}$$

$$C_{gs0P} = 0,09 \text{ fF}/\mu\text{m}$$

OPTIMIZACIÓN EN FREC.

$$L = 0,35 \mu\text{m}$$

(VER TEORÍA)

OPTIMIZACIÓN EN CONSUMO ?

$$\frac{I_o}{W/L} = 3,2 \rightarrow g_m = 3,3 \cdot 10^{-5} \cdot \frac{W}{L} = 9,4 \cdot 10^{-5} \text{ A}$$

$$C_{gs} = 1,18 \cdot 10^{-15} \text{ W} \quad ; \quad \frac{g_m}{C_{gs}} = \frac{R_A}{L_s} = (K) \text{ CTE}$$

$$L_s = 2,8 \cdot 10^{-9}$$

GANANCIA

$$A_V = \frac{g_m}{C_{gs} \cdot S} \cdot \frac{1}{2 R_A} \xrightarrow{\text{ADAP.}} A_V = \frac{Z_T}{2 L_s \cdot S} \approx 10$$

$$L_s \approx 1 \text{ nH}$$

No se puede optimizar consumo y ganancia.

$$L_s = 1,5 \text{ nH} (Q=9,6)$$

+ L bonding.

Fijamos la ganancia como parámetro

COMPROMISOS ADAPTACIÓN

$$\frac{g_m}{C_{gs}} = \frac{R_A}{L_s} = 138,7 \cdot 10^9$$

FIJAROS C_{gs} QUE PERMITA ADAPTACIÓN

$$\frac{1}{C_B \cdot S} + \frac{1}{C_{gs} \cdot S} + \frac{g_m}{C_{gs}} \cdot \frac{R_s}{S} + (L_B + L_s) \cdot S = -75$$

\uparrow \uparrow \uparrow \downarrow
 $C_B = 2 \text{ pF}$ (-40Ω) $b R_s / L_s$ 1 nH
 $(11,74)$
 -23Ω
 13 nH MAX
 $(152,7 \Omega)$

$$C_{gs} \sim 500 \text{ fF}$$

(170)

$$\frac{g_m}{C_{gs}} = 1,38 \cdot 10^{11} \rightarrow g_m \approx 6,9 \cdot 10^{-2}$$

$$W \approx 425 \mu\text{m}$$

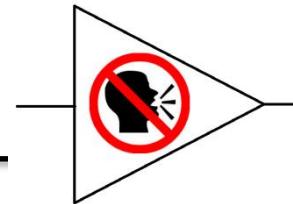
$$g_m = 0,0371 I_o \quad \rightarrow \quad \approx 11,6 \text{ mA}$$

$W 425 \mu\text{m}$

CONSUMO NO PUEDE SER OPTIMO.

GANANCIA POR CORRIENTE

Low Noise Amplifiers (LNA)



Inductive Source Degeneration. EXAMPLE:

NOTAS // Solo se tiene en cuenta el ruido térmico

MOSFET: $i_{ds}^2 = 4kT/g_m$ $R \quad i_{ds}^2 = \frac{4kT}{R}$
 $v_{ds}^2 = 4kT/g_m$ $v_{ds}^2 = 4kTR$

RUIDO Π_2 DESPRECIABLE (CASCODE, VER TEMA RUIDO)

RUIDO Π_B $R_{BIAS} \gg \rightarrow I_{bias} = \frac{4kT}{R}$

$R_{BIAS} \ll \rightarrow$ AISLA LA POLARIZACION

Efecto R_A y Π_1

EN ADAPTACION $\rightarrow G = \frac{g_m \cdot R_A}{2R_A \cdot C_{GS} \cdot S}$

$\hookrightarrow \overline{v_{out}^2} = 4kT_A R_A \left(\frac{g_m \cdot R_A}{2R_A \cdot C_{GS} \cdot S} \right)^2$

$\hookrightarrow \overline{v_{out}^2} = \frac{4kT}{g_m} \left(\frac{g_m \cdot R_A}{2R_A \cdot C_{GS} \cdot S} \right)^2$

Efecto R_s SE PUEDE DEMOSTRAR QUE LA GANANCIA ES LA MISMA ENTRANDO POR PUERTA O POR SURTIODOR

$\overline{v_{out}^2} = 4kT R_s \left(\frac{g_m \cdot R_A}{2R_A \cdot C_{GS}} \right)^2$

$$\overline{v_{out}^2} = 4kT R_A \left(|A_V|^2 \right) + \frac{4kT}{g_m} \left(|A_V|^2 \right) + \\ + \left(4kT R_s \left(|A_V|^2 \right) \right) + 4kT \cdot R_{eff}$$

No se tiene en cuenta por R_s

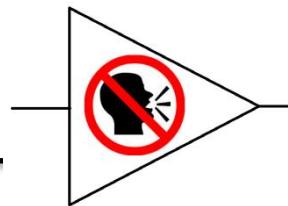
$$\overline{v_{out}^2} = 4kT_A R_A + 4kT \frac{Y}{g_m} + \frac{4kT \cdot R_{eff}}{\left(\frac{g_m \cdot R_{eff}}{2R_A \cdot C_{GS} \cdot W_o} \right)^2}$$

$$\overline{v_{out}^2} = 4kT_A \cdot R_A + 4kT_{LNA} \cdot \frac{Y}{g_m} + 4kT_{LNA} \cdot \frac{4R_A^2 C_{GS}^2 W_o}{R_{eff} \cdot g_m^2}$$

$$NF = \frac{\overline{v_{out}^2}}{4kT_A |A_V|^2 \cdot R_A} = 1 + \frac{Y T_{LNA}}{R_A g_m T_A} + \frac{\left(4R_A \cdot C_{GS}^2 \cdot W_o \right)}{R_{eff} \cdot g_m^2 \cdot T_A}$$

$$NF = 1 + \frac{T_{LNA}}{T_A} \left(\frac{Y}{R_A g_m} + \frac{4R_A C_{GS}^2 \cdot W_o^2}{R_{eff} \cdot (g_m)^2} \right)$$

Low Noise Amplifiers (LNA)



Inductive Source Degeneration. EXAMPLE:

$T_{LNA} > T_{ANTENA}$ → PEOR FACTOR DE RUIDO

$L \downarrow b \rightarrow g^2 \uparrow$ → PEOR FACTOR DE RUIDO

$R_A \uparrow$ → MEJOR, PERO EN 2°
TERMINO PUEDE "PESAR"
DEMASIADO

$R_{eff} \uparrow \uparrow$ → MEJOR

$g_{m1} \uparrow$ → MEJORA, PERO :

$$C_{GS}^2 \rightarrow \propto W^2$$
$$g_{m1}^2 \rightarrow \propto W$$

$\left| \begin{array}{l} \xrightarrow{W \uparrow} \\ \xrightarrow{\text{PEOR NF}} \end{array} \right.$

$$g_m = \sqrt{2 I_D \frac{W}{L} \cdot K_p} \rightarrow g_m \uparrow \rightarrow I_D \uparrow$$

VALOR DE NF

$$1 + \frac{T_{LNA}}{T_A} \left(\frac{2\pi/3}{210 \cdot g_{m1}} \right) + \frac{4 \cdot L_s}{R_{eff} \cdot R_A} \cdot \underbrace{\left(\frac{R_A \cdot C_{GS}}{g_{m1}} \right)^2 \cdot W_0^2}_{L_s = 1 \text{ nH}}$$

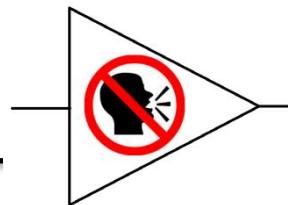
$6 \cdot 10^{-3}$

$$W = 425 \quad I_D = 11,6 \text{ mA} \rightarrow NF = 1 + 0,057 \quad (+ 6 \cdot 10^{-3})$$
$$< 1 \text{ dB}$$

VALOR MUY APROXIMADO

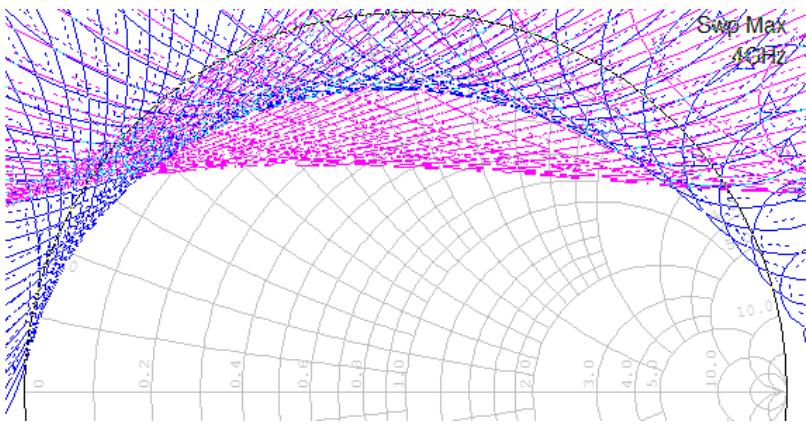
INTERESA LA TENDENCIA
CON OTROS PARAMETROS

Low Noise Amplifiers (LNA)

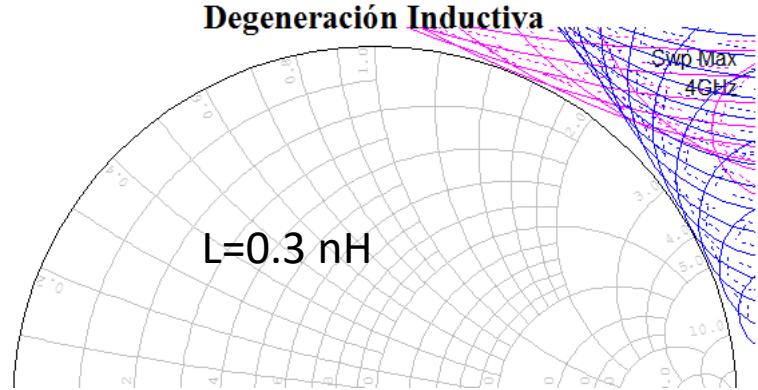


Inductive Source Degeneration.

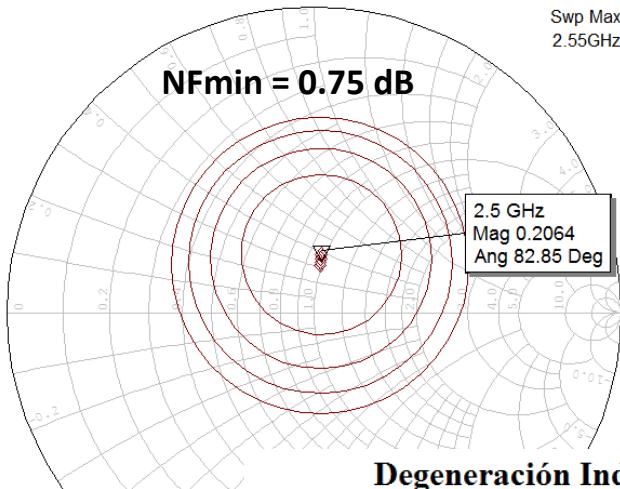
Potentially Unstable



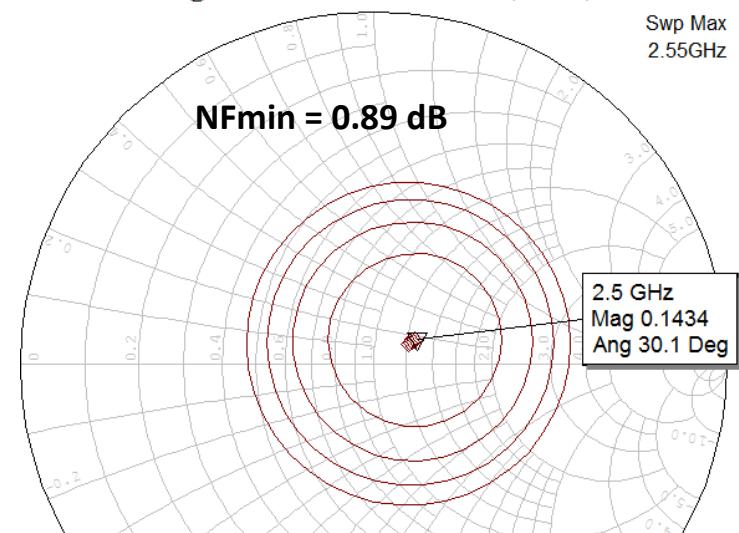
Unconditionally Stable



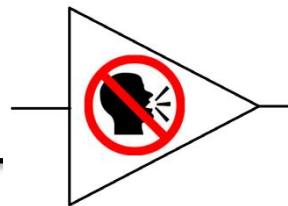
Degeneración Inductiva (0 nH)



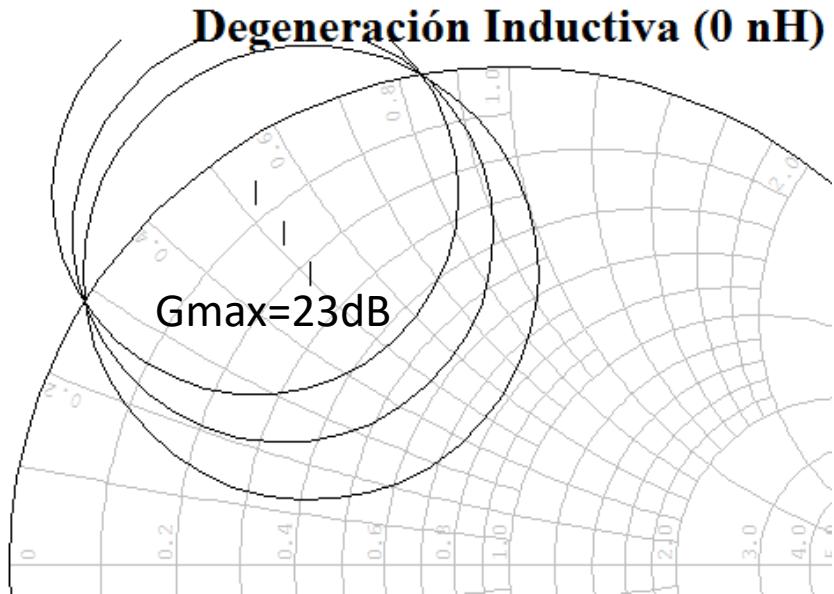
Degeneración Inductiva (1 nH)



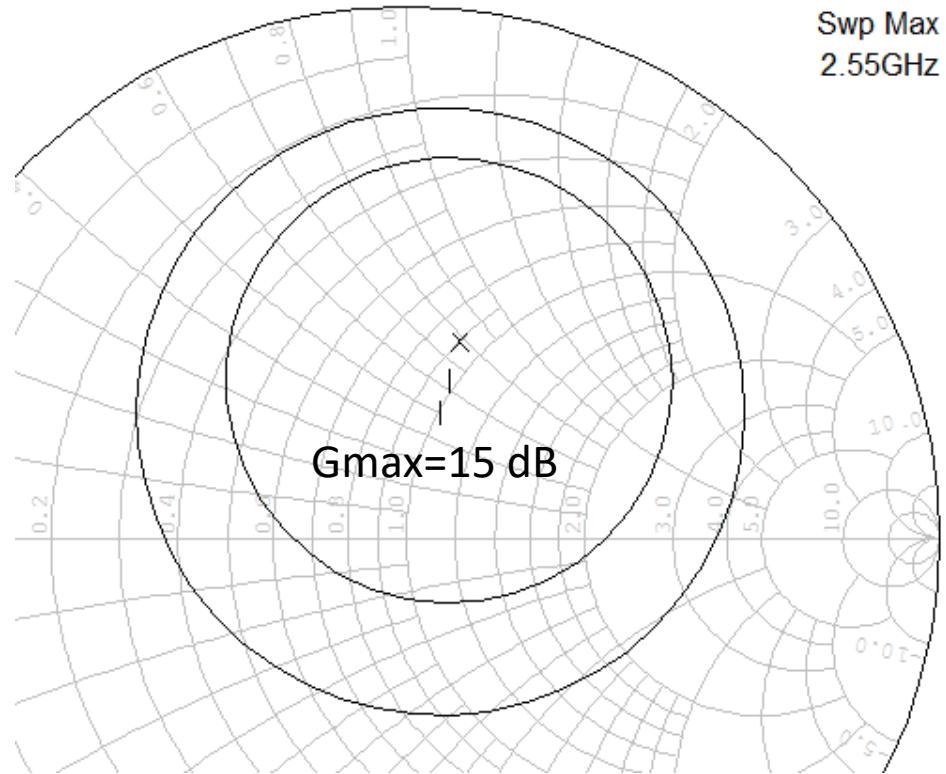
Low Noise Amplifiers (LNA)



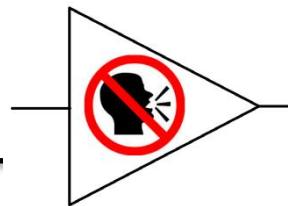
Inductive Source Degeneration. Conclusions



Degeneración Inductiva (0.3 nH)

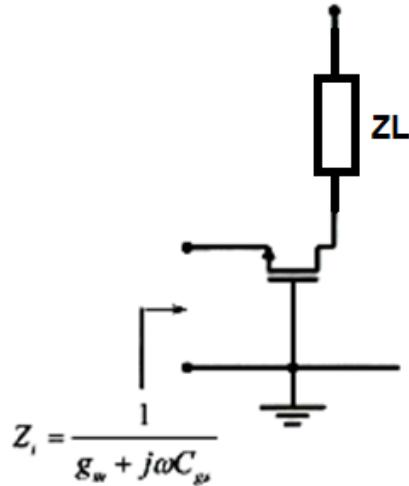


Low Noise Amplifiers (LNA)

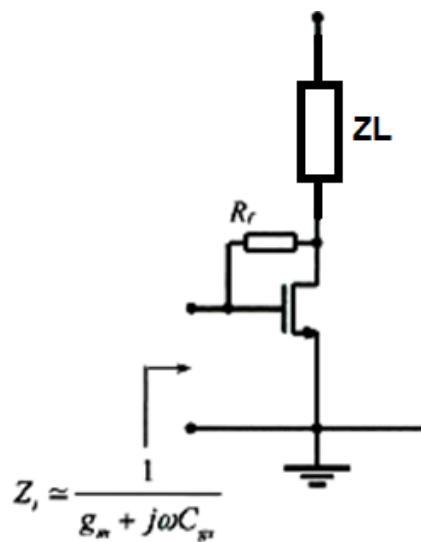


- Device Level Issues

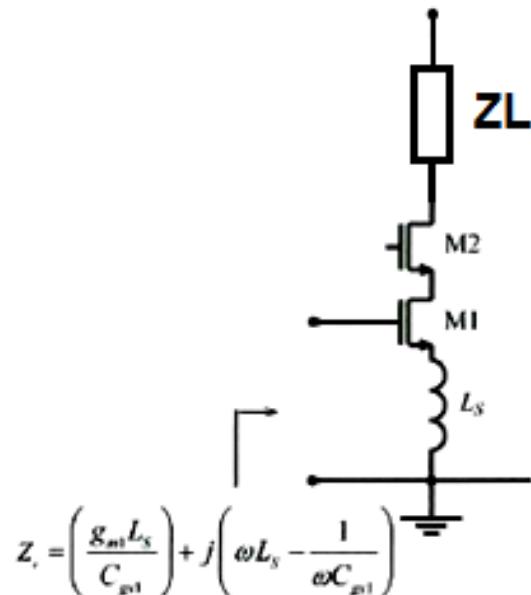
Useful Topologies



Common - GATE

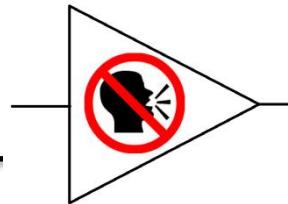


Common - SOURCE
(with negative feedback)



Common – SOURCE CASCODE
Inductive Degeneration

Low Noise Amplifiers (LNA)

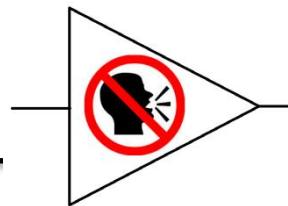


- Device Level Issues

Useful Topologies

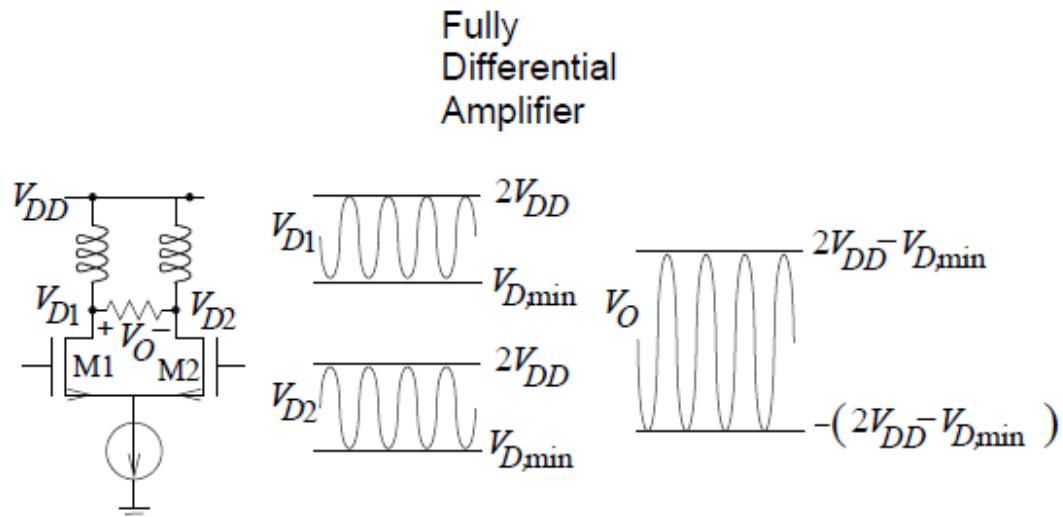
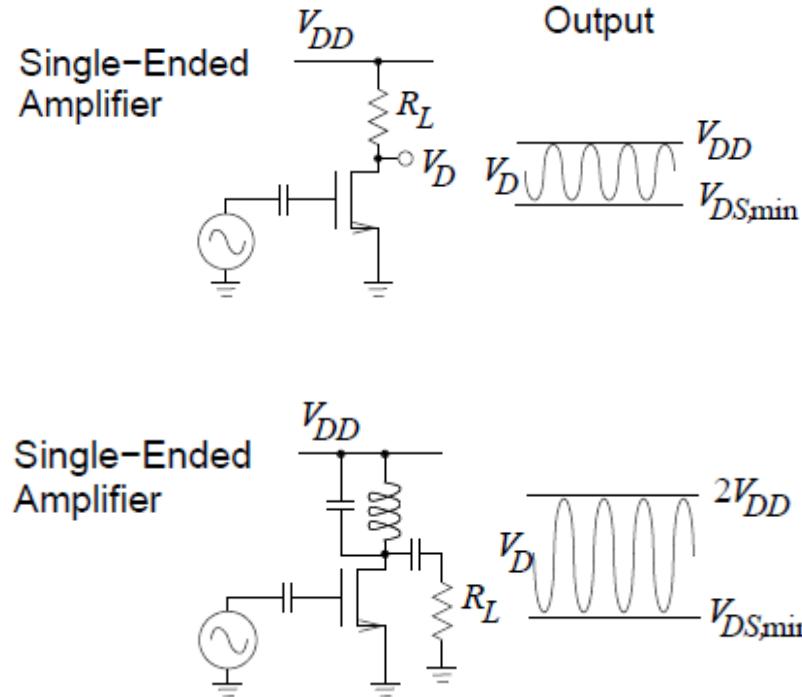
Characteristic	Common-Source	Common-Gate	Cascode
Noise Figure	Lowest	Rises rapidly with frequency	Slightly higher than CS
Gain	Moderate	Lowest	Highest
Linearity	Moderate	High	Potentially Highest
Bandwidth	Narrow	Fairly broad	Broad
Stability	Often requires compensation	Higher	Higher
Reverse Isolation	Low	High	High
Sensitivity to Process Variation, Temperature, Power Supply, Component Tolerance	Greater	Lesser	Lesser

Low Noise Amplifiers (LNA)

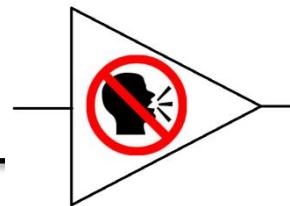


- Other design issues

Load Tuning (Output swing)

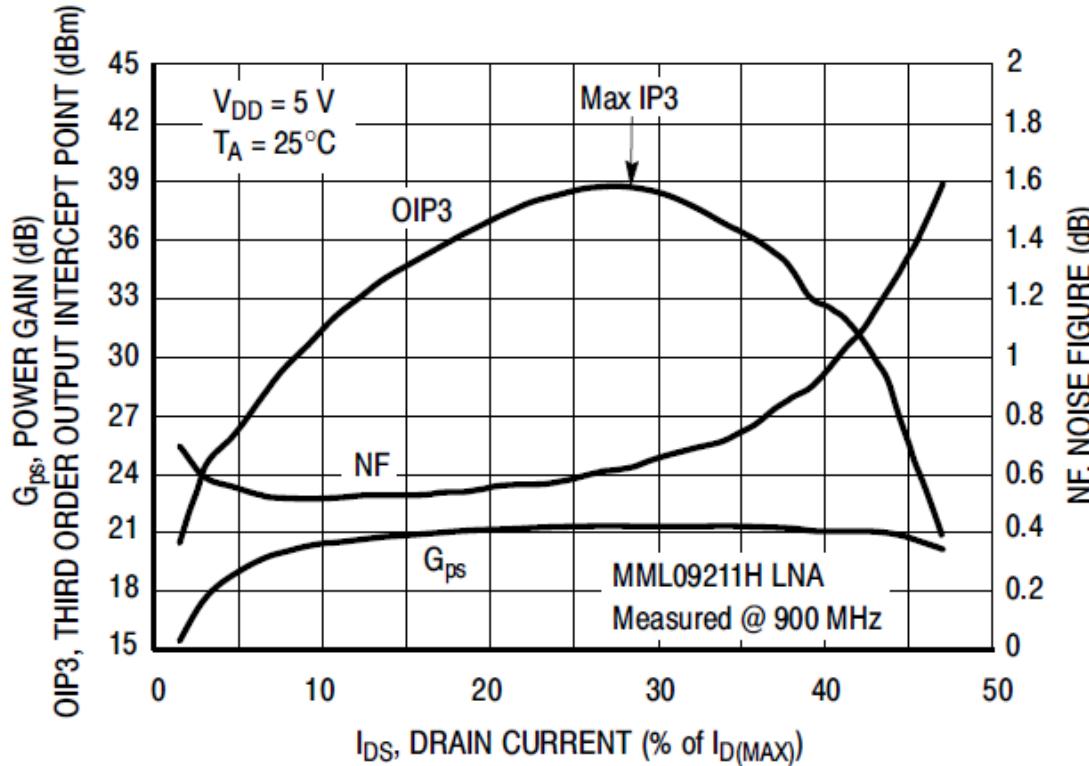


Low Noise Amplifiers (LNA)



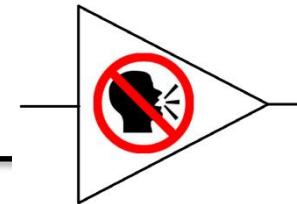
- Other design issues

Device parameter behavior as a function of BIAS point.



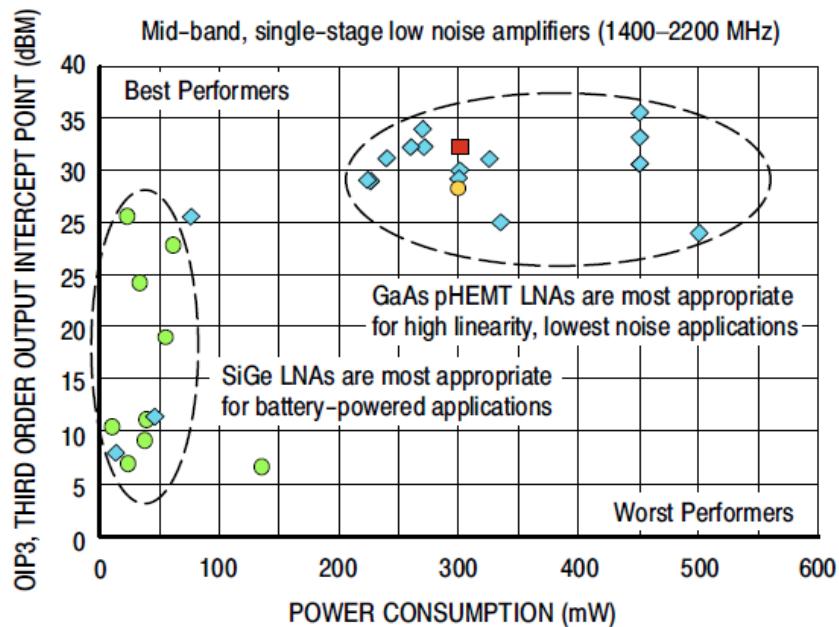
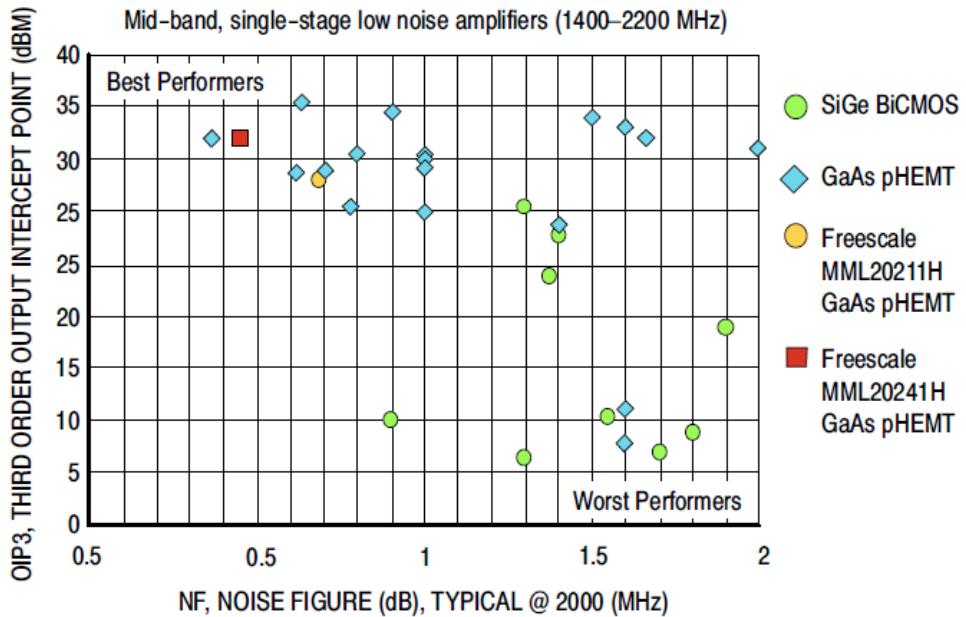
WATCH OUT!! DO NOT FORGET ABOUT POWER CONSUMPTION (BATTERY POWERED DEVICES)

Low Noise Amplifiers (LNA)



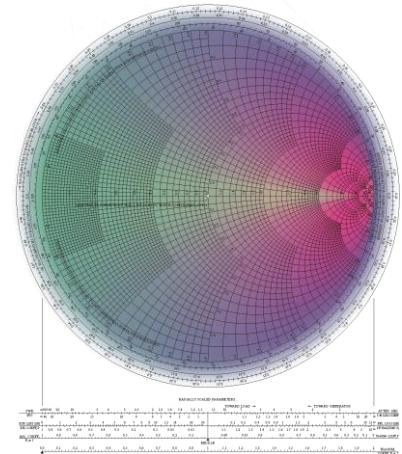
- Other design issues

CHOOSING DEVICE TECHNOLOGY



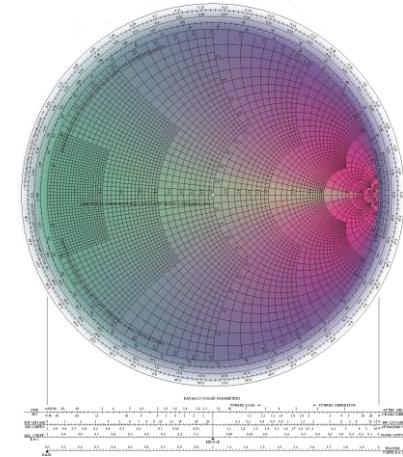
Circuitos Electrónicos De Alta Frecuencia

C4.1 – PASSIVE DEVICES in RF INTEGRATED CIRCUITS
ACTIVE DEVICES FOR RF



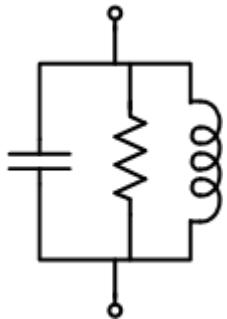
INDEX

- Introduction
 - Resonant circuits
 - Skin Effect
 - Impedance matching
 - Integrated Circuits Manufacturing Process
- Passive Devices in RF IC
- Active Devices
 - Mosfet
 - Bipolar
 - Diodes
- ESD protection
- Packages for IC



Introduction

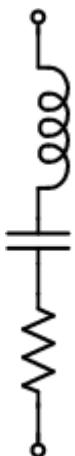
- A few words on Equivalent Circuits (... and Resonant Circuits)



$$Y = \frac{1}{R_P} + jwC + \frac{1}{jwL}$$

Resonance

$$w_0 = \frac{1}{\sqrt{LC}}$$



$$Z = R_S + jwL + \frac{1}{jwC}$$

$$R_p \approx R_S \cdot Q^2$$

NOTE: L & C have approx. same values

$$Q = w_0 \frac{\text{Stored Energy}}{\text{Dissipated Power}}$$

$$E_{tot} = \frac{1}{2} C \cdot (I_{pR} \cdot R_P)^2 \quad P_d = \left(\frac{I_{pR}}{\sqrt{2}} \right)^2 \cdot R_P$$

$$Q = \frac{R_P}{\sqrt{L/C}} = \frac{\sqrt{L/C}}{R_S}$$

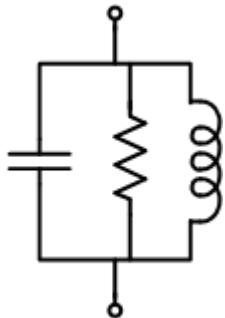
Characteristic Impedance*

$$E_{tot} = \frac{1}{2} L \cdot (V_{pR}/R_S)^2 \quad P_d = \left(\frac{V_{pR}}{\sqrt{2}} \right)^2 / R_S$$

* L & C reactance value at w_0

Introduction

- A few words on Equivalent Circuits (... and Resonant Circuits)



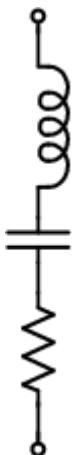
$$Y = \frac{1}{R_P} + jwC + \frac{1}{jwL}$$

$$I_{in} = \frac{|V|}{R_P} \rightarrow I_L = I_C = \frac{I_{in} \cdot R_P}{w_o L} = I_{in} \cdot R_P \cdot w_o C$$

$$I_L = I_C = Q \cdot I_{in}$$

Resonance

$$w_0 = \frac{1}{\sqrt{LC}}$$



$$Z = R_S + jwL + \frac{1}{jwC}$$

$$V_{in} = |I| \cdot R_S \rightarrow V_L = V_C = \frac{V_{in}/R_S}{w_o C} = (V_{in}/R_S) \cdot w_o L$$

$$V_L = V_C = Q \cdot V_{in}$$

L & C in resonant circuits with high Q may handle very high peak currents and voltages

Tesla Coil ++

Introduction

- Skin Effect

$$\nabla \cdot \mathbf{E} = \frac{\rho_v}{\epsilon} \quad (\text{Gauss' Law})$$

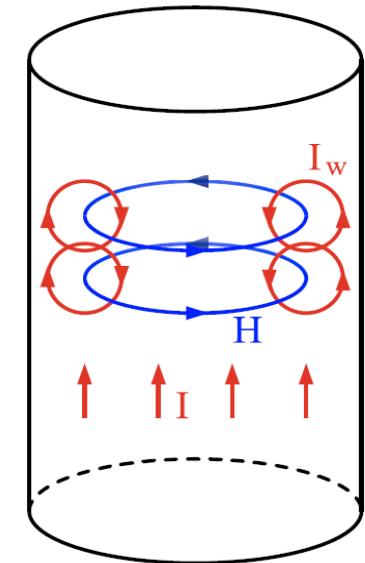
$$\nabla \cdot \mathbf{H} = 0 \quad (\text{Gauss' Law for Magnetism})$$

$$\nabla \times \mathbf{E} = -\mu \frac{\partial \mathbf{H}}{\partial t} \quad (\text{Faraday's Law})$$

$$\nabla \times \mathbf{H} = \mathbf{J} + \epsilon \frac{\partial \mathbf{E}}{\partial t} \quad (\text{Ampere's Law})$$

$$\frac{\partial \mathbf{J}}{\partial t} = \nabla \times \frac{\partial \mathbf{H}}{\partial t} \rightarrow \vec{E}$$

Induced current flows against primary current in the core of the conductor and adds to the primary current close to the surface

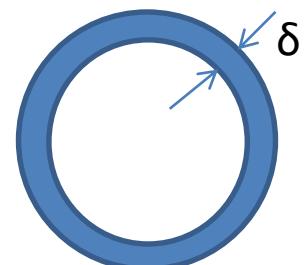


Current falls exponentially with distance to the surface:

$$J = J_S e^{-d/\delta}$$

$$\delta = \sqrt{\frac{2\rho}{\omega\mu_r\mu_0}}$$

Width of the layer where the current flow is confined



Introduction

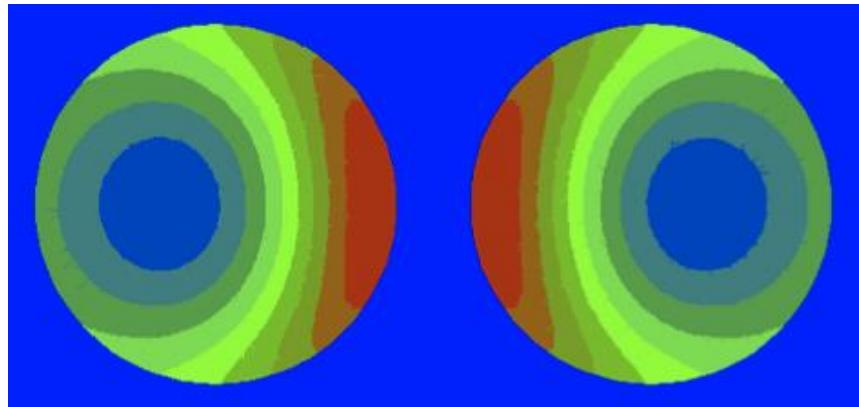
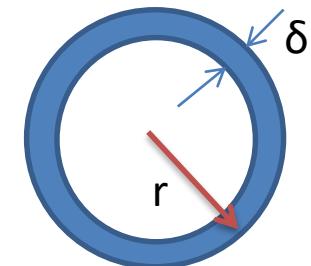
- The Skin Effect

Copper (1GHz) → $\delta=2 \mu\text{m}$
Aluminium (1GHz) → $\delta=2.5 \mu\text{m}$
Gold (1GHz) → $\delta=2.38 \mu\text{m}$

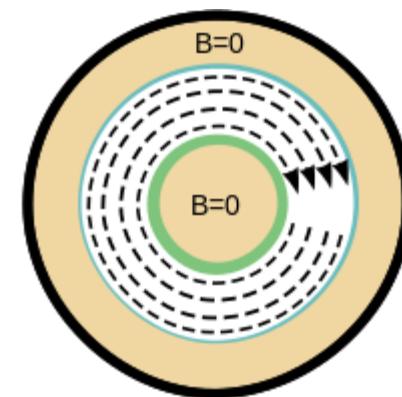


Resistivity
 ρ

$$R \approx \frac{\rho \cdot l}{2\pi r \cdot \delta}$$



Two wires line

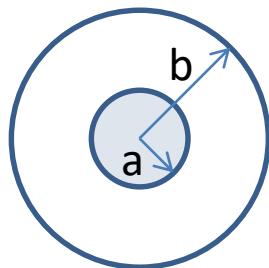


Coaxial

Impedance Matching

- Why 50Ω or 75Ω as “magic” impedance matching goals?

MAXIMUM POWER IN A COAXIAL CABLE



$$E_{max} = \frac{V}{a \cdot \ln(b/a)}$$

$$Z_o = \sqrt{\frac{\mu}{\epsilon}} \cdot \frac{\ln(b/a)}{2\pi}$$

Dielectric
strength limits
Pmax



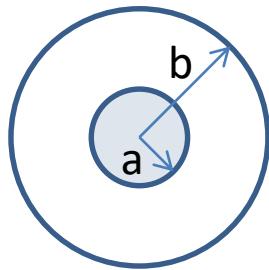
b ↑↑ increases max power but also increases Z_o (OPTIMAL for MAX POWER DELIVERY???)

$$P_{max} \propto \frac{V^2}{Z_o} \rightarrow \frac{b}{a} = \sqrt{e} \rightarrow Z_o = 30 \Omega$$

Impedance Matching

- Why 50Ω or 75Ω as “magic” impedance matching goals?

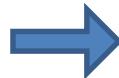
MINIMUM ATTENUATION IN A COAXIAL CABLE



$$\alpha \approx \frac{R}{2Z_o}$$

Attenuation is proportional to resistance per length unit*

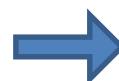
R due to *skin-effect*



a ↑↑ decreases
R but also
decreases Z_o

$$Z_o = \sqrt{\frac{\mu}{\epsilon} \cdot \frac{Ln(b/a)}{2\pi}}$$

Using an approximation for skin effect resistivity and derivating to get the optimum value for a



$$\frac{b}{a} = 3.6$$



$$Z_o \approx 75 \Omega$$

Impedance Matching

- Why 50Ω or 75Ω as “magic” impedance matching goals?

MAXIMUM POWER

$30\ \Omega$



MINIMUM ATTENUATION

$75\ \Omega$



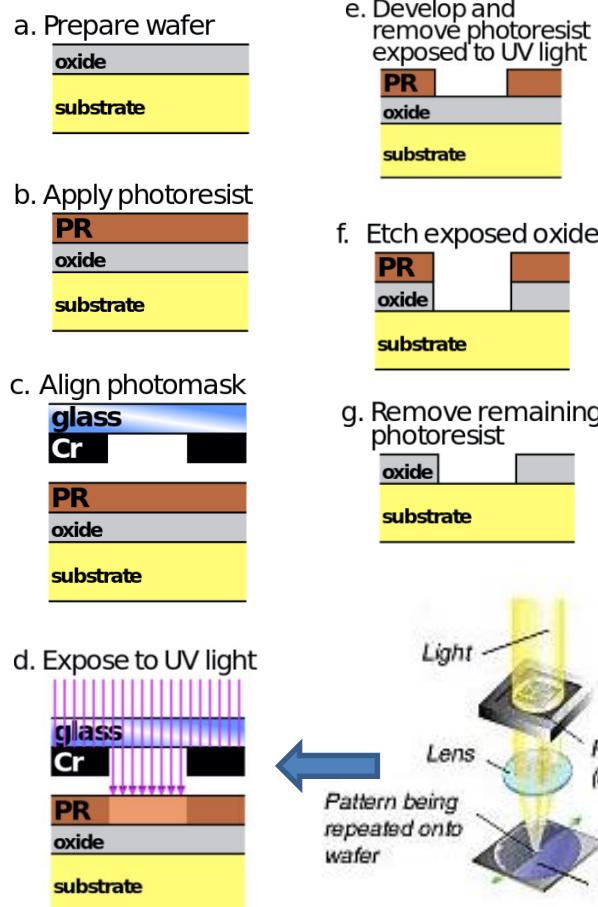
AS A TRADE OFF

$Z_o = 50\ \Omega$

Introduction

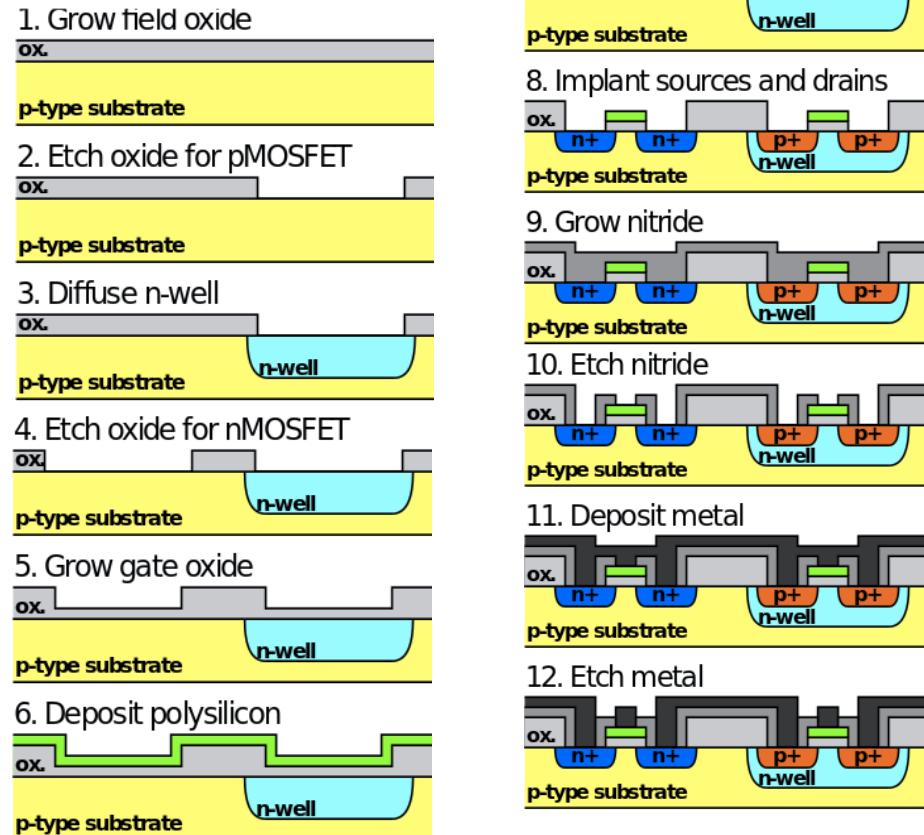
- Integrated Circuit Manufacturing Process

Basic Photolithographic Process

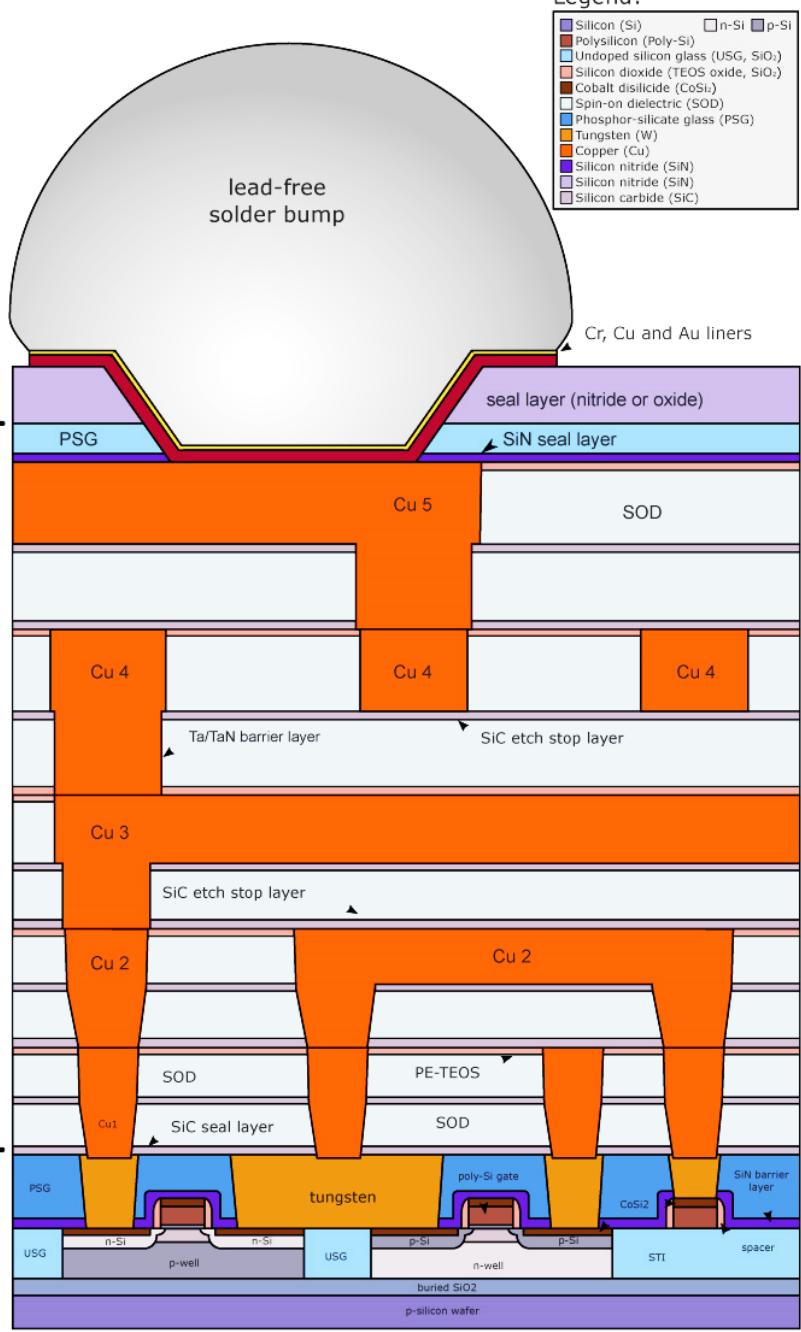


Typical CMOS process

Cada paso que necesita una máscara
(etch, implantación etc.) Proceso de
Fotolitografía



front-end
FEOL | "Advanced Packaging"
back-end /
"Advanced Packaging"



Legend:

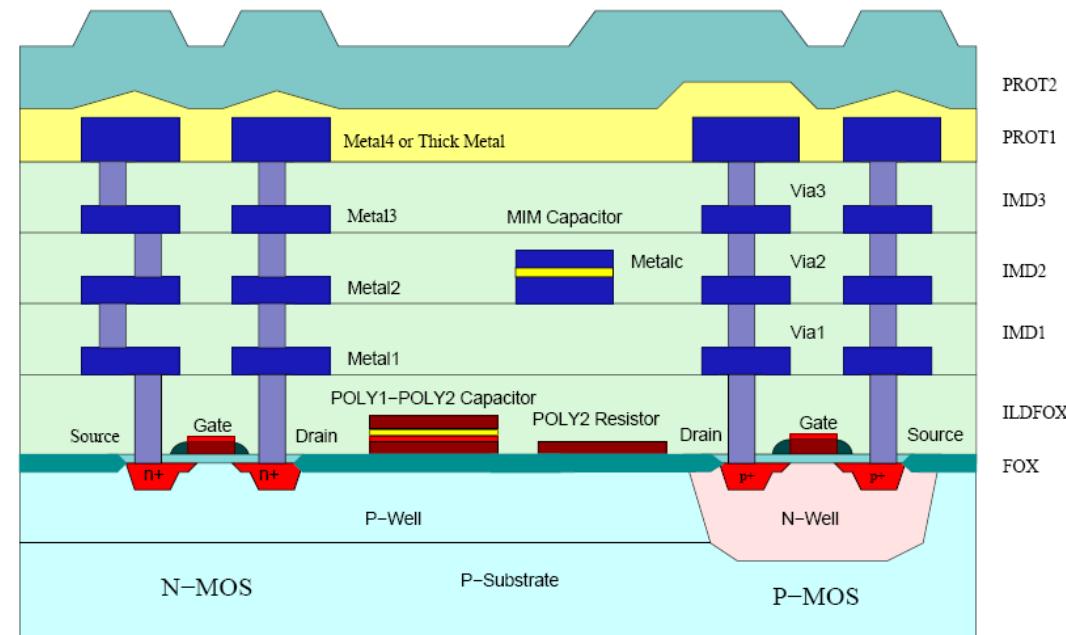
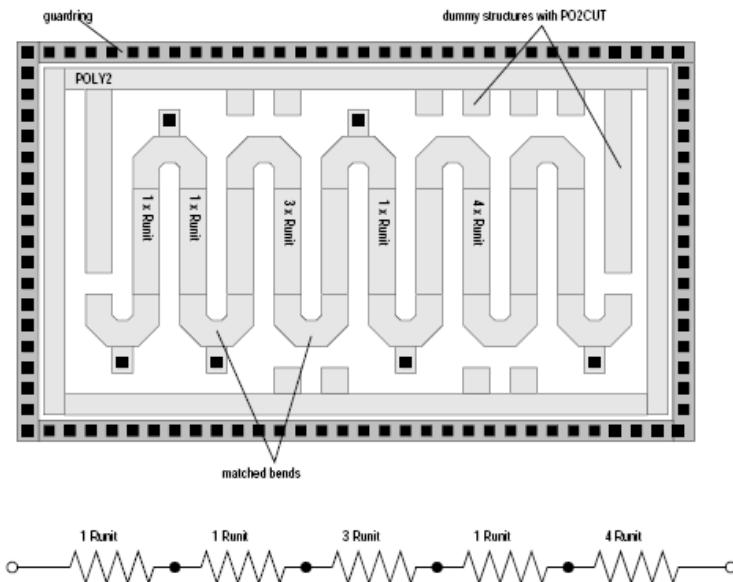
Silicon (Si)	n-Si	p-Si
Polysilicon (Poly-Si)		
Undoped silicon glass (USG, SiO_2)		
Silicon dioxide (TEOS oxide, SiO_2)		
Cobalt disilicide (CoSi_2)		
Spin-on dielectric (SOD)		
Phosphor-silicate glass (PSG)		
Tungsten (W)		
Copper (Cu)		
Silicon nitride (SiN)		
Silicon carbide (SiC)		

Legend:

Silicon (Si)	n-Si	p-Si
Polysilicon (Poly-Si)		
Undoped silicon glass (USG, SiO_2)		
Silicon dioxide (TEOS oxide, SiO_2)		
Cobalt disilicide (CoSi_2)		
Spin-on dielectric (SOD)		
Phosphor-silicate glass (PSG)		
Tungsten (W)		
Copper (Cu)		
Silicon nitride (SiN)		
Silicon carbide (SiC)		

Passive Devices in RF Integrated Circuits

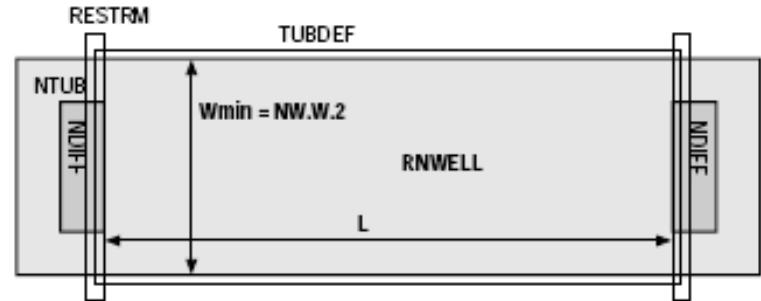
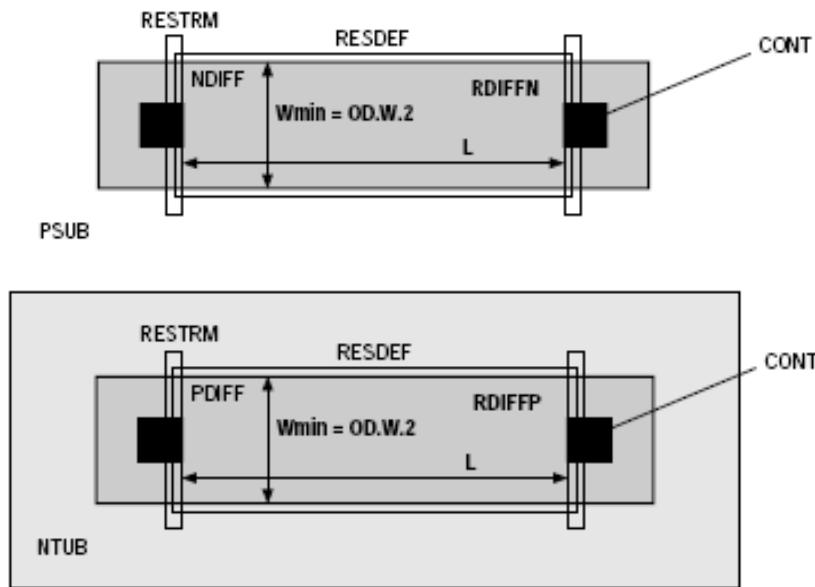
- Resistors in Integrated Circuits
 - Polisilicon Resistors (Po2 & Po2 HR)
 - Tolerance $\pm 20\%$
 - A minimum number of squares and a minimum width is required for a good matching
 - High resistivity doped polisilicon available ($> 1.2 \text{ k}\Omega/\square$)
 - Low temperature coefficients (900 ppm/ $^{\circ}\text{C}$)



Passive Devices in RF Integrated Circuits

- Diffusion and Well Resistors

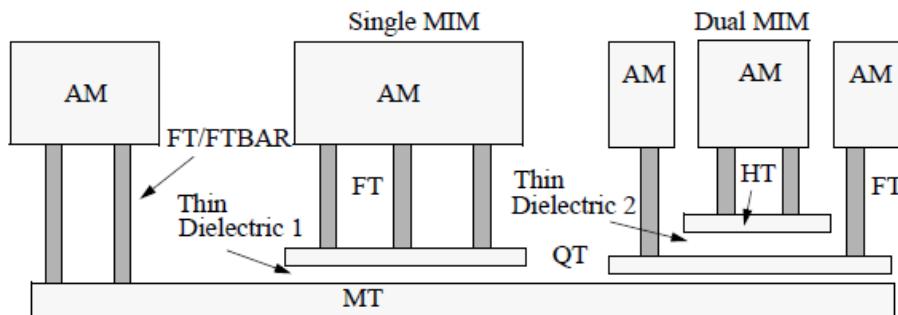
- Tolerance $\pm 10\%$
- Well $\rightarrow 1 \text{ k}\Omega/\square$ / Diffusion $\rightarrow 10 \text{ }\Omega/\square \uparrow \sim 150 \text{ }\Omega/\square \uparrow$
- Well \rightarrow C. Temp. ($6200 \text{ ppm}/^\circ\text{C}$) / Diffusion \rightarrow C. Temp. ($1500 \text{ ppm}/^\circ\text{C}$)
- Noise sensitive. Bad isolation



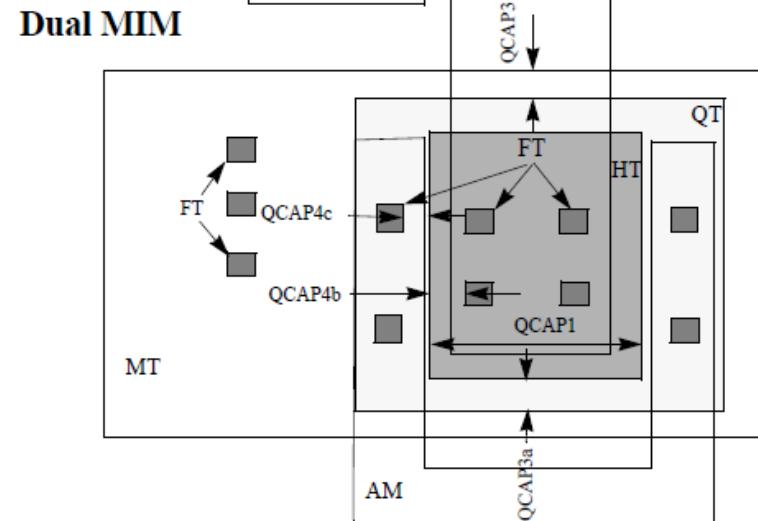
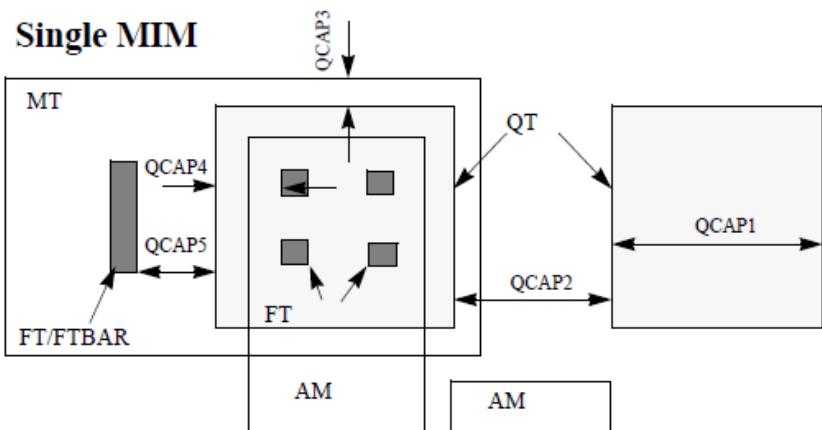
Passive Devices in RF Integrated Circuits

- Integrated Capacitors (MIM)

- ✓ Tolerance $\pm 15\%$
- ✓ MIM lower substrate interference
- ✓ Linear and very Low temperature coefficient ($18 \sim 45$) ppm/ $^{\circ}\text{C}$
- ✓ Very Low BIAS Voltage coefficient (-0.51 ppm/V)



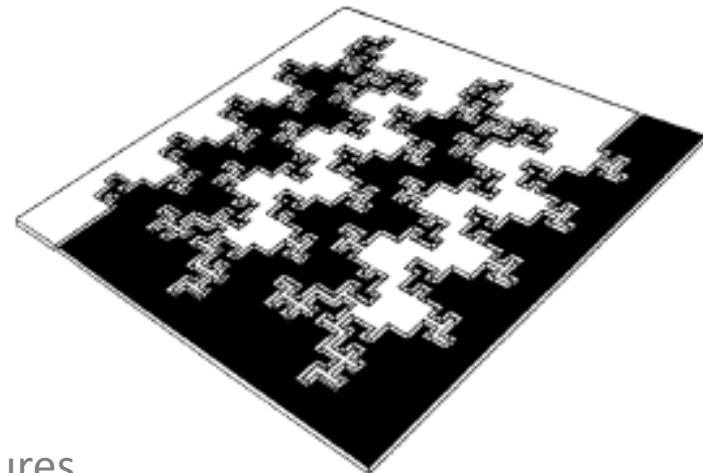
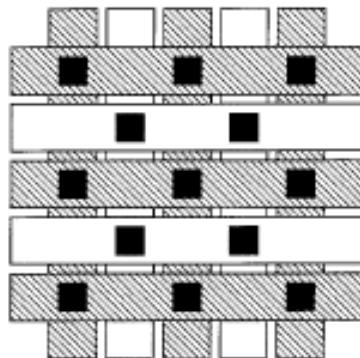
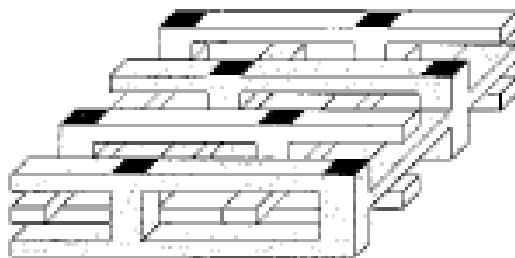
Metal TOP (highest metallization level). Far from substrate



Passive Devices in RF Integrated Circuits

- Integrated Capacitors (MIM). WOVEN & FRACTALS

- ✓ Side capacitances
- ✓ Useful when the manufacturing process doesn't have special dielectrics.

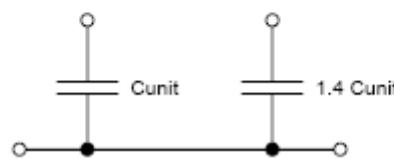
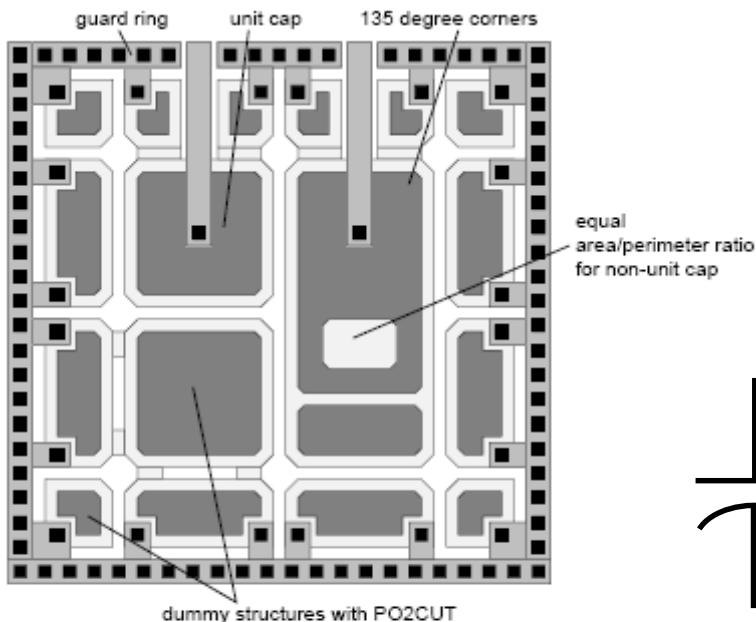


Fractal structures
maximize Perimeter/ Area
ratio.
Higher capacitances can
be achieved
(Koch Cells)

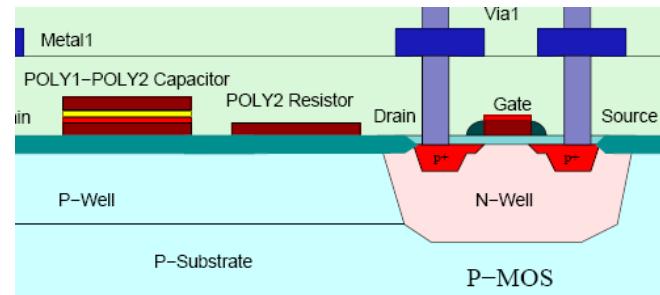
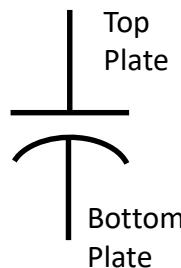


Passive Devices in RF Integrated Circuits

- Integrated Capacitors (CPOLY)



- ✓ Tolerance $\pm 15\%$
- ✓ High substrate effect (up to 30% of the nominal value)
- ✓ Very bad Temp. Coefficient 30000 ppm/ $^{\circ}\text{C}$
- ✓ Low BIAS Voltage coefficient (85 ppm/V)



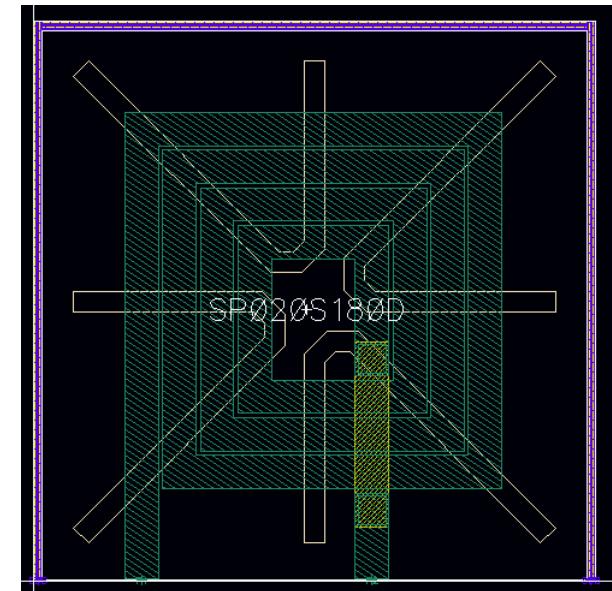
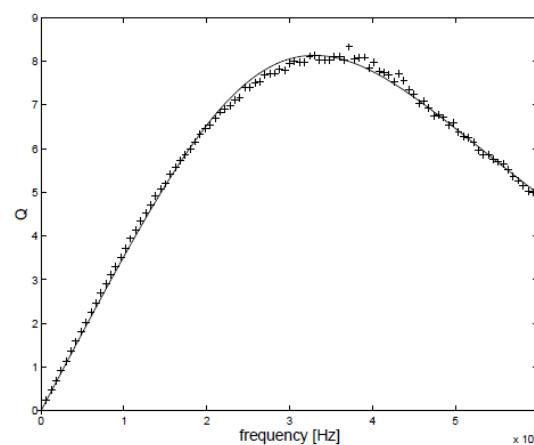
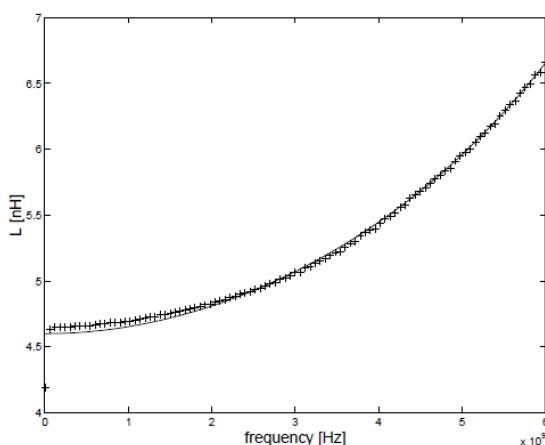
OBSOLETE. Not available in new processes

Passive Devices in RF Integrated Circuits

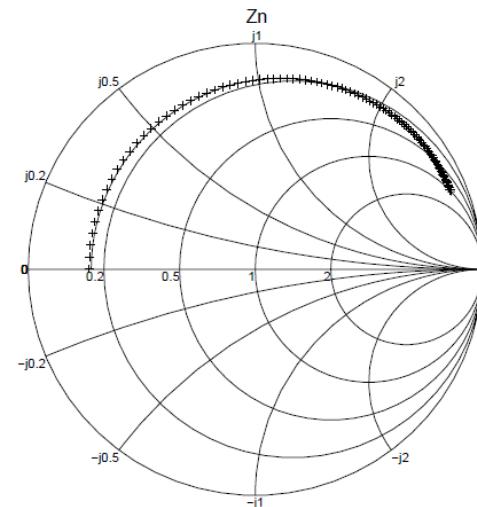
- Integrated Inductors

4.2.3 Main Parameters of Thick Metal Inductors

Inductor	Inductance [nH]			Qmax	Q		fRES [GHz]
	Ls	@2.4GHz	@5.0GHz		@2.4GHz	@5.0GHz	
SP011S200T	1.07	1.04	1.05	11.9@4.4GHz	8.7	11.8	>6
SP015S250T	1.52	1.49	1.58	11.7@3.9GHz	9.6	10.2	>6
SP020S250T	2.02	2.04	2.17	10.4@3.9GHz	7.9	9.8	>6
SP021S200T	2.10	2.10	2.20	9.9@3.9GHz	7.8	9.3	>6
SP024S250T	2.42	2.42	2.75	9.6@2.70GHz	9.3	6.5	>6
SP031S250T	3.07	3.17	3.63	8.9@3.0GHz	8.4	6.6	>6
SP033S150T	3.25	3.32	3.52	9.4@4.33GHz	6.9	9.0	>6
SP037S250T	3.67	3.82	4.58	9.4@2.70GHz	9.0	6.2	>6
SP047S250T	4.66	4.90	5.98	8.3@3.7GHz	7.2	6.4	>6
SP049S300T	4.85	5.30	8.13	8.8@2.46GHz	8.6	3.7	>6
SP060S300T	6.00	6.59	10.09	7.2@3.0GHz	6.8	4.3	>6
SP073S250T	7.25	8.11	12.20	7.7@2.5GHz	7.5	3.2	>6
SP100S250T	10.02	12.08	19.23	7.2@2.0GHz	6.8	1.1	6
SP133S300T	13.30	19.06	-	6.7@1.7GHz	5.5	-	4.3

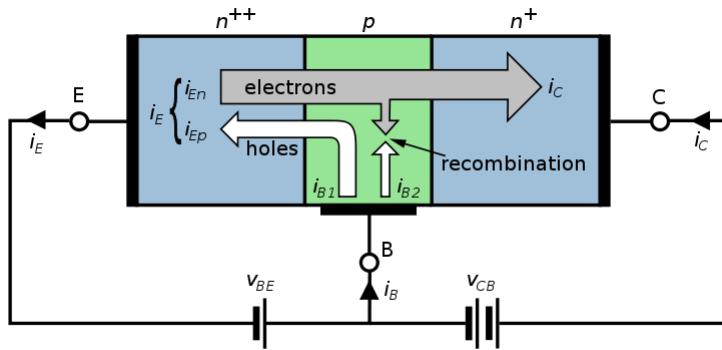


AMS C35 M4 Inductor.
Same level as thick metal but lower thickness



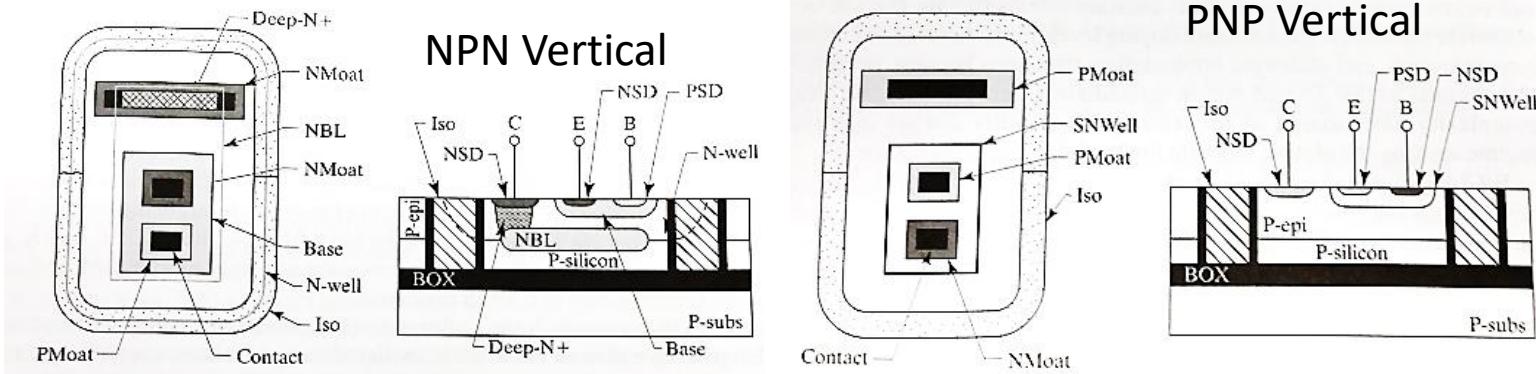
Active Devices

- Bipolar Junction Transistor (BJT)



- BE diode is biased in direct mode. Carrier current flows are generated (h- & e++)
- Minority carriers in base junction (e++) are accelerated by the electric field accross the BC juntion depletion zone
- In order to avoid recombination along B that would reduce minority carriers (e++) flow B must be very thin ($t_{tran} < t_{recomb}$)

Integrated structures are complex and “difficult” to manufacture



Deep-trench or dielectric isolation is required

Active Devices

- Bipolar Junction Transistor (BJT). Linear Model (AC)

$A_E \rightarrow$ Emitter junction area

$D_n \rightarrow$ Diffusion Coefficient

$W_B \rightarrow$ Base junction width

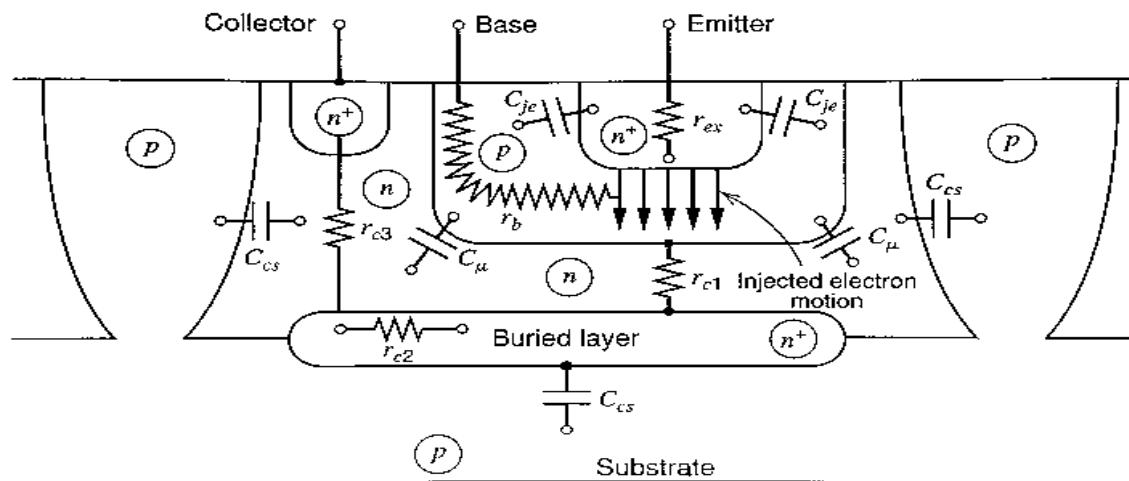
$N_A \rightarrow$ Doping density

$L_p \rightarrow$ Diffusion Length of minority carriers

$\tau_b \rightarrow$ Carriers lifetime at the Base

$$i_c \approx \underbrace{\frac{A_E q D_n n_i^2}{N_A W_B}}_{I_S} \cdot e^{\frac{v_{BE}}{V_T}}$$
$$i_B \approx I_S \cdot \left(\frac{D_p N_A W_B}{D_n N_D L_p} + \frac{W_B^2}{2 D_n \tau_b} \right) \cdot e^{\frac{v_{BE}}{V_T}}$$

$$i_c \approx i_B \cdot \beta$$

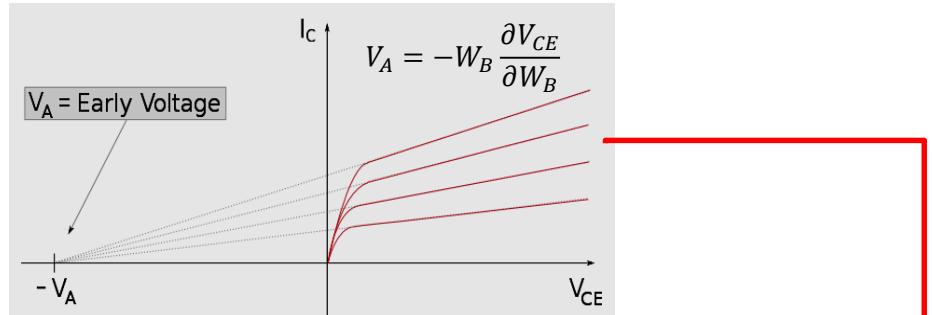


Active Devices

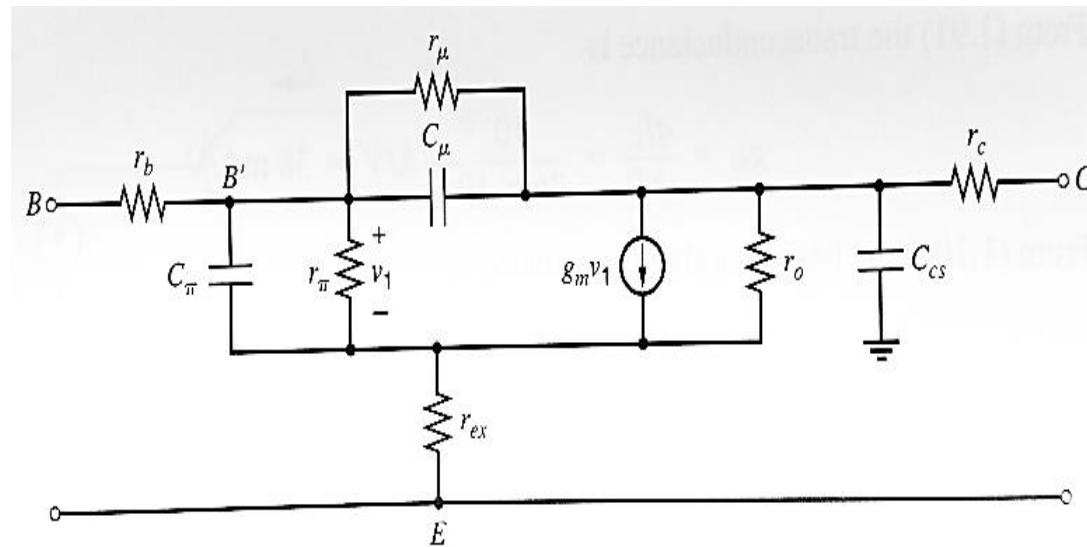
- Bipolar Junction Transistor (BJT). Linear Model (AC)

Early Effect

$$i_c = i_s \left(1 + \frac{v_{CE}}{V_A}\right) e^{\frac{v_{BE}}{V_T}}$$



$$r_\pi = \frac{\beta_{BIAS}}{g_m}$$
$$C_\pi = \tau_F g_m + C_{je}$$



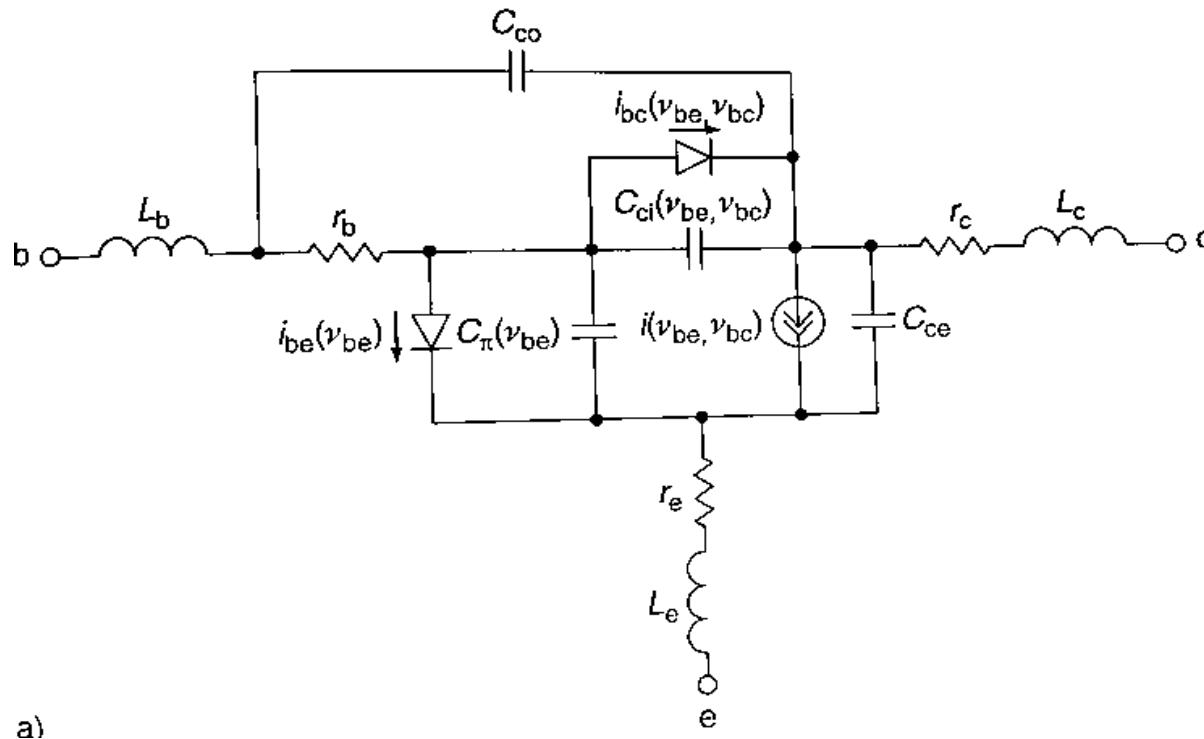
$$g_m = \frac{I_C}{V_T}$$

$$r_o = \frac{V_A}{I_C}$$

$$r_\mu = r_o \beta_{BIAS}$$

Active Devices

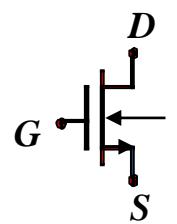
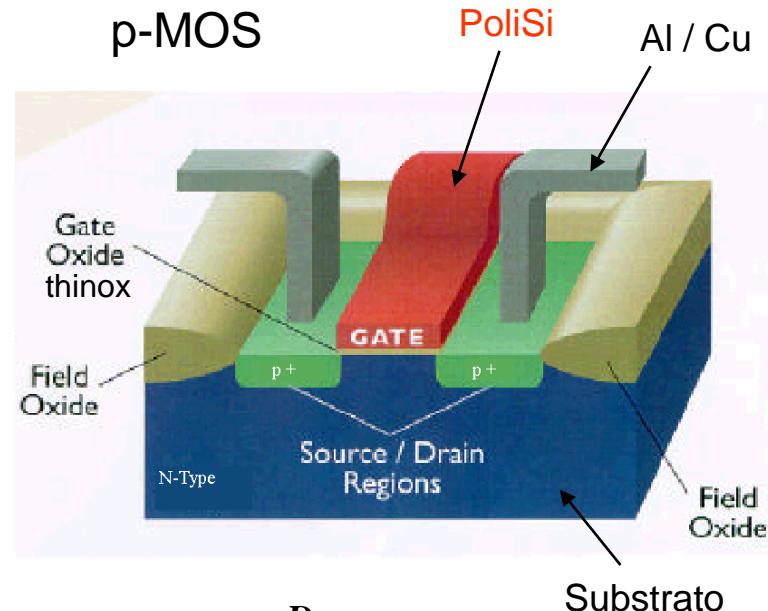
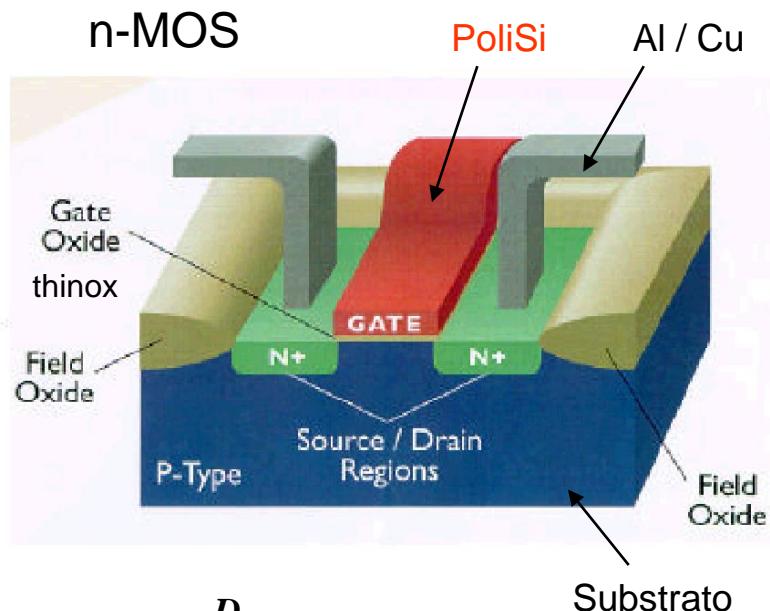
- Bipolar Junction Transistor (BJT). Non Linear Model (Modified Gummel-Poon)



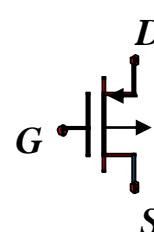
S parameter matrix based model can be obtained directly from Gummel-Poon model *

Active Devices

- Metal Oxide Semiconductor Field Effect Transistor (MOSFET)



n-MOS
Enhancement



p-MOS
Enhancement

Active Devices

- Metal Oxide Semiconductor Field Effect Transistor (MOSFET)

$$I_c = \frac{I_D}{I_0 \cdot (\frac{W}{L})} \quad \text{Inversion Index}$$

STRONG INVERSION ($I_c > 10$)

$$I_{DS} = \frac{Kp}{2.n} \cdot \frac{W}{L} [(V_{GS} - V_T)^2]$$

$$I_{DS} = \frac{Kp}{2.n} \cdot [(V_{GS} - V_T)^2] \cdot [1 + \lambda \cdot V_{DS}]$$

Channel Length Modulation Effect

Linear Region

$$V_{DS} < V_{GS} - V_T$$

$$I_{DS} = \frac{Kp}{2.n} \cdot \frac{W}{L} [(V_{GS} - V_T) \cdot V_{DS} - \frac{V_{DS}^2}{2}]$$

Saturation Region

$$V_{DS} > V_{GS} - V_T$$

MODERATE INVERSION ($I_c = 1$)

$$I_D = I_0 \frac{W}{L} \left(\ln \left(1 + e^{\frac{V_{GS} - V_T}{2nU_T}} \right) \right)$$

WEAK INVERSION ($I_c < 0.1$)

$$I_D = I_0 \frac{W}{L} \left(e^{\frac{V_{GS} - V_T}{nU_T}} \right)$$

$$\text{BJT} \rightarrow I_C = I_S \left(e^{\frac{V_{BE}}{U_T}} \right)$$

Gate areas needed for Moderate and Weak inversion are not compatible with HF requirements

Active Devices

- Metal Oxide Semiconductor Field Effect Transistor (MOSFET)

Saturation Region

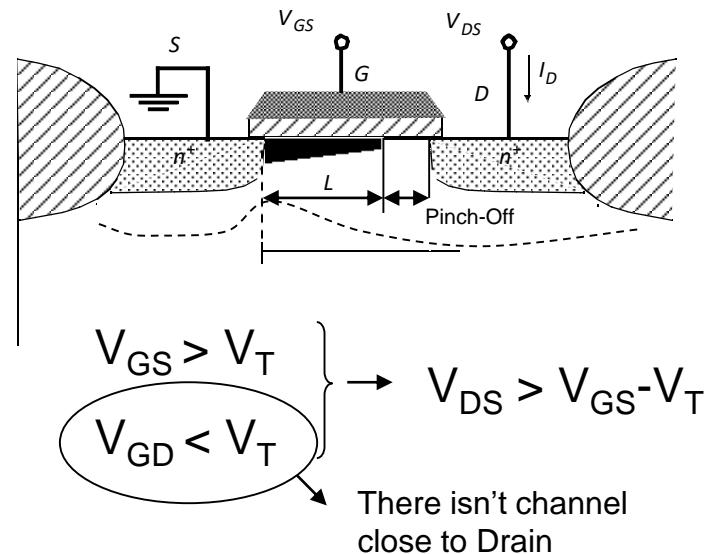
$$V_{DS} > V_{GS} - V_T$$

STRONG INVERSION ($I_c > 10$)

$$I_{DS} = \frac{Kp}{2.n} \cdot \frac{W}{L} [(V_{GS} - V_T)^2]$$

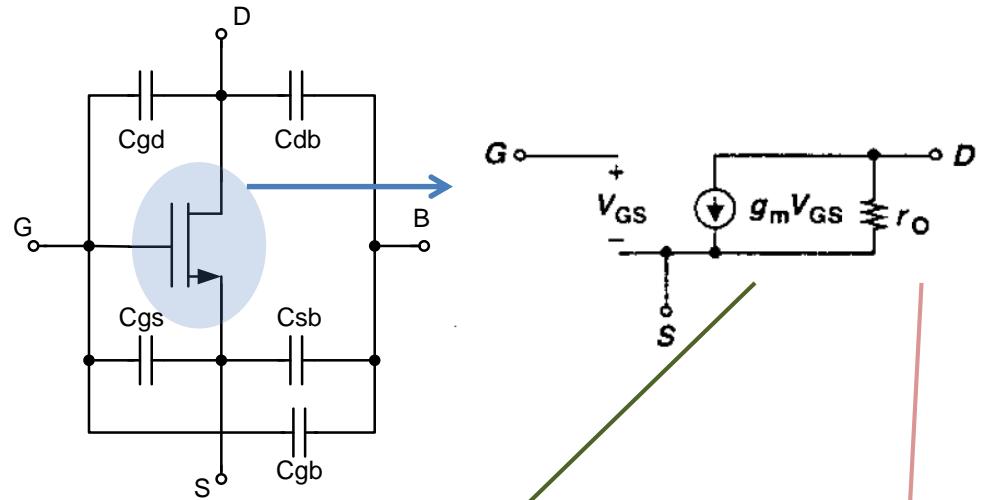
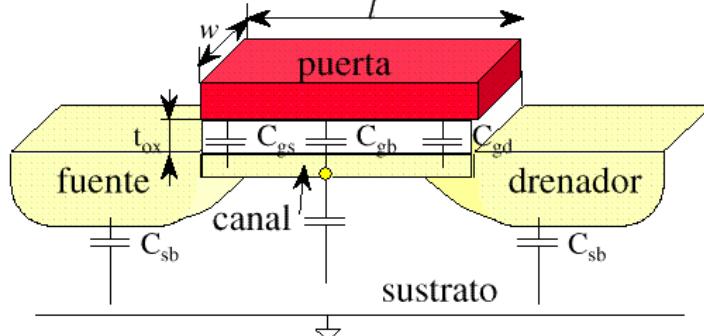
$$I_{DS} = \frac{Kp}{2.n} \cdot [(V_{GS} - V_T)^2] \cdot [1 + \lambda \cdot V_{DS}]$$

Channel Length Modulation Effect



Active Devices

- Metal Oxide Semiconductor Field Effect Transistor (MOSFET)



Strong Inversion

$$g_m = \sqrt{2I_D \frac{K_P}{n} \left(\frac{W}{L} \right)}$$

Weak Inversion

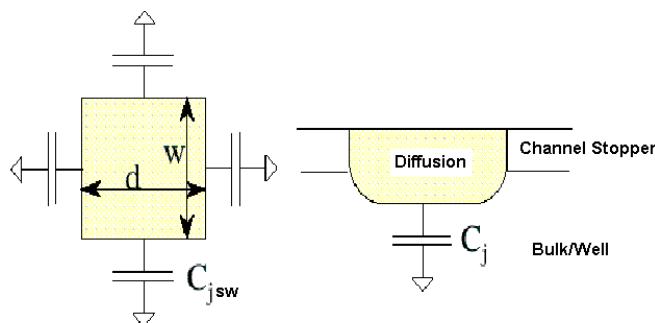
$$g_m = \frac{I_D}{n \cdot U_T}$$

$$r_d = \frac{1}{\lambda \cdot I_D}$$

λ inverse
dependent with L

Moderate Inversion

$$g_m = \sqrt{2(I_D - I_{Dmin}) \frac{K_P}{n} \left(\frac{W}{L} \right)}$$



Active Devices

- Metal Oxide Semiconductor Field Effect Transistor (MOSFET)

Saturation Velocity:

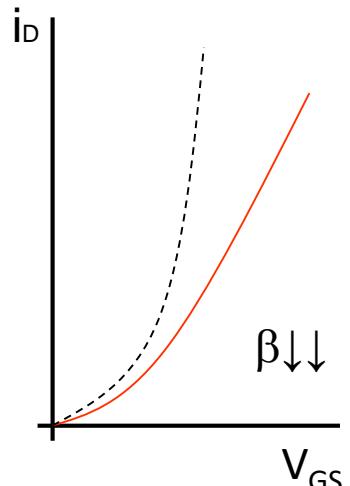
Pinch-off zone fills the channel length L ($V_D \gg V_G$)

$$v = \mu E \quad v_{SAT} \rightarrow \text{Maximum speed of the carriers in the channel}$$

$$Q_C = C_{ox} \cdot W \cdot L \cdot (V_{GS} - V_t)$$

$$I_{DS} = v_{SAT} \cdot C_{ox} \cdot W \cdot [(V_{GS} - V_t)] = \mu_n E_{crn} \cdot C_{ox} \cdot W \cdot [(V_{GS} - V_t)] = \mu_n \cdot C_{ox} \cdot \frac{W}{L} \cdot [(V_{GS} - V_t)] \cdot V_{DS(vsat)}$$

$$V_{DS(vsat)} = \frac{L \cdot v_{SAT}}{\mu_n} \rightarrow \begin{aligned} &\text{nMOS (L=130nm)} \\ &V_{DS(Vsat)} = 0.34V \end{aligned}$$



- For NMOS V_{DS} is the most restrictive variable
- High V_{DS} (AMSC35 > 0.8 V) are dangerous
- With low V_{DS} close to ΔV the effect is only measurable in small technologies $L < 180\text{nm}$

Active Devices

- MOSFET Figure of Merit (f_T)

Frequency where MOSFET input current is equal to the output current

$$f_T = \frac{1}{2\pi} \frac{\mathbf{g}_m \cdot V_{GS}}{\mathbf{C}_{gs} \cdot V_{GS}} \rightarrow f_T = \frac{1}{2\pi} \frac{\sqrt{2 \cdot I_D \cdot \frac{\mathbf{K}_{PP}}{n} \cdot \frac{W}{L}}}{W \cdot L \cdot C_{ox} \left(\frac{2}{3} + \frac{C_{GSO}}{LC_{ox}} \right)}$$

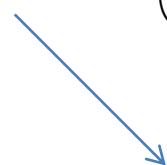
n: Slope Factor



$$f_T \propto \frac{1}{L^{3/2}}$$
$$f_T \propto \frac{1}{W^{1/2}}$$

Channel length L is
the most important
factor in f_T

$$C_{gs} = W \cdot L \cdot C_{ox} \left(\frac{2}{3} + \frac{C_{GSO}}{LC_{ox}} \right)$$



$$\frac{2}{3} \text{ de } C_G + W \cdot C_{GSO}$$

For v_{SAT} conditions dependency with L is
lower but still very important

Active Devices

- MOSFET. Maximum working frequency (f_c)

Frequency for a -3 dB attenuation of the signal at the gate
(Distributed Resistance and Capacitance act as a filter)

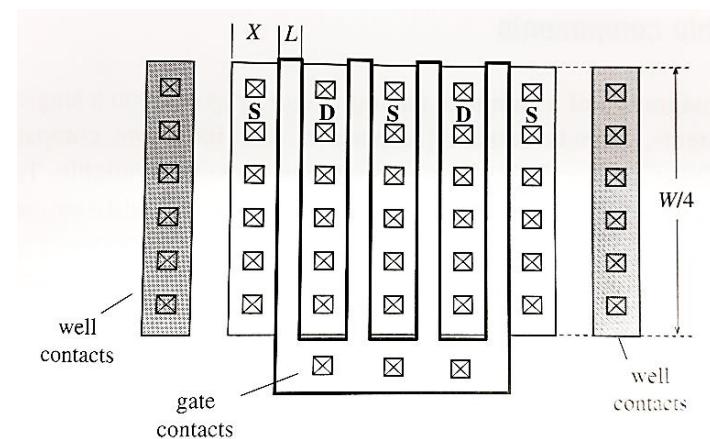
$$* f_c = \frac{0.124}{R_{sh} C_{ox} W^2}$$

Rpoly Gate Oxide capacitance

**Must be careful with W
(MULTIFINGER)**

AMS C35 / NMOS (40 μ /0.35 μ) \rightarrow 1,1 GHz

In order to reach 5GHz \rightarrow 2 x finger $W=20\mu$



Active Devices

- EI MOSFET. TRADE-OFFs

$$\uparrow f_T = \frac{1}{2\pi} \frac{\sqrt{2 \cdot I_D \cdot \frac{K_{PP}}{n} \cdot \frac{W}{L}}}{W \cdot L \cdot C_{ox} \left(\frac{2}{3} + \frac{C_{GSO}}{LC_{ox}} \right)}$$

L minimum

$$\uparrow f_c = \frac{0.124}{R_{sh} C_{ox} W^2}$$

Reduced W

Gm ↓
Then how is gain obtained?

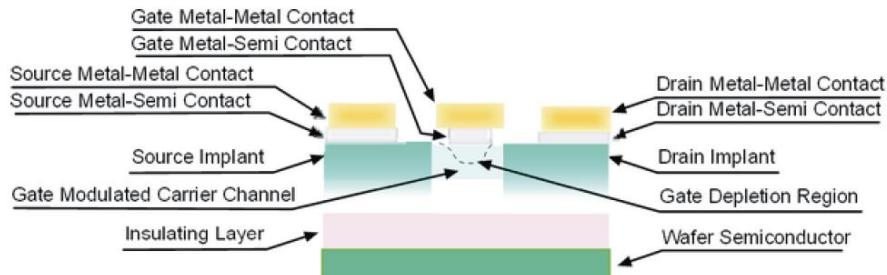
Id ↑
Higher Power consumption

V_{GS} ↑
V_{DS} ↑

**Watch Out
Saturation
Velocity**

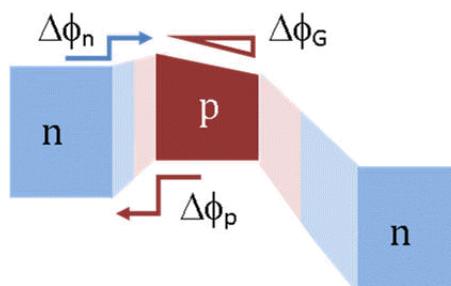
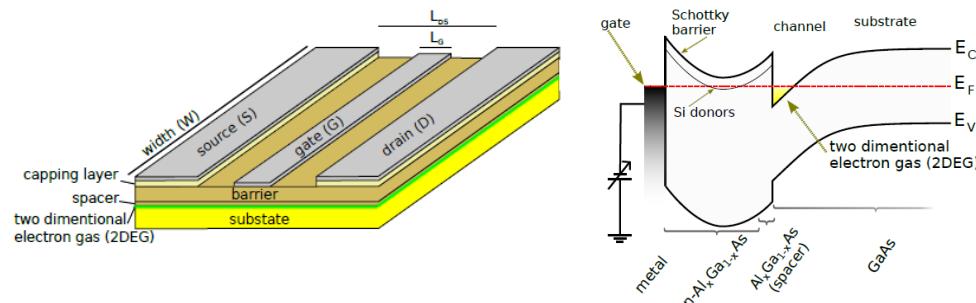
Active Devices

- HF specific transistors. HBT, MESFET y HMET



MESFET: MOSFET without gate isolation.

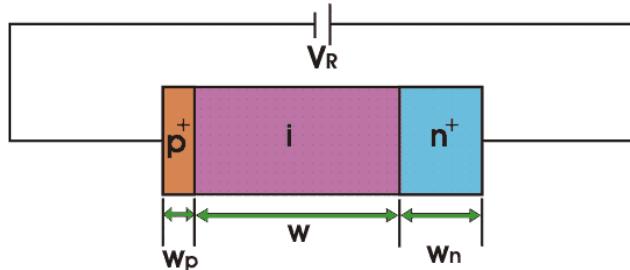
Gate conduction must be avoided, so certain bias voltages are not allowed
(JFET with a Metal-SemiCond junction)



HBT: Bipolar Transistor with different semiconductors and doping profiles that help majority carrier mobility reducing base resistance

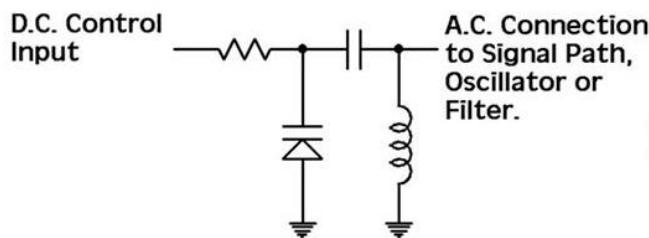
Active Devices

- PIN diodes



- Diode with an intrinsic semiconductor in the middle of the junction.
- Works when intrinsic zone is flooded.
- The charge density creates a BIAS controled equivalent resistance (the response is very linear)
- It can be used as a switch or as a variable attenuator

- Varactors

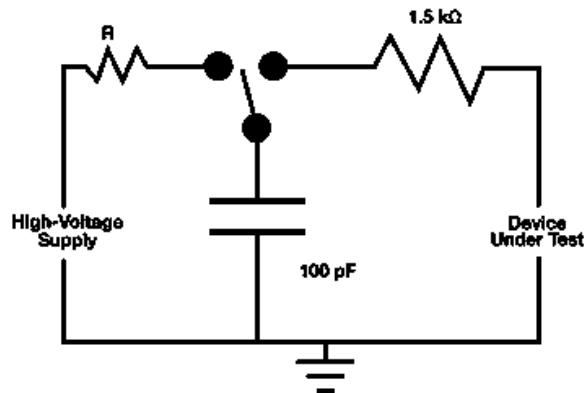


- Inverse biased diode.
- Bias voltage modulates the width of the depletion zone. A lower voltage creates a swallower depletion zone thus increasing the equivalent capacitance
- Show a high dependence with temperature
- Q parameter is not constant along the C range.
Usually is associated to another fixed value capacitor of higher quality.

ESD Protections for RF IC

- Human Body Model

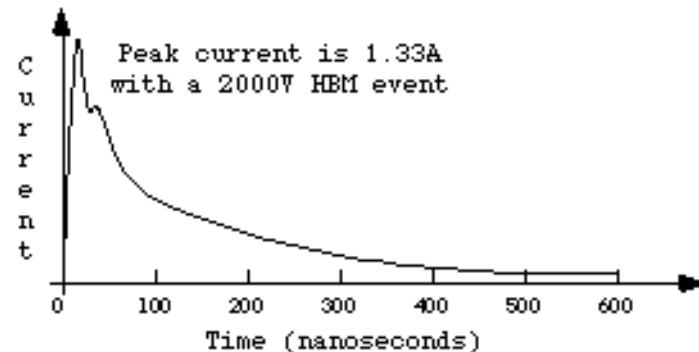
Electrostatic discharge event generated by a typical human body electrically charged when a device is touched



CLASS 1 HV < 2kV
CLASS 2 2kV < HV < 4kV
CLASS 3 HV > 4kV



ESD Protective Symbol

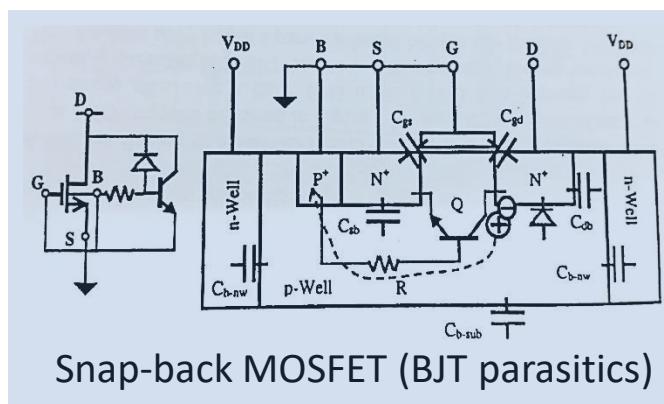
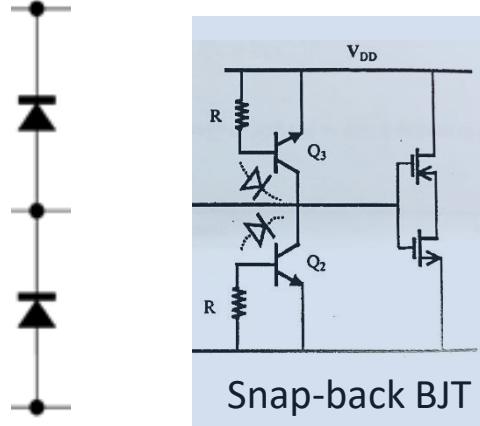
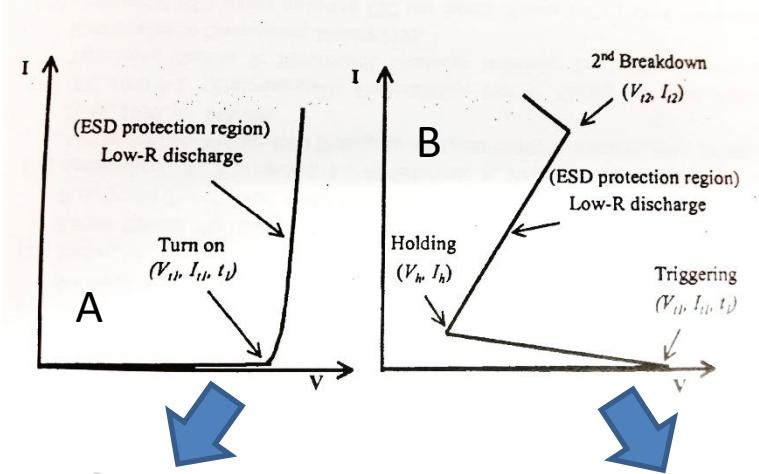


There are other models related to different events such as:

- Machine Model
- Charged Device Model

ESD Protections for RF IC

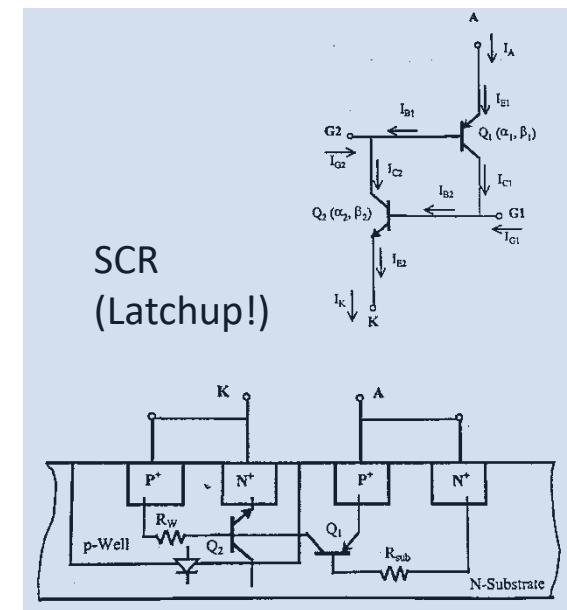
- ESD protection in integrated circuits



Diodos (Clamp)

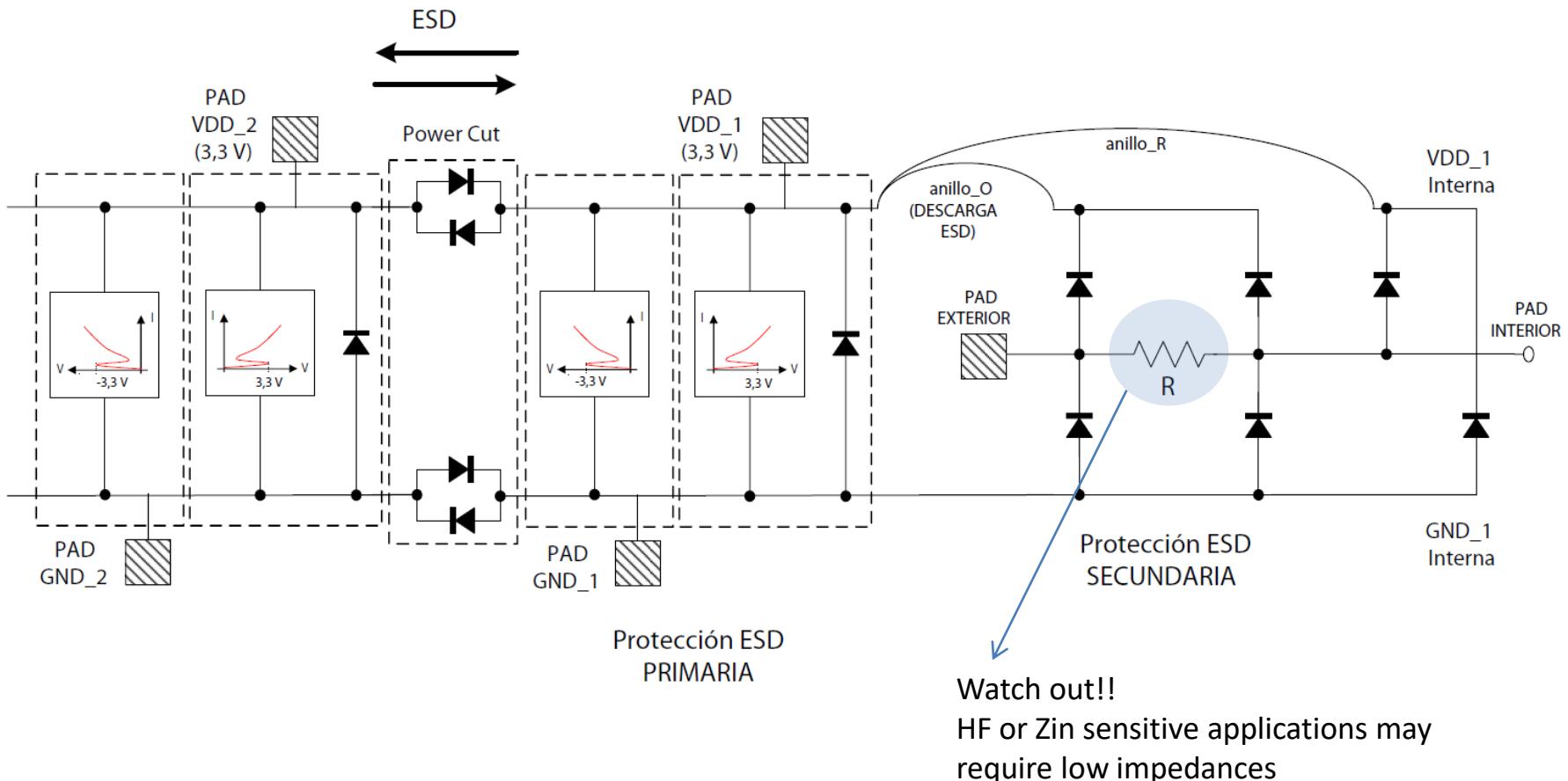
The basic idea is to use special components that allow high current discharges:

- High current conduction for a fixed voltage (A)
- A trigger voltage generates a high current conduction with a lower voltage drop (reduces dielectric stress < $V_{breakdown}$ (B))

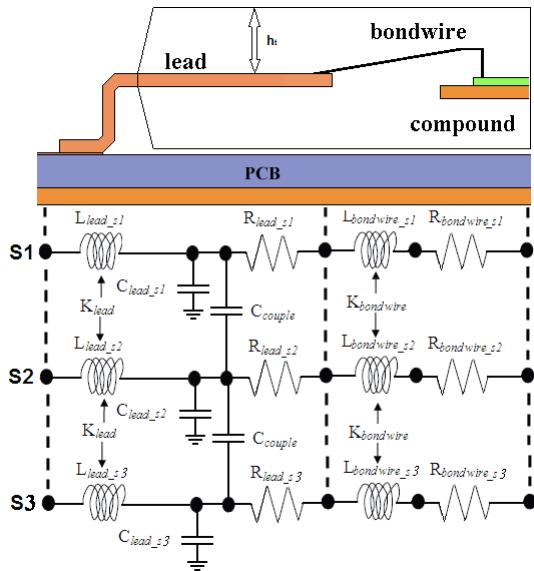


ESD Protections for RF IC

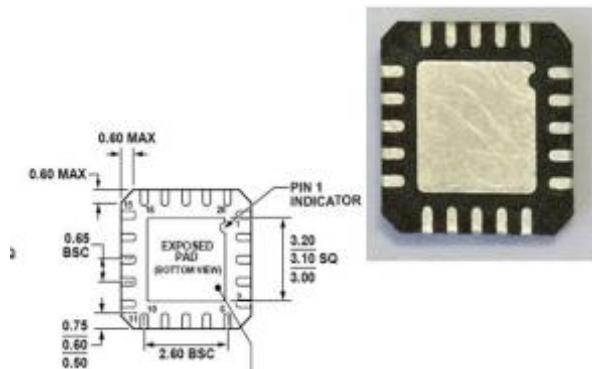
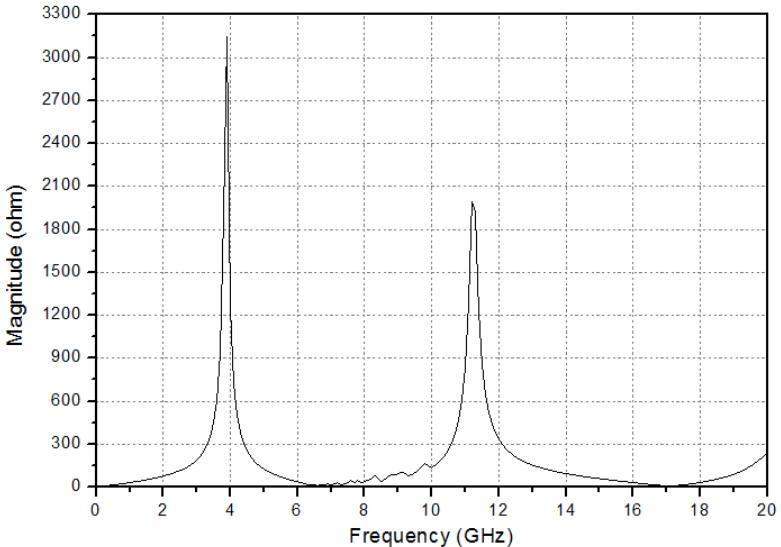
- ESD protection basic scheme



Packages for RF IC



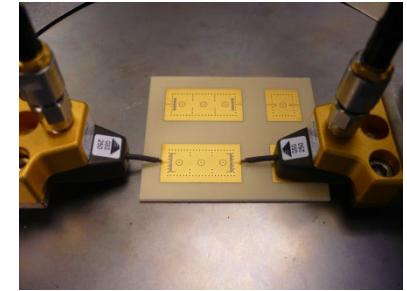
QFP-80



QFN: Reduced size, no pin structure.
Lower parasitics
Frequencies up to 20 GHz



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Chapter 6 – Power Amplifiers

Circuitos Electrónicos de Alta Frecuencia
Máster Universitario en Ingeniería de Telecomunicación (MUIT)

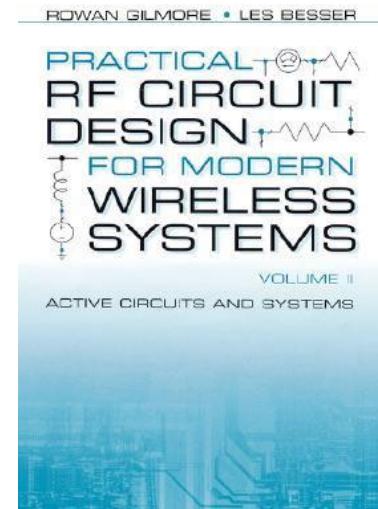
Bibliography

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RF Power Amplifiers for Wireless Communications

Second Edition

Steve C. Cripps



Practical RF for modern Wireless systems (Volume II). Rowan Gilmore, Les Besser. Artech house publishers

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- 
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 - Power levels
 - IPN and 1dB CP
 - Linear and non-linear amplifiers
 2. Optimum impedance to maximize the output power
 - Conjugate match technique
 - Loadline match
 - Conjugate vs Loadline matching
 3. Efficiency
 4. Intermodulation Distortion
 5. Thermal Analysis
 6. Categories of Amplifiers
 - Different classes: A, B, AB, C...
 - Comparison
 7. Design procedure
 - Stability
 - Load Pull contours



■ Introduction



Fundamentals

5/2

■ Power Levels

- **Power Amplifier (PA):** Amplifier intended to provide maximum power at the output for a given active element.
- High output power levels are involved:
 - Voice/data mobile communication systems: 100mW up to 0.5W
 - RADAR or Radio communications: 10-100W
- This high signal level in high frequency has a lot of implications at circuit and system levels

Note:

$$P(\text{dBm}) = 10 \log(P(\text{mW})/1\text{mW})$$

$$30 \text{ dBm} = 1\text{W} = 1000\text{mW}$$

$$20 \text{ dBm} = 0.1\text{W} = 100\text{mW}$$

$$10\text{dBm} = 10\text{mW}$$

$$0 \text{ dBm} = 1\text{mW}$$

Freq.	S21 (Gain)	S12 (Feedback)	Comments
Low	High	Small	To achieve a gain value without loosing stability is possible
High	Decreases	Increases	To achieve high gain values and to have a stable system is rather difficult

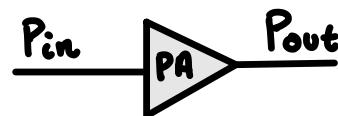
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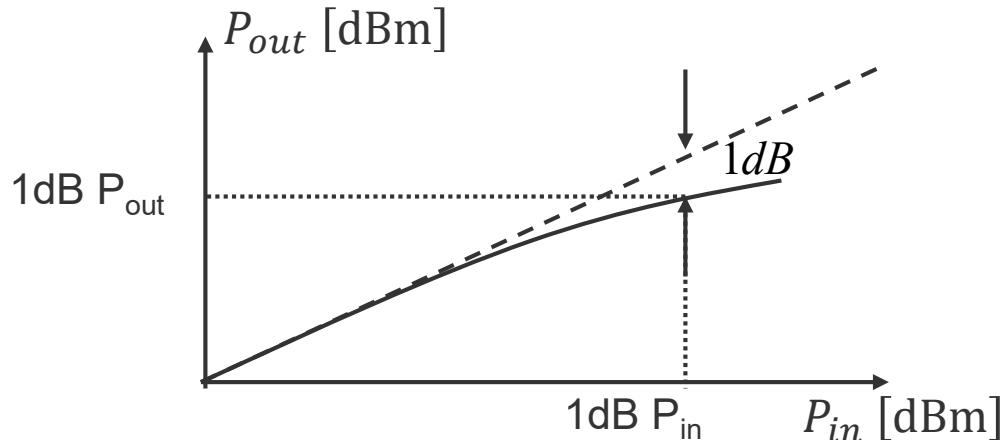
■ N-order Interception Point (IPN) and 1dB Compression Point (1dB CP)

- In general, power amplifiers operate with high-level input signals close to the 1dB Compression Point or the third order Interception Point (3IP), thus making the transistors to operate with a non-linear behavior.

1dB-Compression Point (1dB CP)



- The input 1dB compression point is the input power at which output power drops by 1dB from the predicted output power based on small signal power gain.



Dynamic linear range (Linear DR) is the input power range between the minimum detectable signal (noise floor) and maximum signal level which keeps the device in the linear regime, that is, 1dB-CP

Introduction

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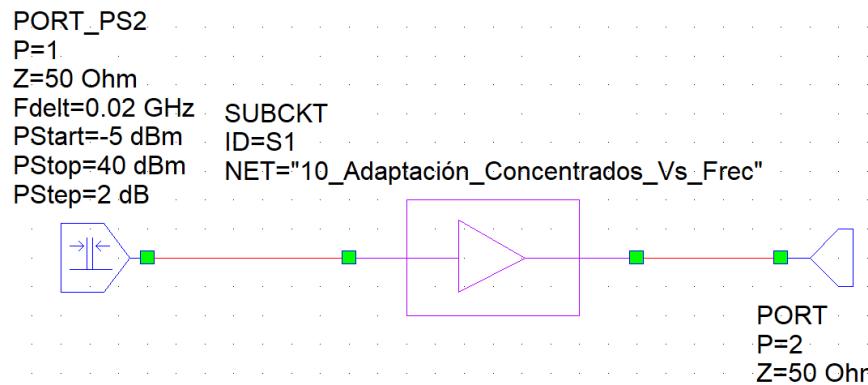
■ N-order Interception Point (IPN) and 1dB Compression Point (1dB CP)

N-order Interception Point (IPN)

- To obtain the interception point of an amplifier, two tones with the same level and very close frequencies ω_{in1} and ω_{in2} are introduced in the amplifier. Both tones will appear at the output with a similar gain (as they are very close in frequency), and other tones, resulting from the non-linear behavior of the amplifier, will also appear in the same frequency slot as the signal of interest with a slightly different gain.

Third order: $2\omega_{in1} - \omega_{in2}$, $2\omega_{in2} - \omega_{in1}$

Fifth order: $3\omega_{in1} - 2\omega_{in2}$, $3\omega_{in2} - 2\omega_{in1}$



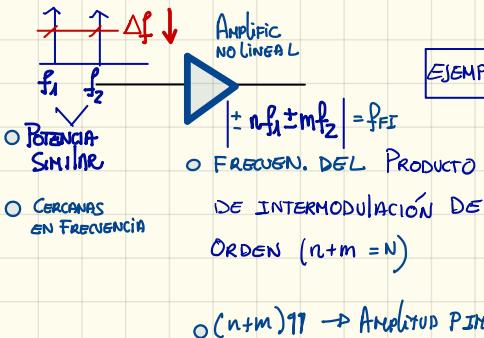
- N-order Interception Point (IPN) and 1dB Compression Point (1dB CP)

Los PI son señales que aparecen en sistemas no lineales cuando reciben en su entrada señales de diferente frecuencia. La frecuencia de los PI es una combinación lineal de las freq. de las señales de entrada.

$$f_{PI} = |\pm m f_1 \pm n f_2| \text{ ORDER } N = m+n$$

PUNTO DE INTERCEPCIÓN DE ORDEN N

IPN



<u>ORDEN 3</u> $2f_1 - f_2 \approx f_1, f_2$ $2f_2 - f_1 \approx f_1, f_2$
<u>ORDEN 5</u> $3f_1 - 2f_2 \approx f_1, f_2$ $3f_2 - 2f_1 \approx f_1, f_2$

¿QUÉ CUANTIFICA EL PUNTO DE INTERCEPCIÓN? LA PIN A LA QUE LA POUT DE LOS PRODUCTOS DE INTERMODULACIÓN GENERADOS POR EL AMPLIFICADOR ES COMPARABLE A LA POUT DE LA SEÑAL A AMPLIFICAR.

Introduction

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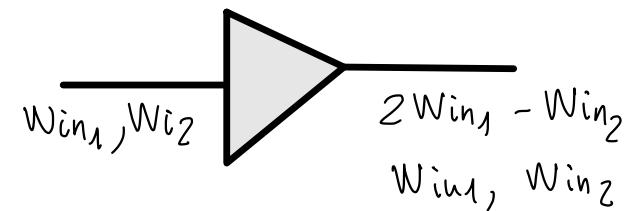
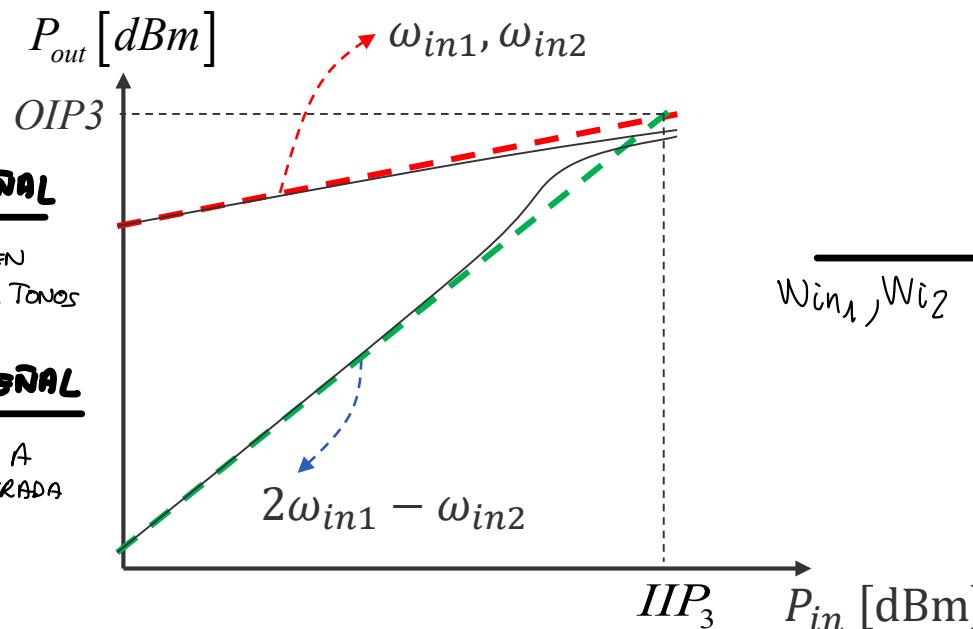
- N-order Interception Point (IPN) and 1dB Compression Point (1dB CP)
- Third-Order Input Intercept Point (IIP3) is defined as the input signal level at which the extrapolated output signal levels at ω_{in1} or ω_{in2} and $2\omega_{in1} - \omega_{in2}$ or $2\omega_{in2} - \omega_{in1}$ equal each other. The corresponding output power level is referred to as the Third-Order Output Intercept Point (OIP3).

EN LA ZONA DE PEQ SEÑAL

P_{OUT} DEL IIP3 ES DESPRECiable EN COMPARACIÓN CON LA P_{OUT} DE LOS TONOS DE ENTRADA

EN LA ZONA DE GRAN SEÑAL

P_{OUT} DEL IIP3 ES COMPARABLE A P_{OUT} DE LOS TONOS DE ENTRADA



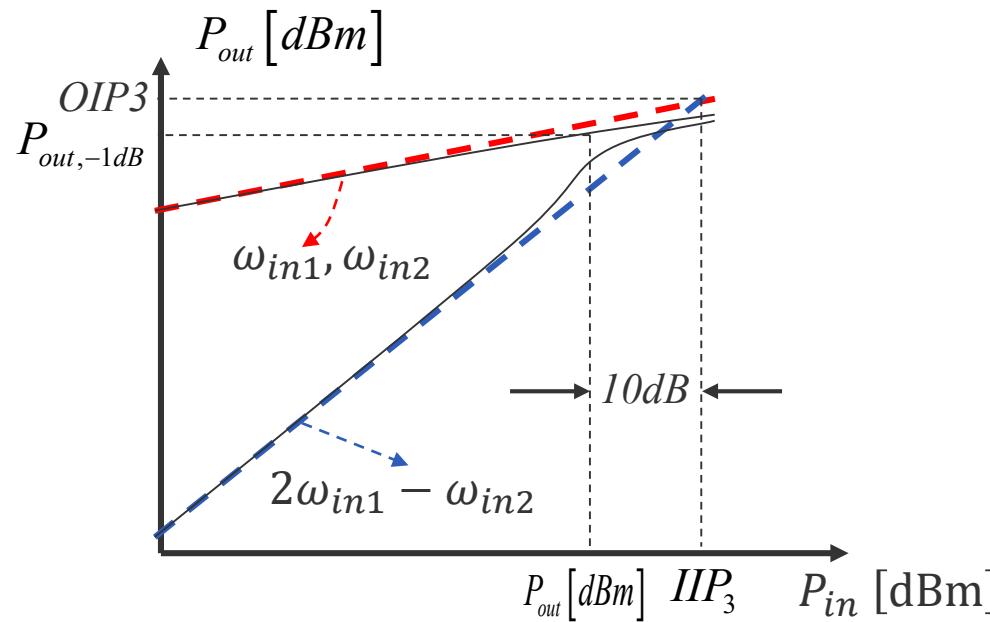
Introduction

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- N-order Interception Point (IPN) and 1dB Compression Point (1dB CP)

Relationship between 1dB CP and IIP3

- Both characteristic input power signals differ in $\approx 10\text{dB}$



- Linear regime: The device behavior does not change although the magnitude and phase of the input signal does.
- Non-linear regime: The device behavior changes with amplitude and phase of the input signal since the input and output amplifier impedances change.
- Quasi-linear regime: Approximation made to operate in linear regime although the operating regime is non-linear

- RÉGIMEN LINEAL
 - PEQUEÑA SEÑAL \rightarrow PIN BAJA
 - POR DEBAJO DEL $1dB-PC$
 - EL TRT PUEDE MODELARSE CON LOS PARÁMETROS S
 - LAS IMPEDANCIAS DEL TRT NO CAMBIAN CON EL NIVEL DE LAS SEÑALES DE ENTRADA \rightarrow EL COMPORTAMIENTO DEL CIRCUITO NO DEPENDE DEL NIVEL DE PIN

- RÉGIMEN NO LINEAL
 - ENTRE $1dB-PC$ Y $IP3 +$ CERCANO A ESTE ÚLTIMO)
 - PIN ALTA \rightarrow GRAN SEÑAL
 - NO SE PUEDE MODELAR EL TRT CON PARÁMETROS S
 - Z_{IN}^{TRT} Y Z_{OUT}^{TRT} DEPENDEN DEL NIVEL DE PIN \rightarrow POR TANTO, TAMBIÉN SU COMPORTAMIENTO (COMPLEJIDAD DE DISEÑO)
 - MAYOR DISTORSIÓN

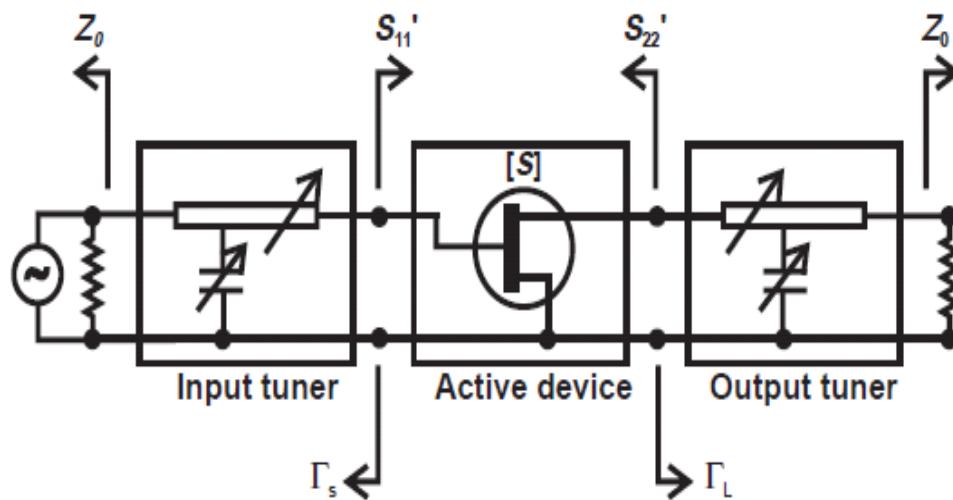
- RÉGIMEN QUASI LINEAL
 - CERCA $1dB-PC$ (Por Dibujos)
 - YA NO PUEDE CONSIDERARSE QUE EL TRT ESTÉ EN LA REGIÓN LINEAL pq SE OBSERVAN EFECTOS NO LINEALES, PERO NO HA ENTRADO EN LA REGIÓN FUERTEMENTE NO LINEAL
 - PUEDEN UTILIZARSE LOS PARÁMETROS S PARA MODELAR EL TRT, SIENDO CONSCIENTES QVÉ SE ESTÁ COMETIENDO UN ERROR

Introduction

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■ Linear and non-linear Amplifiers

- **Transistor:** 2-port network that can be represented by its S-parameters assuming linear regime
- Input and output impedance levels (or reflection coefficient) can be adjusted using passive networks

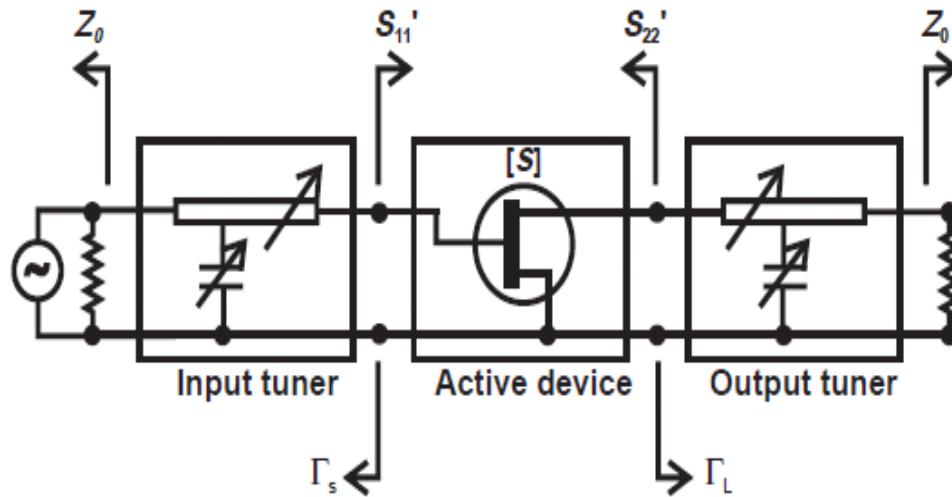


$$0 < |\Gamma_{s,L}| < 1$$

Introduction

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■ Linear amplifier



$$s'_{11} = s_{11} + \frac{s_{21} \cdot s_{12} \cdot \Gamma_L}{1 - s_{22} \cdot \Gamma_L}$$

$$s'_{22} = s_{22} + \frac{s_{21} \cdot s_{12} \cdot \Gamma_S}{1 - s_{11} \cdot \Gamma_S}$$

- In general, a solution can be found satisfying:

$$\Gamma_S = \overline{S'_{11}} \quad \Gamma_L = \overline{S'_{22}}$$

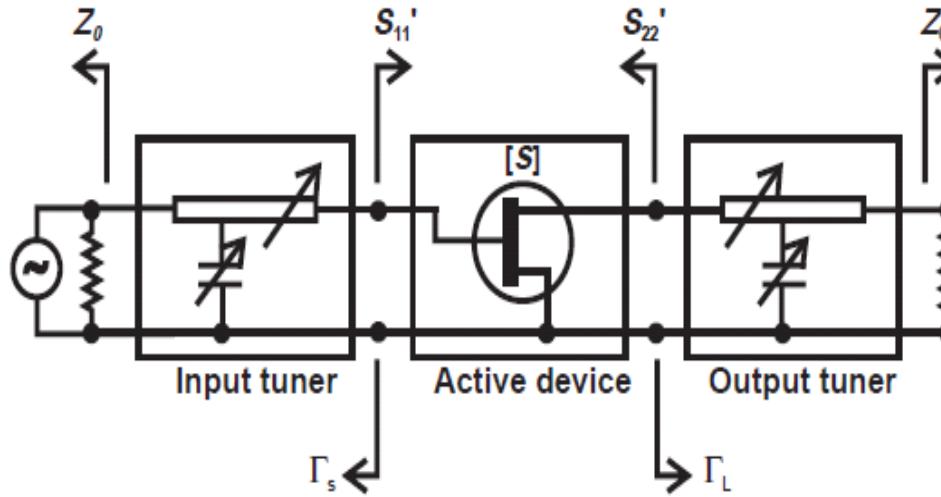
Complex Conjugate Match

Introduction

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□ Other types of amplifiers

For example, the problem of obtaining low noise performance and also maximum dynamic range from an amplifier used in the front end of a receiver could be considered to be a PA design problem, and the techniques described in Chapter 2 would be applicable. But in practice, by a "PA," we really mean an amplifier that can do damage to something.



OJO!! Un LNA también puede ser diseñado para dar el máximo margen dinámico y hay que aplicar técnicas de PA en la salida y de bajo ruido en la entrada (ver nota del Cripps que aparece arriba)

Amplifier Type	Input network design	Output network design
Linear amplifier	$MTP \Rightarrow \Gamma_S = \overline{S'_{11}}$	$MTP \Rightarrow \Gamma_L = \overline{S'_{22}}$
Low Noise	$Minimum\ NF \Rightarrow \Gamma_S = \Gamma_{S\ min\ NF}$	$MTP \Rightarrow \Gamma_L = \overline{S'_{22}}$
Power Amplifier	$MTP \Rightarrow \Gamma_S = \overline{S'_{11}}$	Max. output power $\Rightarrow \Gamma_L = \Gamma_{L\ max\ power}$

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□ Power Amplifiers design parameters

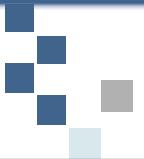
Power Amplifiers Specifications:

- Maximum output power
- Efficiency
- Intermodulation Distortion
- Thermal Effects

Amplifier Type	Input network design	Output network design
Linear amplifier	$MTP \Rightarrow \Gamma_S = \overline{S'_{11}}$	$MTP \Rightarrow \Gamma_L = \overline{S'_{22}}$
Low Noise	Minimum NF $\Rightarrow \Gamma_S = \Gamma_{S \min NF}$	$MTP \Rightarrow \Gamma_L = \overline{S'_{22}}$
Power Amplifier	$MTP \Rightarrow \Gamma_S = \overline{S'_{11}}$	Max. output power $\Rightarrow \Gamma_L = \Gamma_{L \max power}$

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 - Load Pull contours



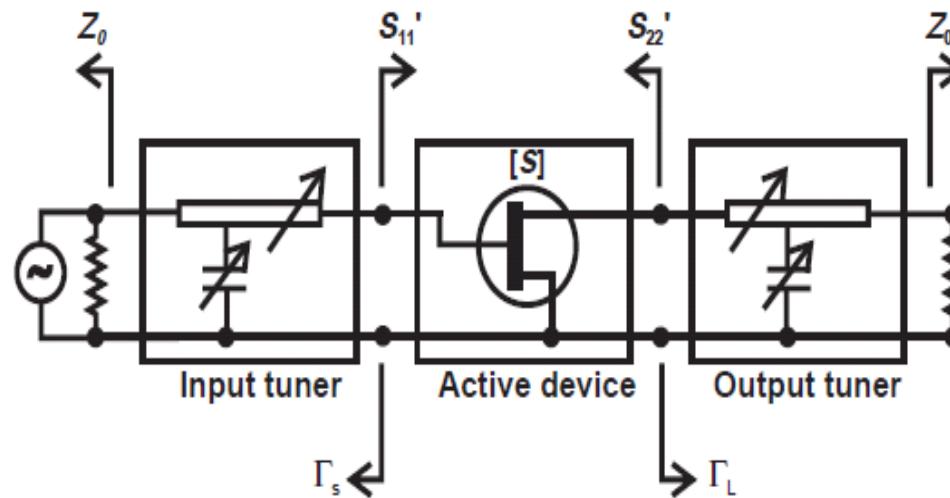
- Optimum impedance to maximize output power

Optimum impedance for maximum output power

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■ Conjugate Match

- A generator delivers maximum power into an external load when:
 - Load resistance is equal to the real part of the generator impedance
 - Reactive parts have been resonated out



$$\Gamma_S = \overline{S'_{11}}$$

$$\Gamma_L = \overline{S'_{22}}$$

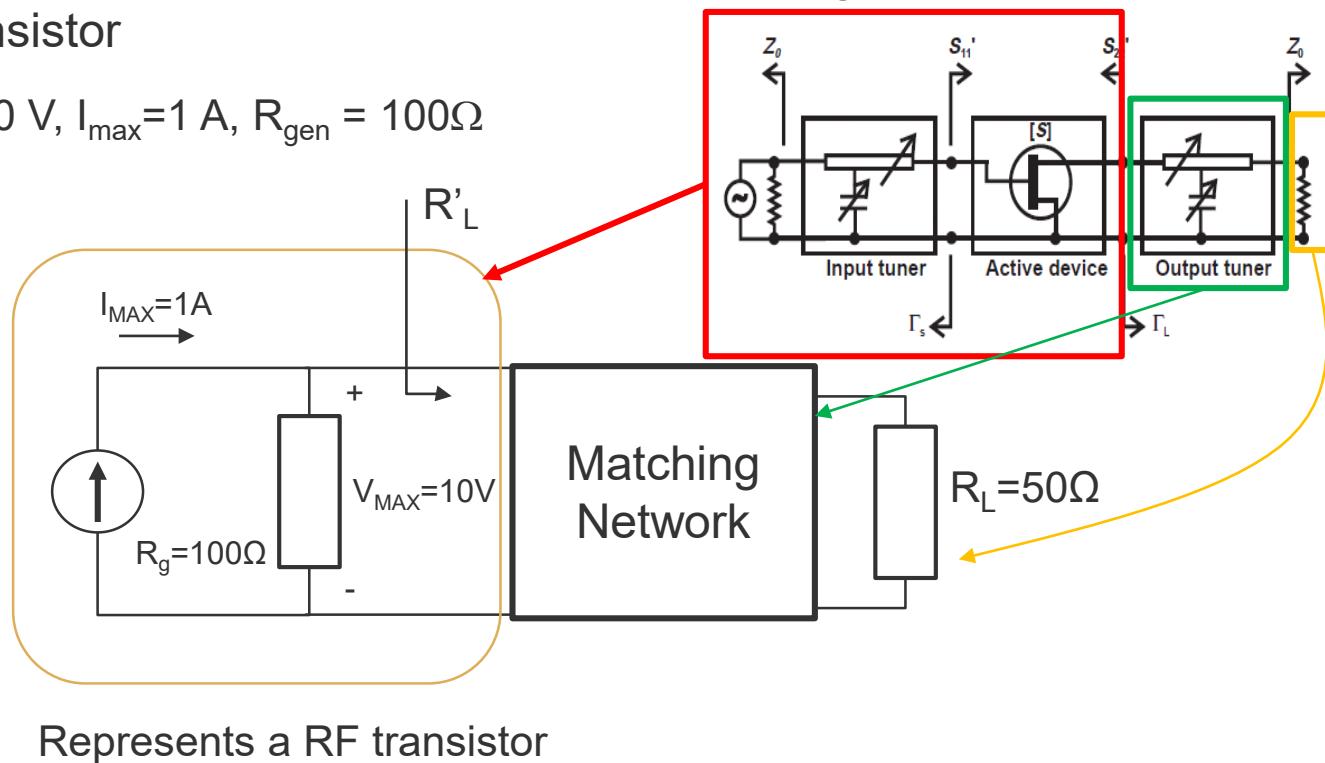
Optimum impedance for maximum output power

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■ Loadline Match

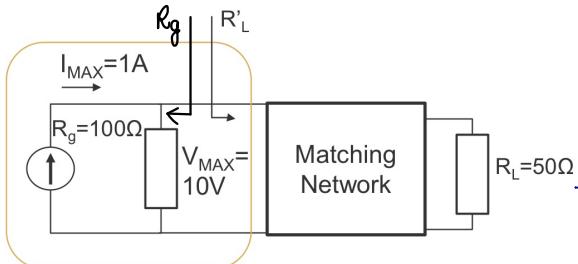
- The *loadline match* is a fundamental trade-off for obtaining the maximum power from a particular RF transistor

Example: $V_{\max} = 10 \text{ V}$, $I_{\max} = 1 \text{ A}$, $R_{\text{gen}} = 100\Omega$



■ Loadline Match

Example: $V_{max} = 10 \text{ V}$, $I_{max} = 1 \text{ A}$, $R_{gen} = 100\Omega$



Represents a RF transistor

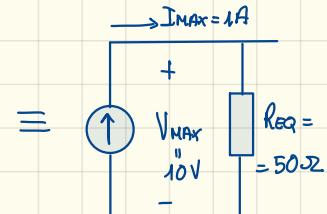
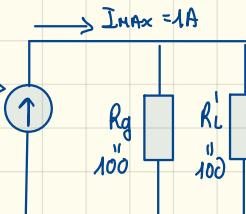
¿QUÉ RESISTENCIA R'_L DEBE "VERSE" DESDE LA SALIDA DEL TRT EN DIRECCIÓN A LA CARGA PARA CONSEGUIR QUE EL TRT LOGRE LA MÁXIMA POTENCIA QUE (DENTRO DE SUS LIMITACIONES) PUEDE GENERAR EL TRT?

$$P_{max}^{TRT} = I_{max} \times V_{max} = 1A \times 10V = 10W$$

DISEÑO PARA MTP

$$R'_L = \overline{R_g} = 100\Omega$$

CIRCUITO EQUIVALENTE

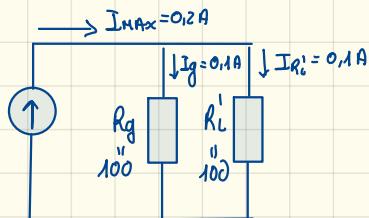


CON ESTA $R_{req} = 50\Omega$ NO ES POSIBLE EXTRAYER AL MUNDO TANTO I_{max} Y V_{max} DEL TRANSISTOR (TRT)

O SI HACEMOS QUE EL TRT GENERE V_{max} : $I_{req} = \frac{V_{max}}{R_{req}} = \frac{10}{50} = 0,2(A) < I_{max} = 1(A)$

O SI HACEMOS QUE EL TRT GENERE I_{max} : $V_{req} = I_{max} \cdot R_{req} = 50(V) > 10(V) = V_{max}$
NO PUEDE TRABAJAR CON I_{max} !

POR TANTO, CON MTP ($R_{QG} = 50\Omega$) NO SE PUEDE EXTRAER LA MÁXIMA CORRIENTE DEL TRT Y, EN CONSECUENCIA, NO SE PUEDE EXTRAER DE ÉL LA POTENCIA MÁXIMA QUE PUEDE ENTREGAR.



$$P_{Rg} = 0,1^2 \times 100 = 1W$$

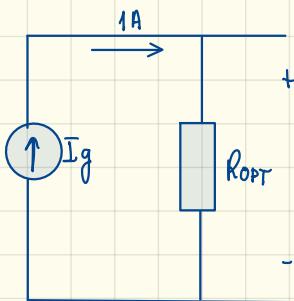
$$P_{R_L} = 0,1^2 \times 100 = 1W$$

POTENCIAS
DISTRIBUIDAS

$$\begin{aligned} P_{TRT} &= 10(V) \times 0,2A = \\ &= 2W \end{aligned}$$

POTENCIA
GENERADA

EXISTE UNA CARGA PARA LA QUE SÍ ES POSIBLE EXTRAER DEL TRANSISTOR LA MÁXIMA POTENCIA QUE ÉSTE PUEDE GENERAR. ¿CUÁL ES ESTA CARGA?



$$R_{OPT} = \frac{10}{1} = 10\Omega = \frac{R_g R_l'}{R_g + R_l'}$$

$$R_g = 100\Omega \quad 10 = \frac{100 R_l'}{100 + R_l'}$$

$$10(100 + R_l') = 100 R_l'$$

$$1000 = (100 - 10) R_l'$$

$$R_l' = \frac{1000}{90} = 11,1\Omega$$

$$R_l' = 11,1\Omega$$

PERMITE EXTRAER LA MÁXIMA POTENCIA DEL TRANSISTOR

$$P_{R_l'} = \frac{10^2}{11,1} = 9(W) \quad I_{R_l'} = \frac{10}{11,1} = 0,9(A)$$

$$P_{Rg} = \frac{10^2}{100} = 1(W) \quad I_{Rg} = \frac{10}{100} = 0,1(A)$$

$$R_l' > R_l$$

CON HATCH

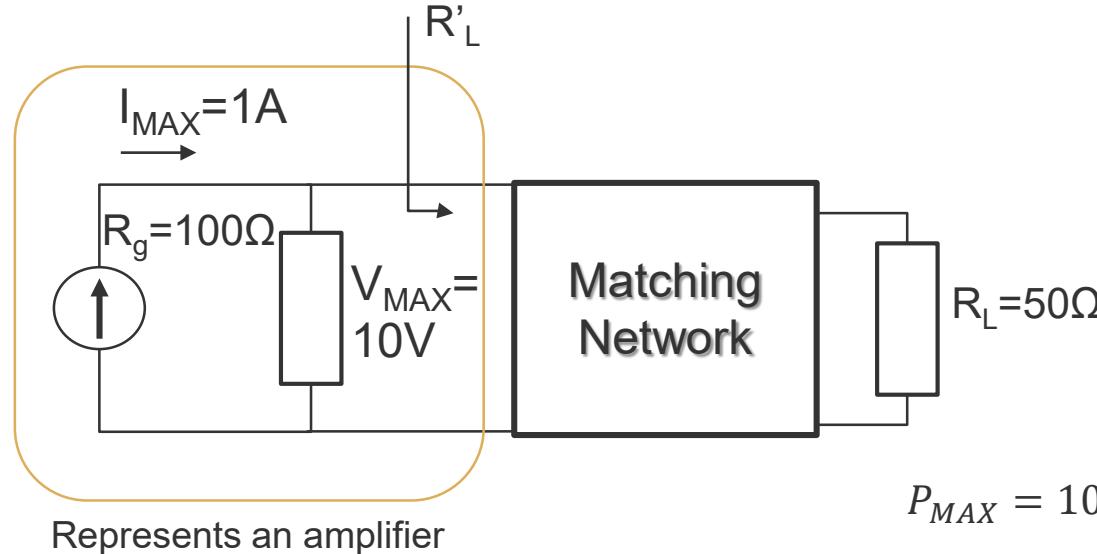
P_{MAX}

Optimum impedance for maximum output power

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■ Loadline Match

Example: $V_{max} = 10 V$, $I_{max}=1 A$, $R_{gen} = 100\Omega$



Maximum Transfer Power

$$R'_L = 100\Omega \rightarrow \begin{cases} V_{MAX} = 10V \\ I_{MAX} = 0,2A \\ P_{MAX} = 2W \end{cases}$$

Maximum output Power

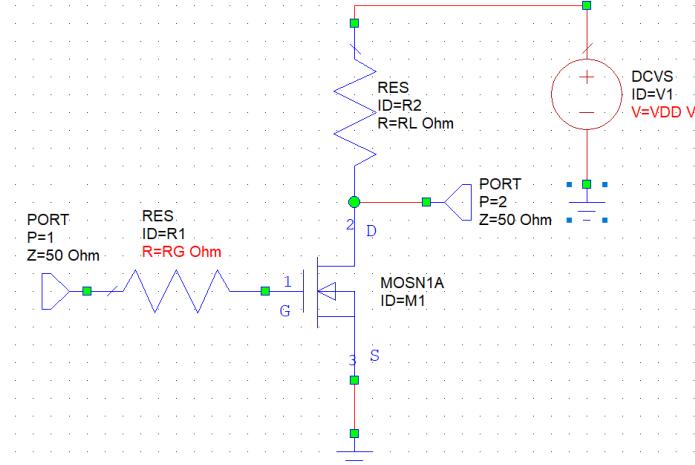
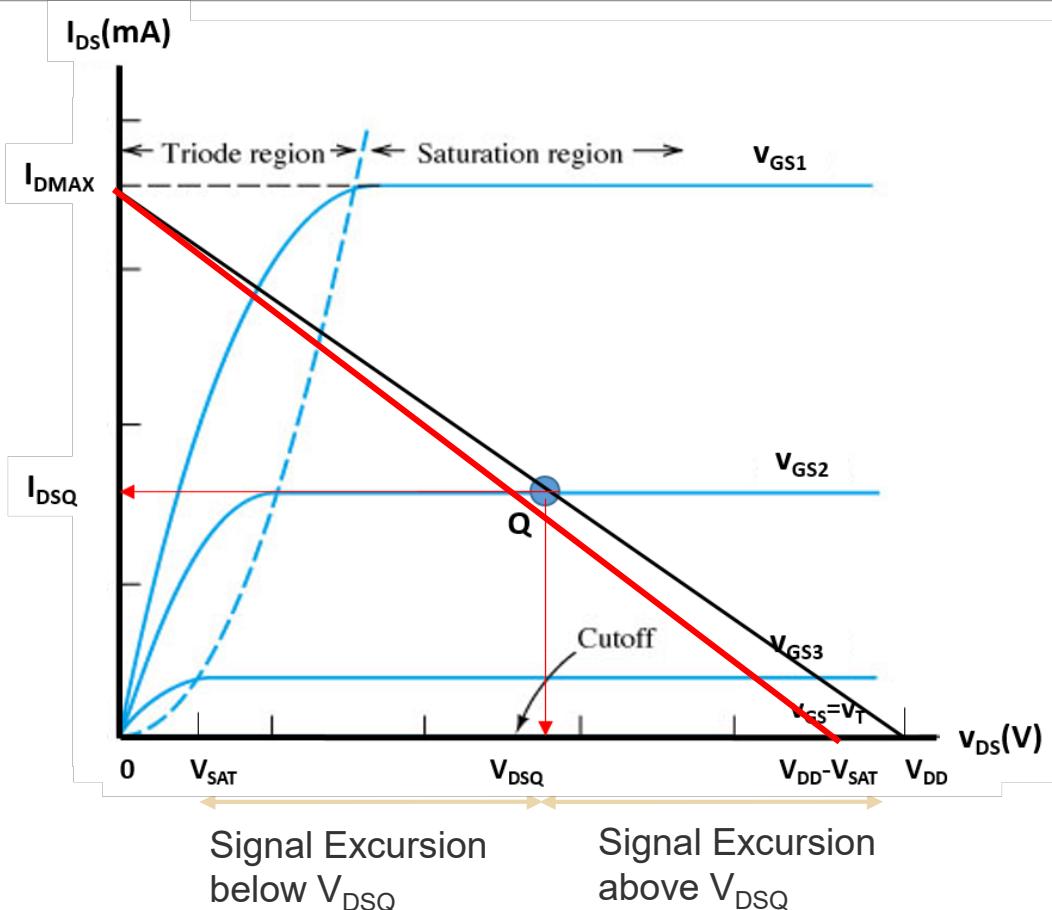
$$P_{MAX} = 10W \rightarrow R_{opt} = \frac{R_g R'_L}{R_g + R'_L} = \frac{V_{MAX}}{I_{MAX}} = 10\Omega$$
$$R_g = 100\Omega \rightarrow R'_L = 11\Omega$$

- A lower value of load resistance, than that used for the conjugate match, is needed to employ the maximum current and the maximum voltage swing

Optimum impedance for maximum output power

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Loadline Match



Circuit: Loadline

$$V_{DD} - I_D R_L - V_{DS} = 0 \rightarrow I_D = \frac{V_{DD} - V_{DS}}{R_L}$$

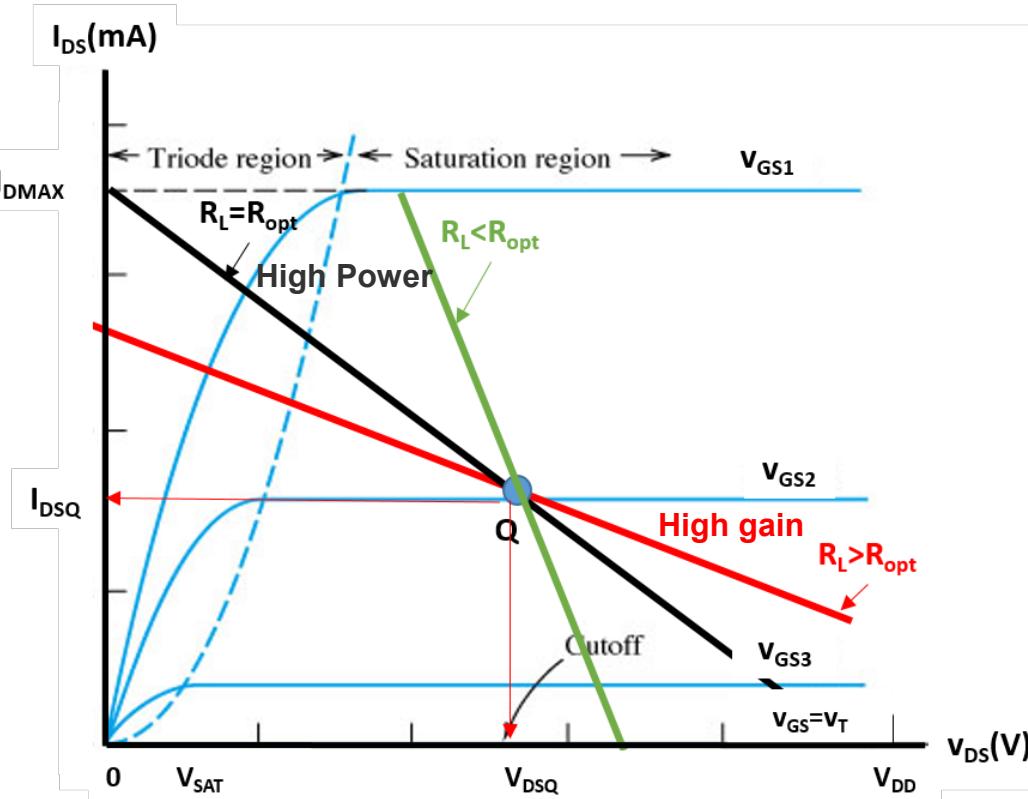
Optimum RL for symmetric signal excursion

$$R_{opt} = \frac{V_{DD} - V_{SAT}}{I_{DMAX}}$$

Optimum impedance for maximum output power

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Loadline Match

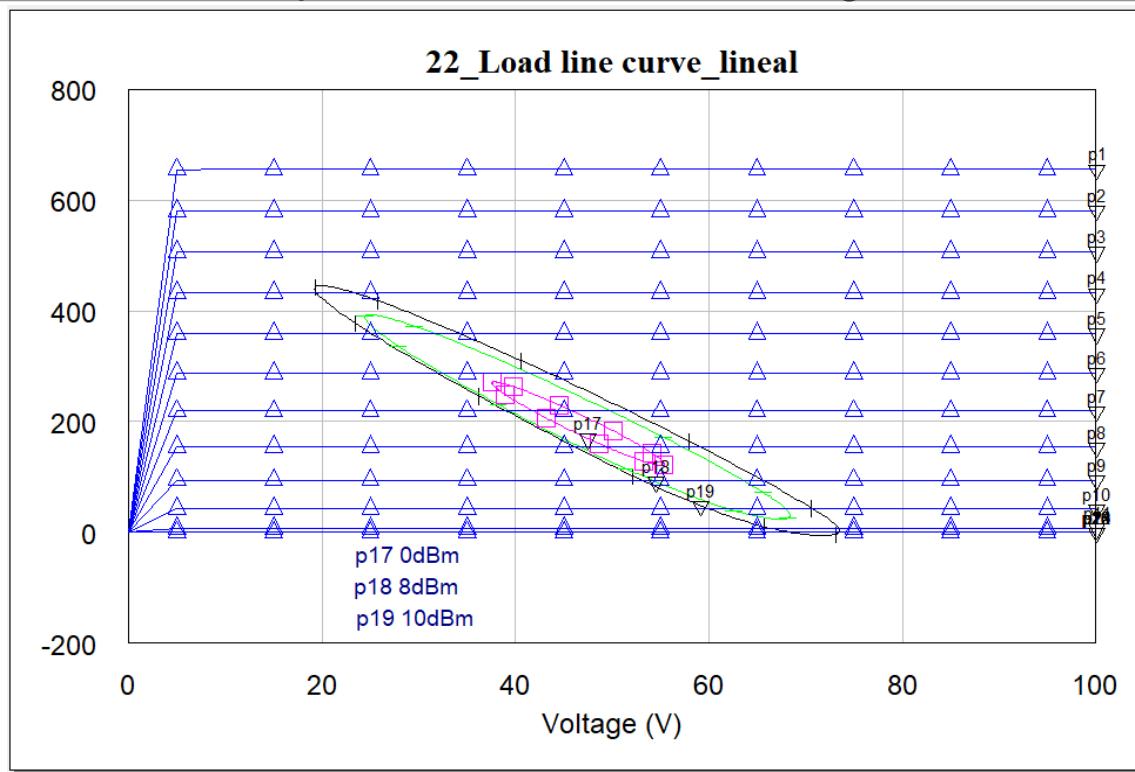


- Optimum RL (symmetric excursion)
$$R_{opt} = \frac{V_{DD} - V_{SAT}}{I_{DMAX}}$$
- $RL < R_{opt}$: Voltage limited
 $RL > R_{opt}$: Current limited
- The higher gain is provided by the red loadline: $RL > R_{opt}$
- The Q point can be moved towards cut-off to increase the Efficiency of the amplifier.

Optimum impedance for maximum output power

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- Loadline match (dynamic): With complex Load Impedance, a elliptic shape appears due to the phase shift between voltage and current

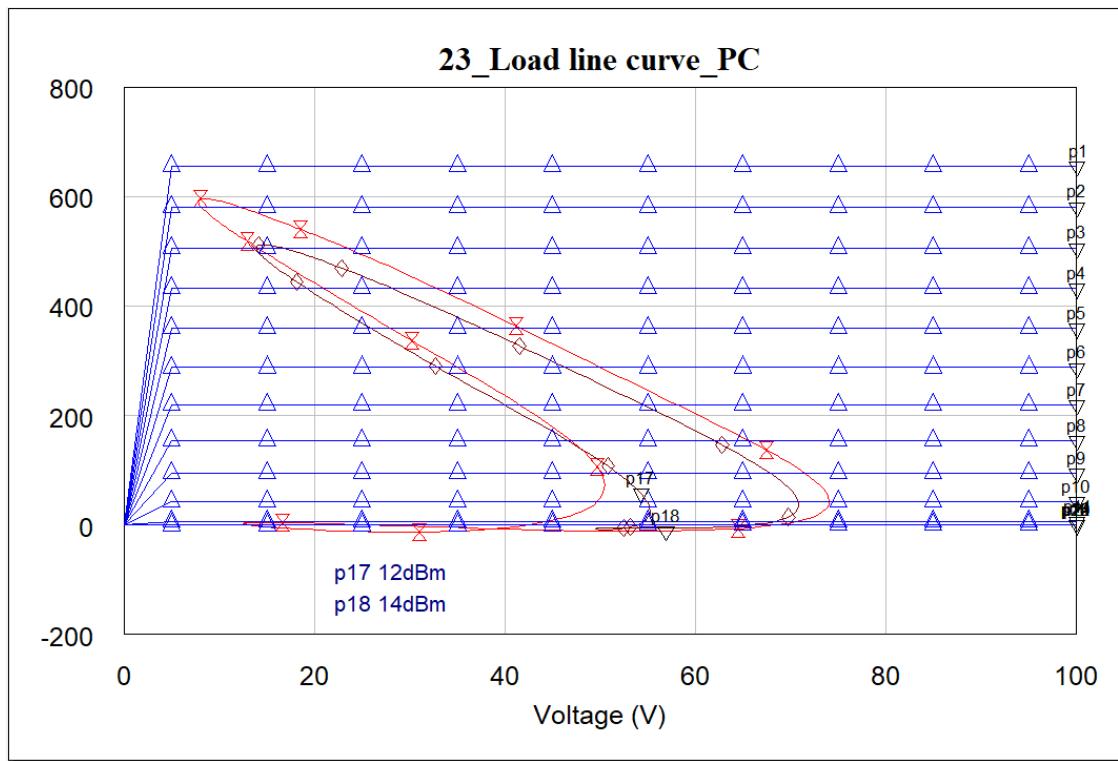


Dynamic movement of the operation point in the active device with a complex impedance (frequency dependent) and different input powers

Optimum impedance for maximum output power

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- Loadline Match (dynamic): Complex Load Impedance and Non-linear effects

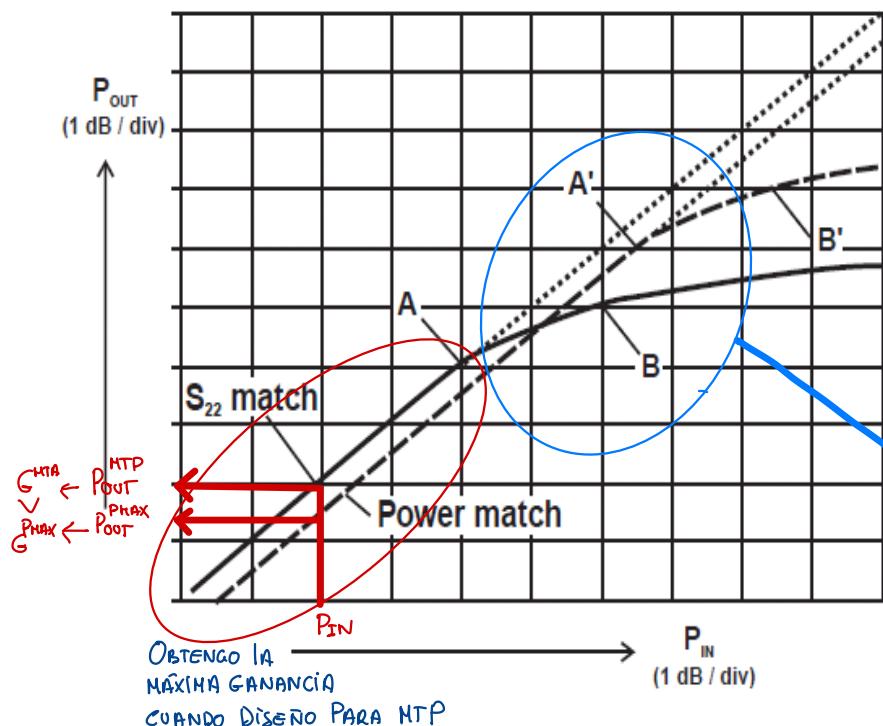


Optimum impedance for maximum output power

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Conjugate and Loadline

- Power transfer function for conjugate (*solid*) and loadline (*dashed*) matching conditions



- 1dB compression point is lower for the conjugate matched device
- For stronger driving signals the power match can guarantee 2 – 4 dB of increased output power

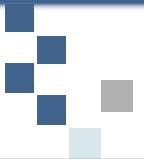
Es + EFICIENTE EL DISEÑO Realizado PARA MAXIMIZAR LA P_{OUT} Porque EL PC SE PRODUCE DESPUES. ESTO ES Porque EL DISEÑO ESTÁ PREPARADO PARA QUE EL TRT FUNCIONE en la Rección no lineal

EN PEQUEÑA SEÑAL, CUANDO DISEÑO PARA MTP OBTENGO
MÁS GANANCIA QUE SI DISEÑO PARA OPTIMIZAR LA
POTENCIA MÁXIMA QUE PUEDE PROPORCIONAR EL DISPOSITIVO
ACTIVO.

EN CAMBIO, A MEDIDA QUE AUMENTA EL NIVEL DE LA SEÑAL
DE ENTRADA, EL DISEÑO QUE MAXIMIZA LA POTENCIA DE
SALIDA ES + EFICIENTE QUE EL DE MTP.

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■ Efficiency

Thermal Analysis

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■ Data sheet

**polyfet rf devices****GP041****General Description**

Polyfet's [GAN on SiC] HEMT power transistors contain no internal matching; making them suitable for both broadband and narrow band applications. The use of a thermally enhanced package enables this device to have superior heat dissipation properties. The high drain break down voltage permits this device to operate over a wide voltage range of up to 50V.

Sustrato de elevada conductividad térmica

FACILITA LA DISTRIPACIÓN TÉRMICA EVITANDO EL AUTOCALIENTAMIENTO DEL TRANSISTOR
RF POWER GAN TRANSISTOR

10.0 Watts Single Ended

Package Style GP

HIGH EFFICIENCY, LINEAR

HIGH GAIN, LOW NOISE

ROHS COMPLIANT

Efficiency

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■ Power Added Efficiency

- Universal Definition of Efficiency:

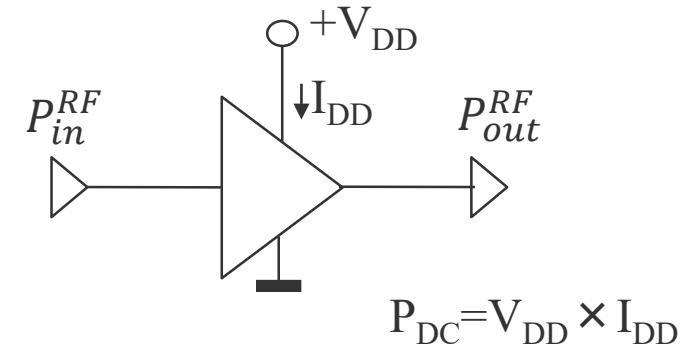
$$\eta_{TOTAL} = \frac{P_{RF,out}}{P_{DC} + P_{RF,in}}$$

- The most used definition in Radio Frequency:

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

Where $\eta = \frac{P_{RF,out}}{P_{DC}}$ and $G = \frac{P_{RF,out}}{P_{RF,in}}$

PAE: Power Added Efficiency



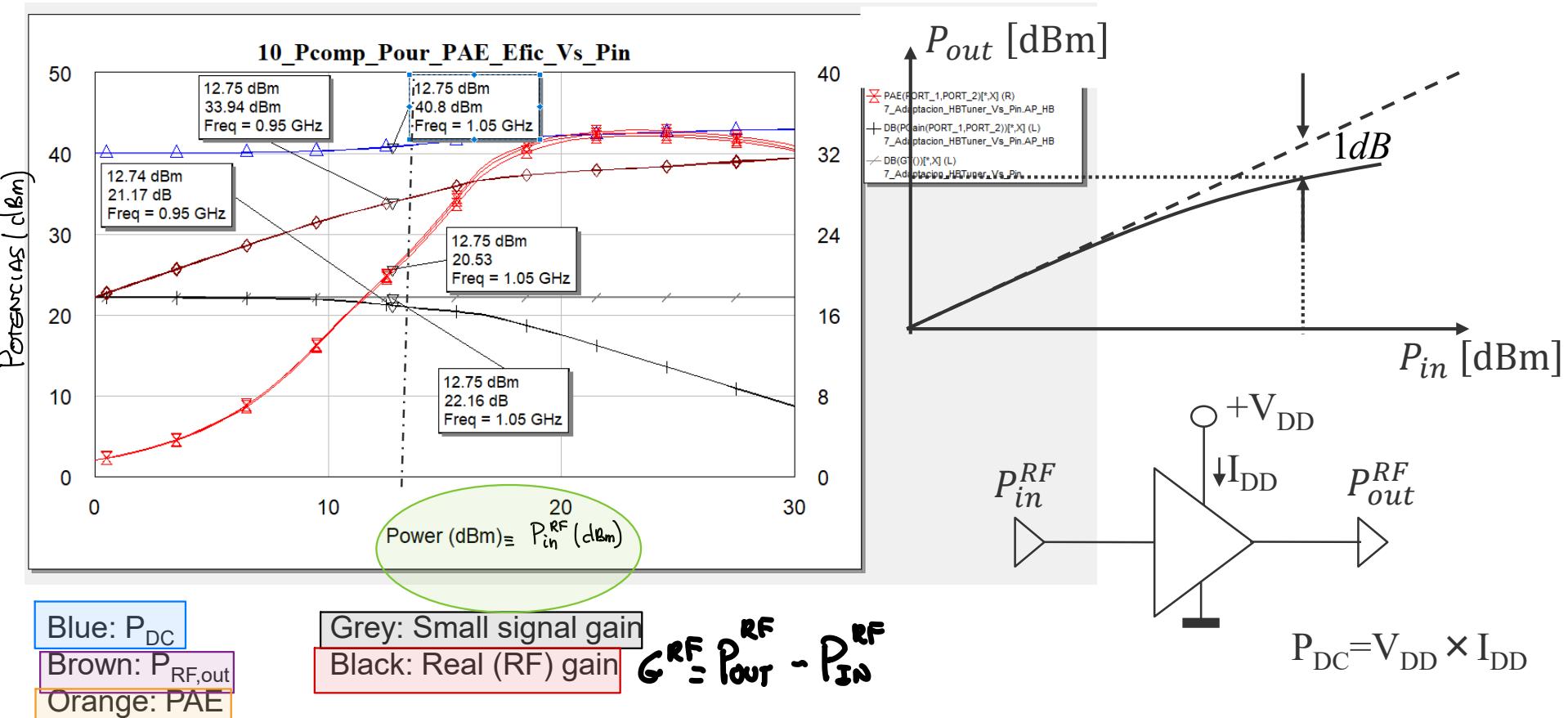
- For high gains ($P_{RF,in} \ll P_{RF,out}$):

$$\eta_{TOTAL} = \eta_{PAE} = \eta = \frac{P_{RF,out}}{P_{DC}}$$

Efficiency

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■ Power Added Efficiency



EN LA ZONA DE PEQUEÑA SEÑAL

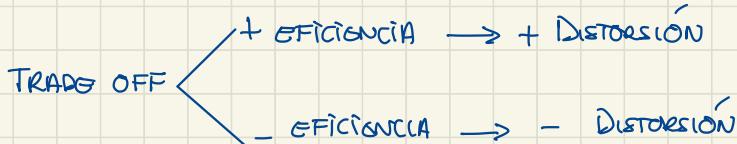
- LA RELACIÓN $P_{in} - P_{out}$ ES CONSTANTE
- P_{DC} NO CAMBIA CON P_{in}

EN LA ZONA DE GRAN SEÑAL

- LA G^{RF} SE ALEJA DE LA G EN PEQ SEÑAL, CUANDO LA DIFERENCIA ENTRE ELLAS SEA 1dB EL AMPLIFICADOR ESTARÁ EN SU PUNTO DE COMPRESIÓN
- POR OTRO LADO P_{DC} CAMBIA PORQUE LAS IMPEDANCIAS DEL TRANSISTOR CAMBIAN CON P_{in} Y EL PUNTO DE POLARIZACIÓN CAMBIA

EFICIENCIA

CLASE A (35%) NOS PIDEN DISEÑAR PARA MÁXIMA EFICIENCIA, ESTO NOS VA A OBLIGAR A TRABAJAR EN RÉGIMEN NO LINEAL Y HABRÁ DISTORSIÓN



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■ Intermodulation distortion

Intermodulation Distortion

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■ Definition

PORT_PS2

P=1

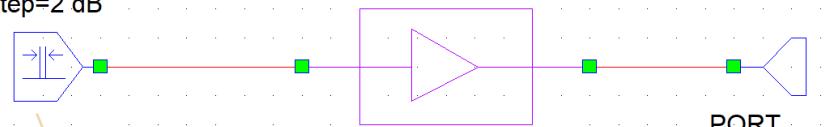
Z=50 Ohm

Fdelt=0.02 GHz SUBCKT

PStart=-5 dBm ID=S1

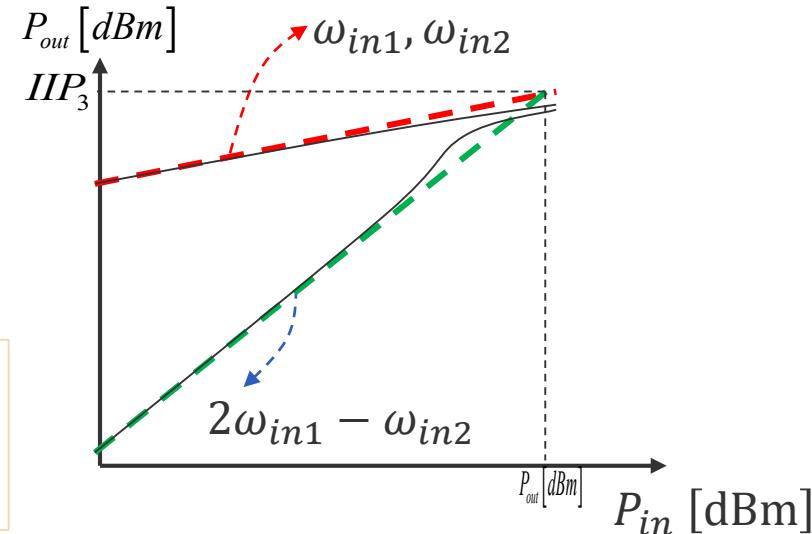
PStop=40 dBm NET="10_Adaptación_Concentrados_Vs_Frec"

PStep=2 dB



Two tones with similar level and very close frequencies ($\omega_{in1}, \omega_{in2}$) are introduced.

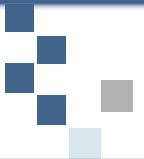
Due to the non-linear behaviour, several intermodulation products appear at the output:
Third order: $2\omega_{in1} - \omega_{in2}, 2\omega_{in2} - \omega_{in1}$
Fifth order: $3\omega_{in1} - 2\omega_{in2}, 3\omega_{in2} - 2\omega_{in1}$



The power difference (in dBs) between the desired output signal and the intermodulation product is known as Intermodulation Distortion.

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■ Thermal analysis

Thermal Analysis

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- Data sheet

**polyfet rf devices****GP041****General Description**

Polyfet's GAN [on SiC] HEMT power transistors contain no internal matching; making them suitable for both broadband and narrow band applications. The use of a thermally enhanced package enables this device to have superior heat dissipation properties. The high drain break down voltage permits this device to operate over a wide voltage range of up to 50V.

**Sustrato de elevada conductividad térmica****RF POWER GAN TRANSISTOR**

10.0 Watts Single Ended

Package Style GP

HIGH EFFICIENCY, LINEAR

HIGH GAIN, LOW NOISE

ROHS COMPLIANT

Thermal Analysis

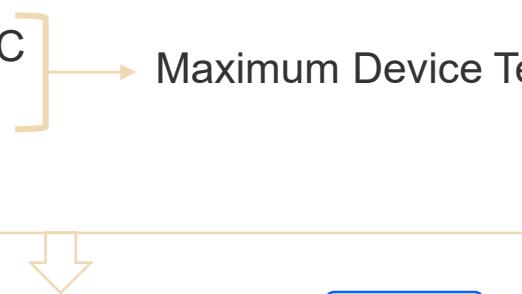
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■ Data sheet

ABSOLUTE MAXIMUM RATINGS ($T = 25^{\circ}\text{C}$)

Total Device Dissipation	Junction to Case Thermal Resistance	Maximum Junction Temperature	Storage Temperature	Drain to Source Voltage	Gate to Source Voltage
30 Watts	5.20 $^{\circ}\text{C/W}$	200 $^{\circ}\text{C}$	-65 $^{\circ}\text{C}$ to 150 $^{\circ}\text{C}$	180 V	-10 V to + 2 V

- Maximum Power (data sheet): $P_{\text{MAX}}=30\text{W}$
- Maximum Junction Temperature: $T_{\text{MAX}}=200^{\circ}\text{C}$
- Room Temperature: $T_{\text{room}}=25^{\circ}\text{C}$
- Thermal Resistance: $R_{\text{TH}}=5,20^{\circ}\text{C/W}$



$$\text{Maximum Power: } P_{\text{MAX}} = \frac{T_{\text{MAX}}}{R_{\text{TH}}} = \boxed{33,64\text{W}}$$

Safety margin

GaN PolyFET

GaN → Densidad de potenciales del orden de $\frac{W}{mm^2}$

SiC → MATERIAL DE ALTA CONDUCTIVIDAD TÉRMICA
FAVORECIENDO LA DISIPACIÓN TÉRMICA Y EVITANDO
LOS PUNTOS CALIENTES

$$\text{LEY OHM: } V = I \times R$$

$$\text{LEY DE OHM TÉRMICA: } \Delta T = P_D \times R_{TH}$$

$$R_{TH} = 5,20 \frac{^{\circ}\text{C}}{W}$$

$$T_{MAX TET} = 200^{\circ}\text{C}$$

↓

$$P_D = \frac{\Delta T}{R_{TH}} = \frac{200^{\circ} - 25^{\circ}}{5,20 \frac{^{\circ}\text{C}}{W}} = 33,64 W$$

MÁXIMA QUE PUEDE SOPORTAR, ES NECESARIO REGISTRAR LA DEL AMBIENTE

$$P_D^{\text{FAB}} = 30 W$$



ligeramente superior porque el fabricante siempre deja un margen de seguridad por si hubiera defectos en el material y se generaran puntos calientes

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■ Categories of Amplifiers

Categories of Amplifiers

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■ Introduction

■ Two main categories:

- **A, AB, B and C:** Controlled conduction angle amplifiers, which are defined by the length of their conduction state over some portion of the output waveform.
- **D, E, F, G, S, T...:** So-called “switching” amplifier classes, which use digital circuits and pulse width modulation (PWM) to constantly switch the signal driving the output hard into the transistors saturation and cut-off regions.

■ We are going to focus on A, AB, B and C.

- **Class A** are the less efficient, but have the smallest distortion, and are easier to stabilize at higher frequencies.
- **Classes B, AB and C** have higher efficiencies, but also higher distortion (they can only be used with those signals which do not carry the information in the amplitude). They are more difficult to stabilize, since the input and output impedances depend on the input signal level.

CLASE A, AB, B, C

LOS COMPARAMOS EN TÉRMINOS DE :

EFICIENCIA

DISTORSIÓN

RELACIONADOS CON EL HECHO DE QUE EL TRT TRABAJE EN LA REGIÓN DE GRAN SEÑAL (NO LINEAL)

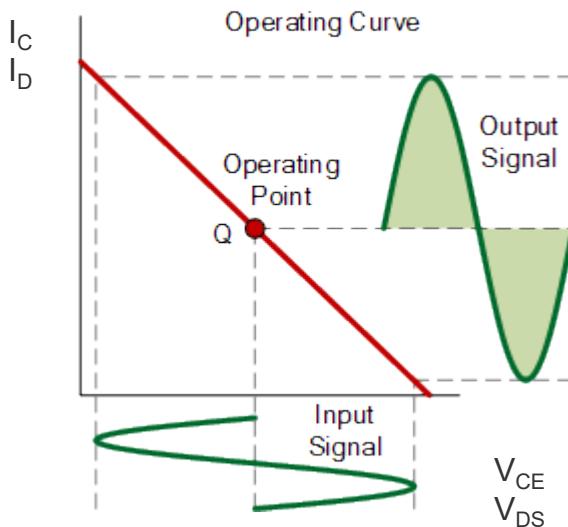
ESTABILIDAD

EN ESTA REGIÓN SE PRODUCEN CAMBIOS EN Z_{IN} Z_{OUT} DEL TRT LO QUE AFECTA AL PUNTO DE POLARIZACIÓN, ESTABILIDAD Y DISTORSIÓN

Categories of Amplifiers

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Class A



The conduction angle α indicates the proportion of the RF cycle for which conduction is happening

- Linear regime: It does not work in the strong non-linear regions. It works below, but close to, 1dB-PC. Linear models can be applied with an (small) error.
- To achieve quasi-linearity and gain, the output stage is biased “ON” (conducting) all the time (conduction angle: 360 degrees).
- As the output device is “ON” at all times, it is constantly carrying current, which represents a continuous loss of power in the amplifier.
- Due to this continuous loss of power, tremendous amounts of heat are generated adding to their very low efficiency at around 30%, making them impractical for very high-power amplifications.
- Small distortion

Categories of Amplifiers

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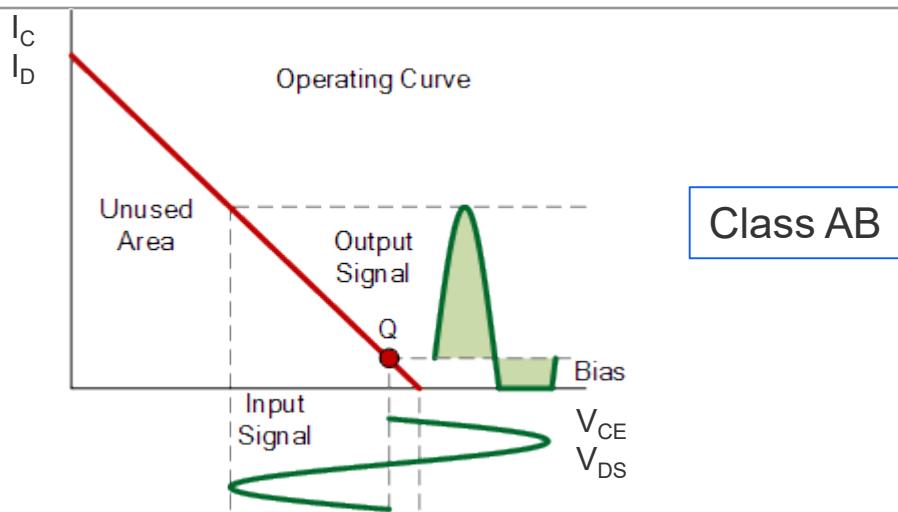
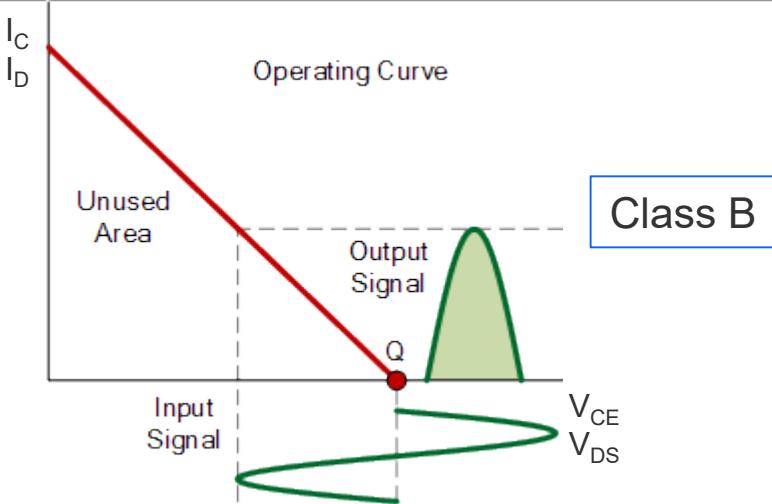
■ Reduced conduction angle

- Highly efficient PAs are demanded in several applications (e.g. mobile handset) in order to conserve battery power
- Increase the amplifier efficiency by **reducing the quiescent current** with a properly biased active device
- Additional cost is to be paid:
 - Driving power level must be substantially increased in some cases
 - Harmonic content must be shorted out

Categories of Amplifiers

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Class B and AB



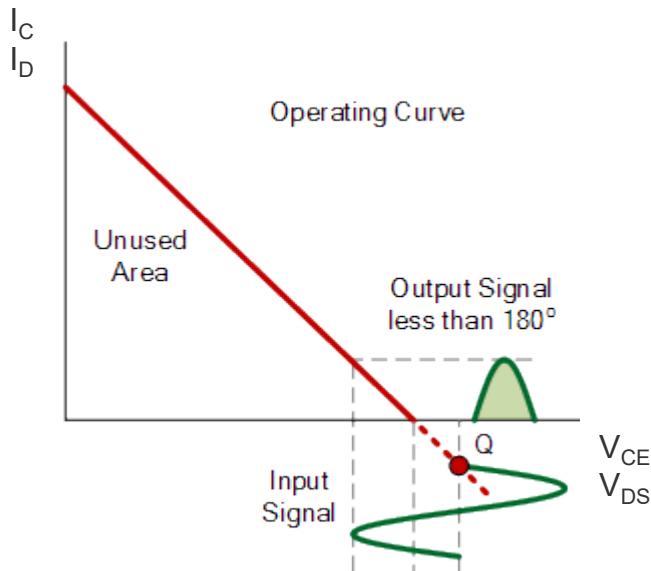
- Positive input signal: TRT conducts.
- Negative input signal: TRT is switched off (Biased in cut-off).
- Transistor conducting angle is 180°
- It is more efficient than Class A
- Higher distortion

- A combination of the “Class A” and the “Class B” type amplifiers.
- Transistors are biased just above its cut-off point.
- Conducting angle will be between 180° and 360° .
- Better efficiency than A and less distortion than B

Categories of Amplifiers

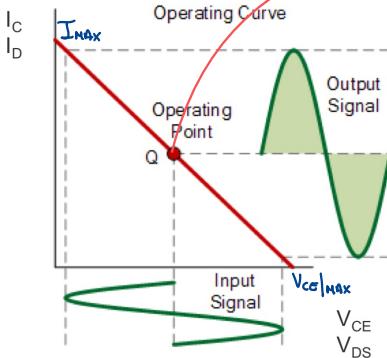
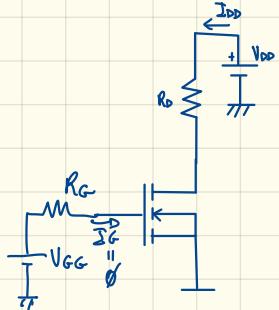
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Class C



- It has the greatest efficiency (around 80%) but the poorest linearity (higher distortion) of the previous ones.
- The conduction angle for the transistor is significantly less than 180 degrees

CLASE A



$\theta = 360^\circ \rightarrow \text{EL TRT ESTÁ EN CONDUCCIÓN DURANTE TODO EL PERÍODO DE } V_{IN}$

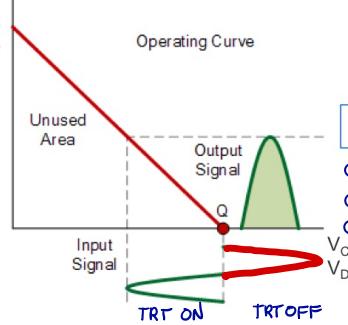
- Poco EFICIENTE
- Distorsión BAJA
- RÉGIMEN DE TRABAJO QUASI-LINEAR DENTRO DE $P \geq (1\text{dB} - PC)$

REQUISITO EFICIENCIA ALTA \rightarrow AHORRO CONSUMO DE BATERÍAS

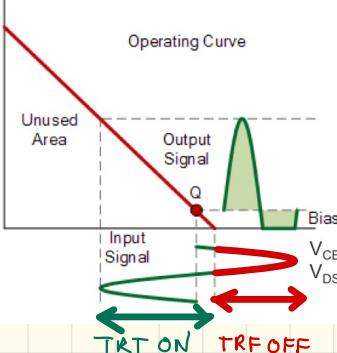
PROBLEMAS

- ① SE REQUIEREN P_{IN} MAYORES (Non-Linear)
- ② AUMENTA la Distorsión

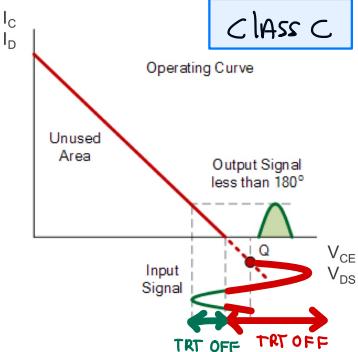
→ Reducir Periodos en los que el TRT CONDUCE
↓
MEJOR EFICIENCIA
¿COMO? CORTANDO EL TRT



- Class B
- $\theta = 180^\circ$
 - + EFICIENTE
 - + DISTORSIÓN



- Class AB
- $\theta \in [180^\circ - 360^\circ]$
 - MEJOR η QUE A
PEOR η QUE B
 - MENOS Distorsión QUE B

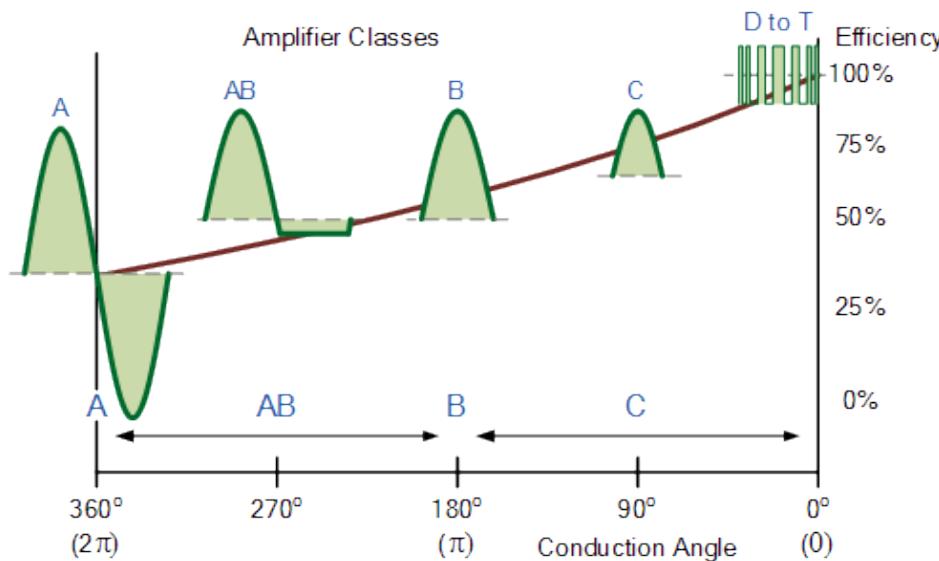


- Class C
- $\theta < 180^\circ$
 - MEJOR EFICIENCIA
 - $P_{IN} > 1\text{dB} - PC$ (PARA CONSEGUIR P_{OUT} GRANDES)
 - - ① Z_{IN}, Z_{OUT} CAMBIAN CON P_{IN} DE FORMA SIGNIFICATIVA
 - ↳ Complican DISEÑO Y ESTABILIZACIÓN
 - ② + DISTORSIÓN

Categories of Amplifiers

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■ Comparison

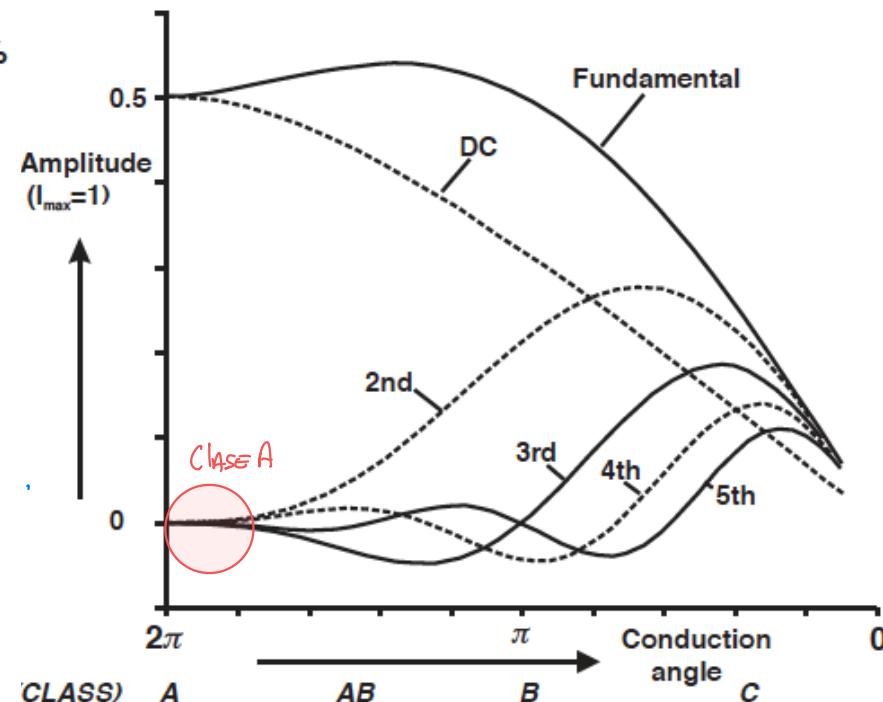
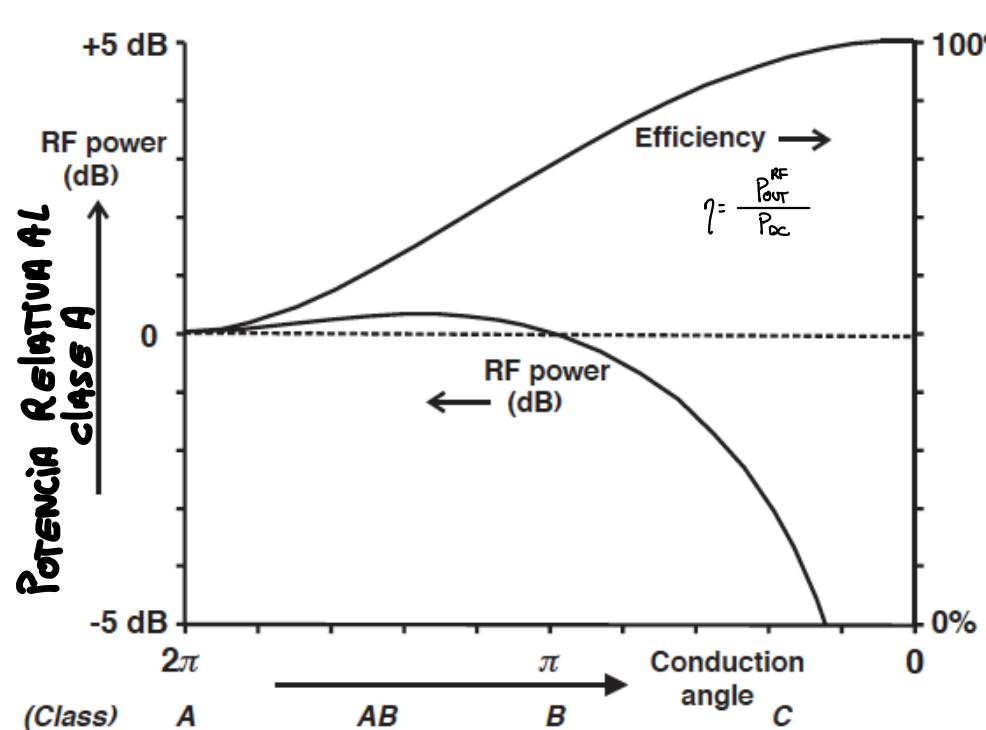


- A and AB are considered quasi-linear amplifiers, as the output signals amplitude and phase are linearly related to the input signals amplitude and phase.

Categories of Amplifiers

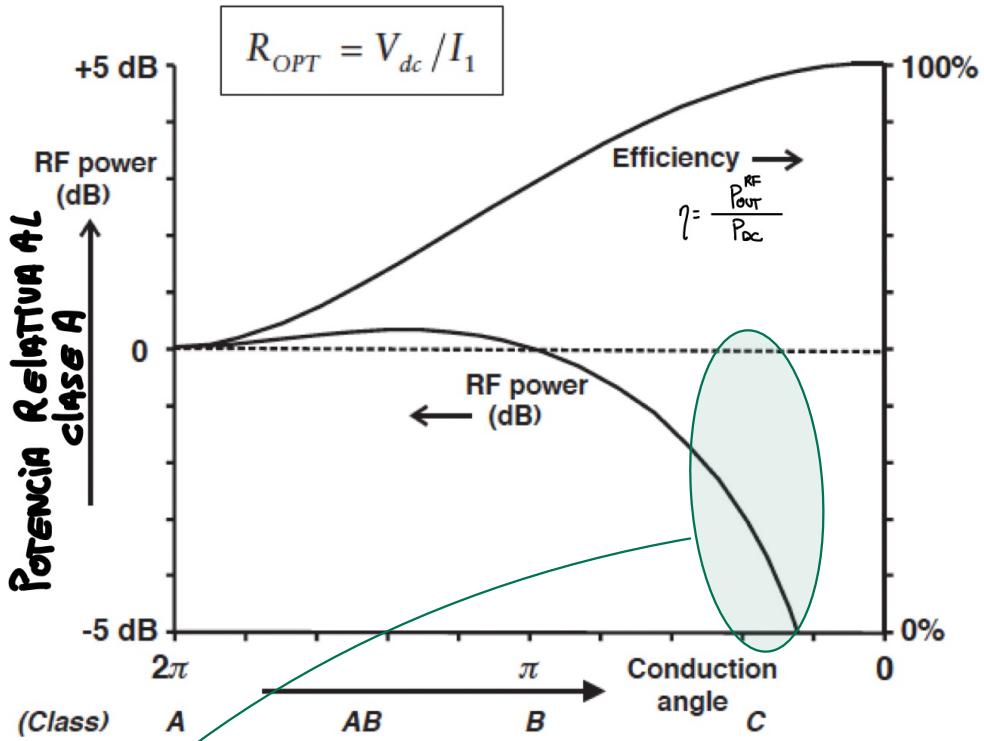
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- Comparison Results for a fixed input power



Results for perfect harmonic shorts,
maximum linear current and voltage swings

$$R_{OPT} = V_{dc} / I_1$$



PARA ALCUAR LA POTENCIA DE LA SEÑAL DE SALIDA
HAY QUE ALCUAR LA POTENCIA DE LA SEÑAL DE
ENTRADA. ESTO LLEVA AL TRT A TRABAJAR POR
ENCIMA DE 1 dB-PC Y ESTO PROVOCAR UN
AVVENTO DE LA DISTORSIÓN

Categories of Amplifiers

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■ Comparison

- Class A: Harmonic power negligible and comparable DC and RF levels.
- Class AB operation shows an almost constant fundamental RF output power, similar or slightly higher than Class A. The power of the 2nd harmonic starts increasing.
- Class B condition delivers the same output power as Class A but with a DC supply reduced by a factor $\pi/2$, therefore the efficiency increases. The level of the harmonics increases.
- Class C condition shows an ever-increasing efficiency as the conduction angle is reduced to lower values; however this is accompanied by a substantial reduction on the output RF power
- Harmonic components at the output, which also consume power, can be eliminated by using a filter. In RF, sometimes the own output impedance of the amplifier behaves as a filter.

Categories of Amplifiers

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■ Comparison

Clase	Ángulo de conducción	Eficiencia	Distorsión	Régimen de trabajo
A	360º	Baja (30%)	+baja	Quasi-linear
AB	360º-180º	Media (40%)	Media-baja	Quasi-linear
B	180º	Media (50%)	Media-alta	Non-linear
C	<180º	+Alta (80%)	+alta	Strong non-linear

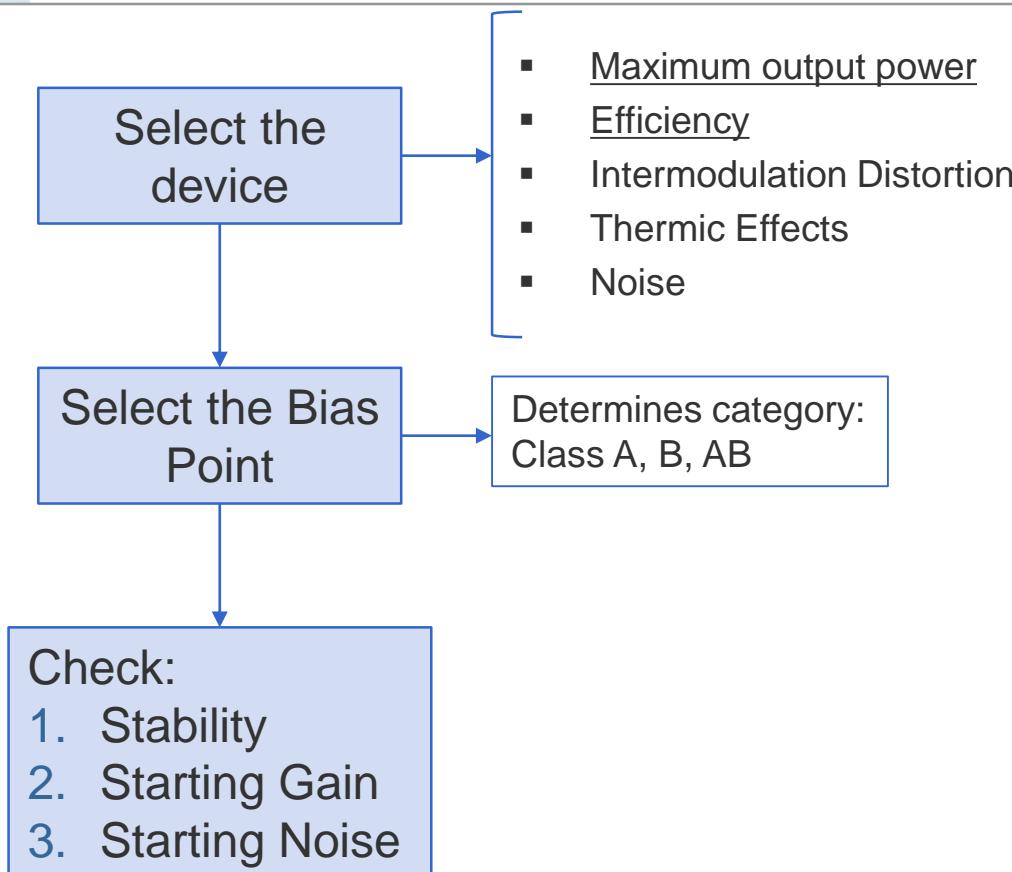


■ Design Procedure

Class A Amplifier

2/2

■ Design Procedure





Class A Amplifier

■ Stability

■ **K-factor (Rolleff Factor)**

$$k = \frac{1 - |s_{11}|^2 - |s_{22}|^2 + |D|^2}{2 \cdot |s_{21}| \cdot |s_{12}|}$$

$$D = s_{11} \cdot s_{22} - s_{12} \cdot s_{21}$$

- If $k > 1$ then the device will not present an input/output reflection coefficient higher than unity
- The former result is verified for any source and/or load (i.e. the amplifier is *unconditionally stable*)

Class A Amplifier

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- **Stability**

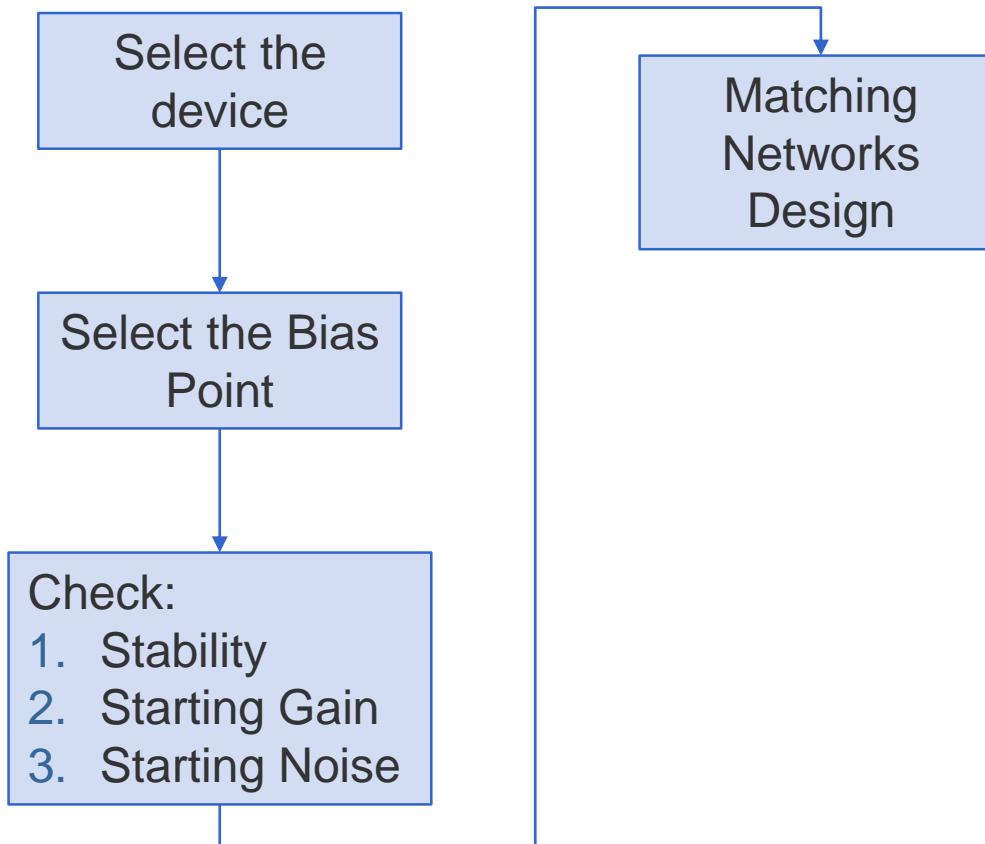
- **Some practical considerations about Rollet factor**

- 1) If k is slightly higher than unity, device gain/matching are higher than the unilateral equivalent
- 2) Circuit losses can compensate a k -factor slightly lower than unity for a given active device
- 3) Circuit implementation and parasitic can impact the active device S-parameters (in particular S_{12}) and then stability
- 4) Rollet factor is applicable strictly to single-stage amplifiers

Class A Amplifier

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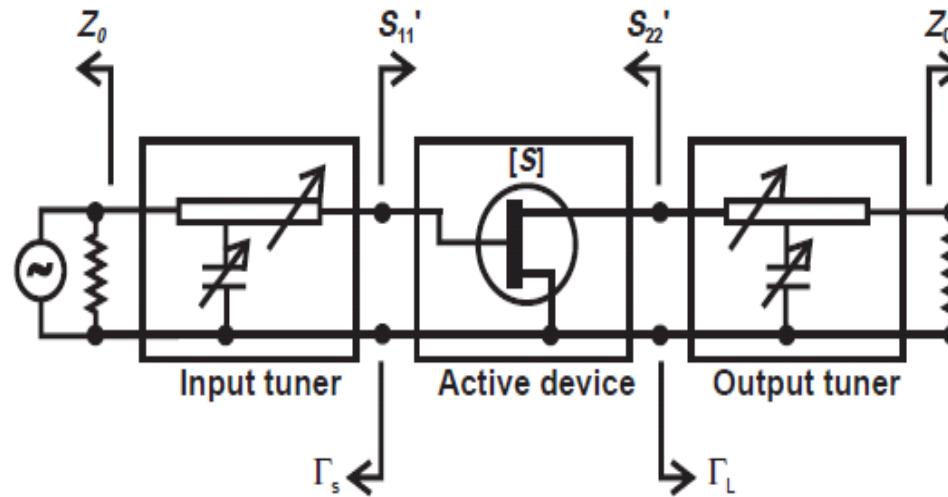
■ Design Procedure



Class A Amplifier

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■ Matching networks



Power Amplifier

$$\text{MTP} \Rightarrow \Gamma_s = \overline{S'_{11}}$$

Max. output power $\Rightarrow \Gamma_L = \Gamma_L \text{ max power}$

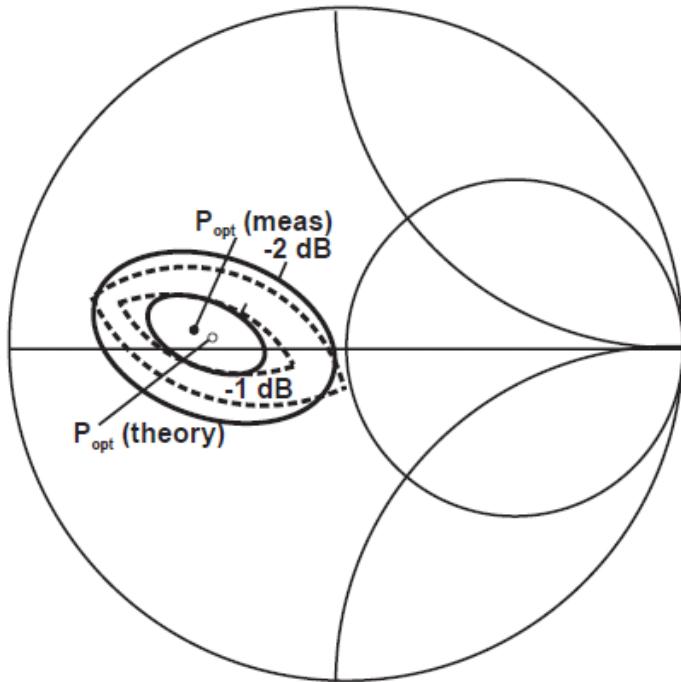
$$R_{opt} = \frac{V_{DD} - V_{SAT}}{I_{DMAX}}$$

Class A Amplifier

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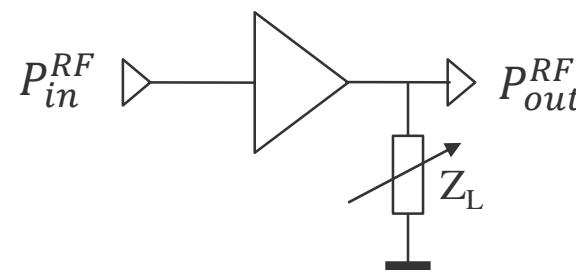
■ Load-Pull measurements

- Relationship between power and output matching represented in the Smith Chart



Solid experimental-dashed theoretical

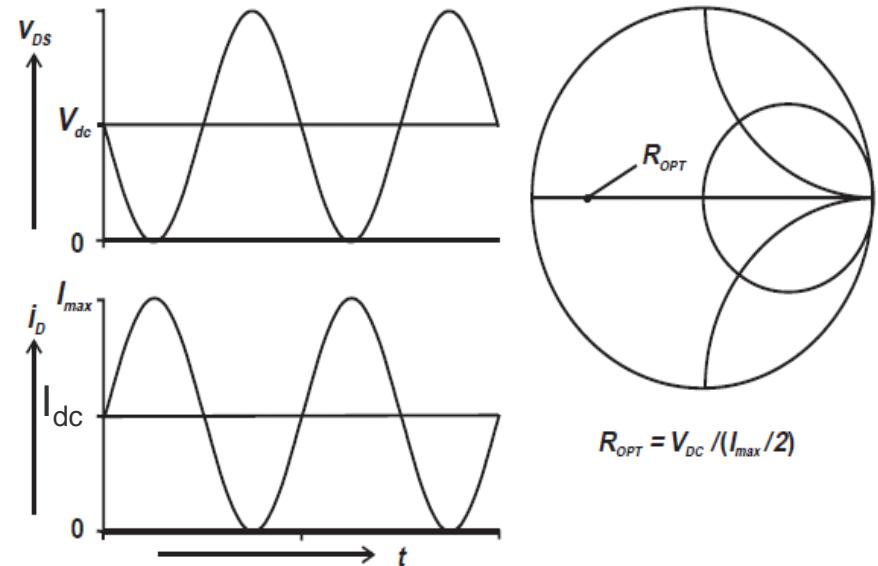
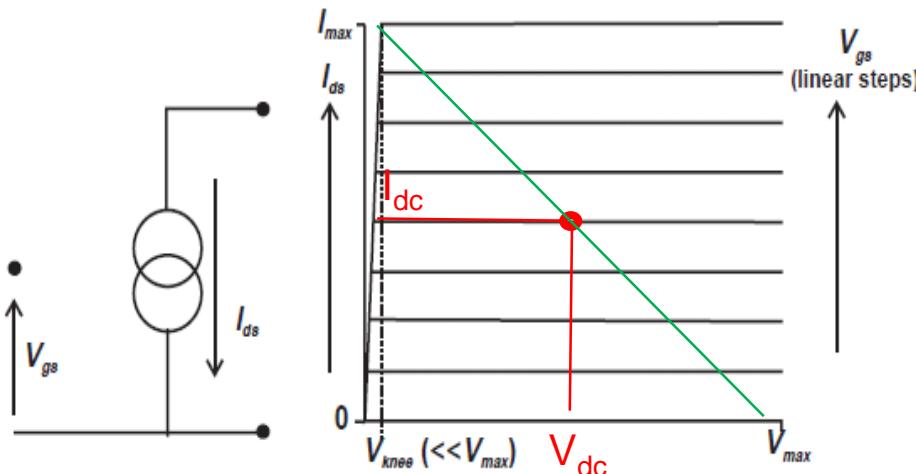
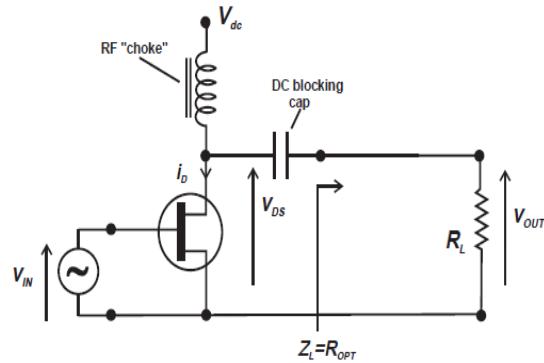
- Measurement setup comprises the DUT and an output impedance tuner
- Contours show boundaries of specific output power levels
- Load-Pull contours at 1dB and 2dB are generally used



Class A Amplifier

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- **Loadline match:** Optimum load resistor for power match



$$R_{opt} = V_{dc} / (I_{max}/2) = V_{dc} / I_{dc}$$

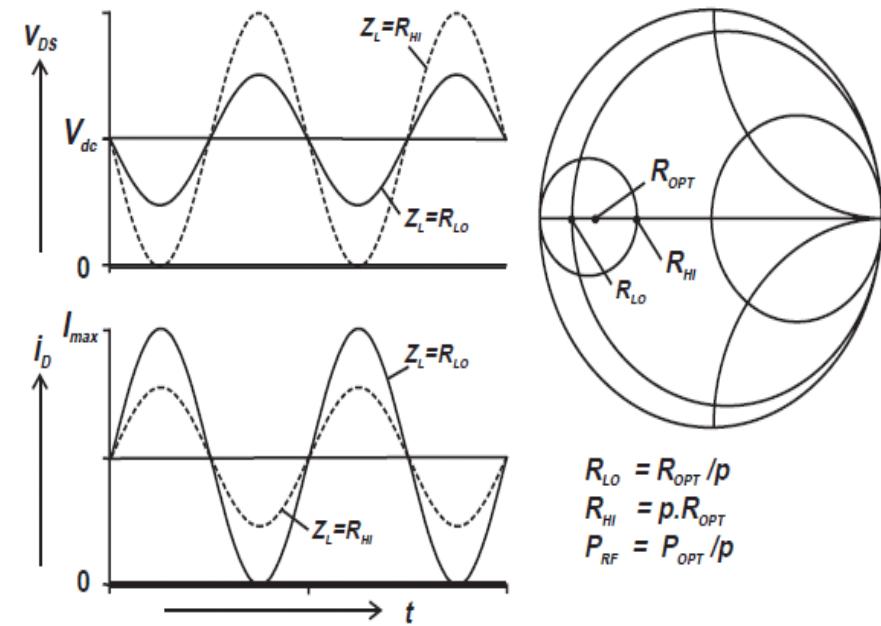
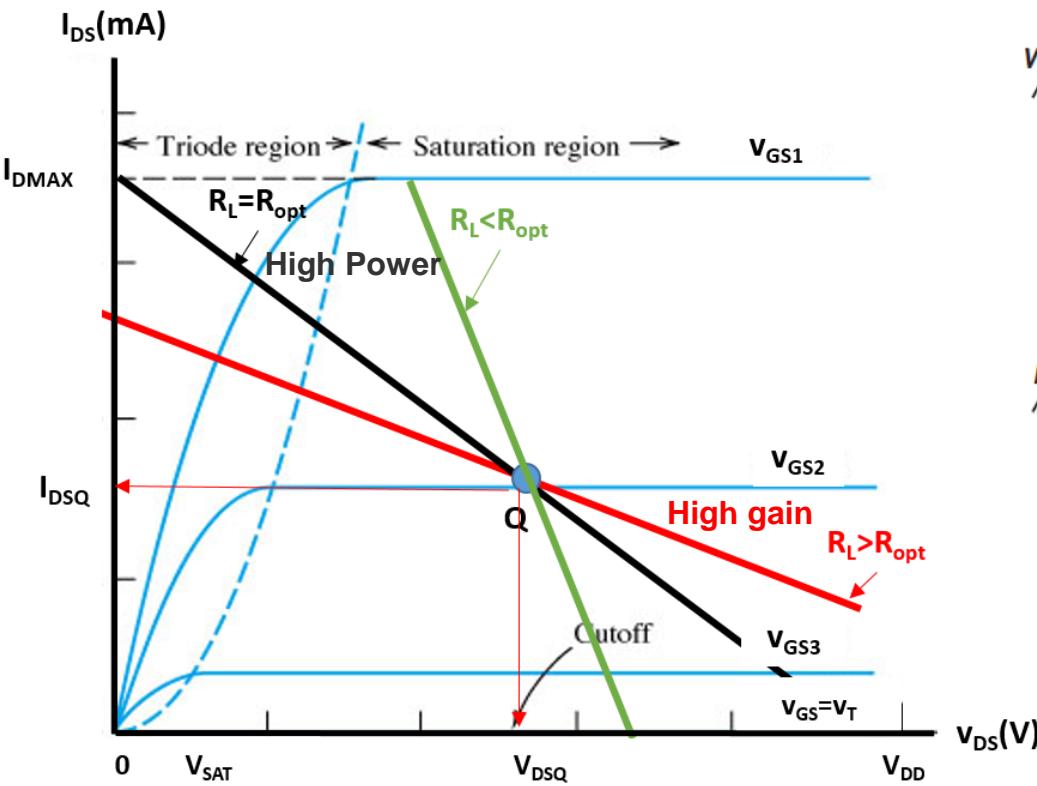
$$P_{opt} = (1/2) \cdot V_{dc} \cdot I_{dc}$$

Class A Amplifier

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- Loadline Theory

A) Resistive Load terminations giving an output power P_{opt}/p



$\underline{RL < R_{opt}}$: Voltage limited
 $\underline{RL > R_{opt}}$: Current limited

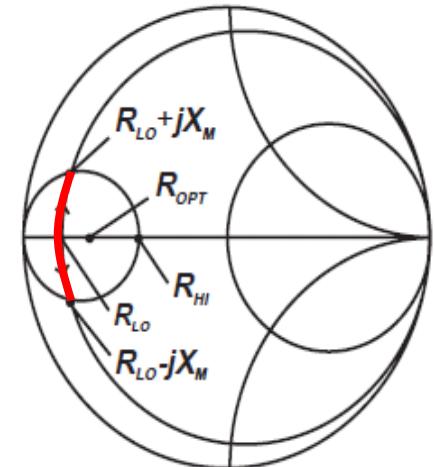
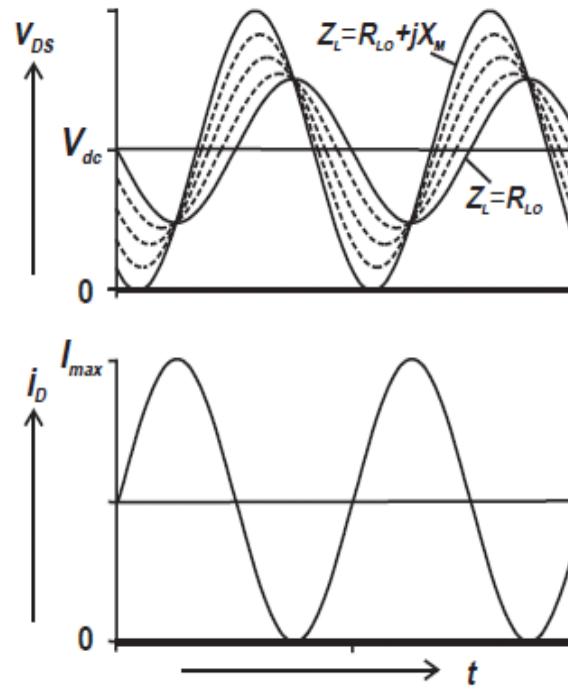
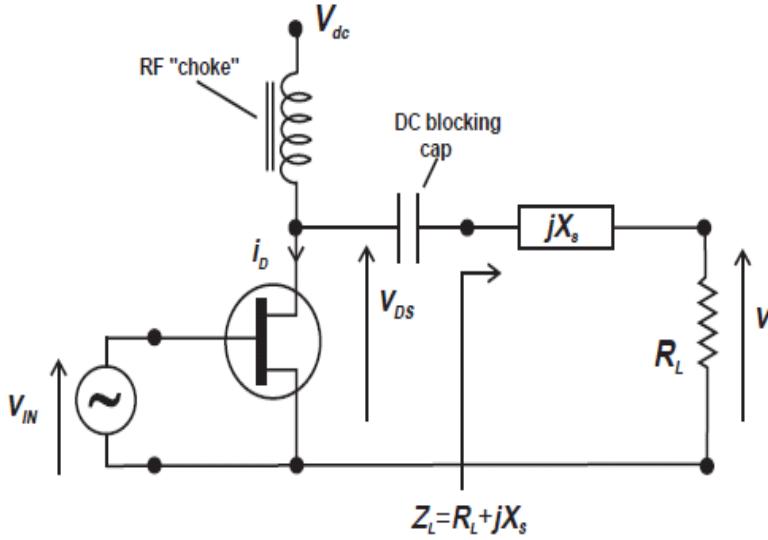
Class A Amplifier

10/2

- Loadline Theory

B) Series reactance

Constant current excursion while variable voltage swing



$$R_L = R_{LO} = R_{OPT} / p$$

$$R_{HI} = p \cdot R_{OPT}$$

$$P_{RF} = I_{MAX}^2 \cdot R_L / 8 \text{ for } -X_M < X_s < X_M$$

$$P_{RF} = P_{OPT} / p \text{ for } -X_M < X_s < X_M$$

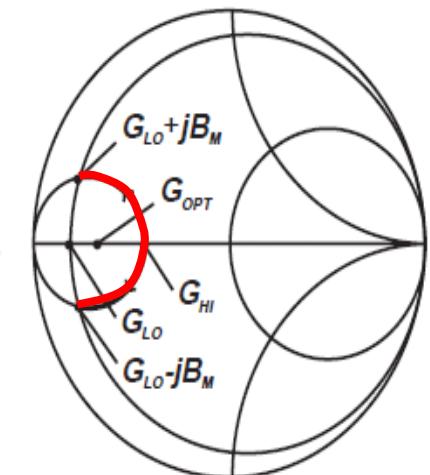
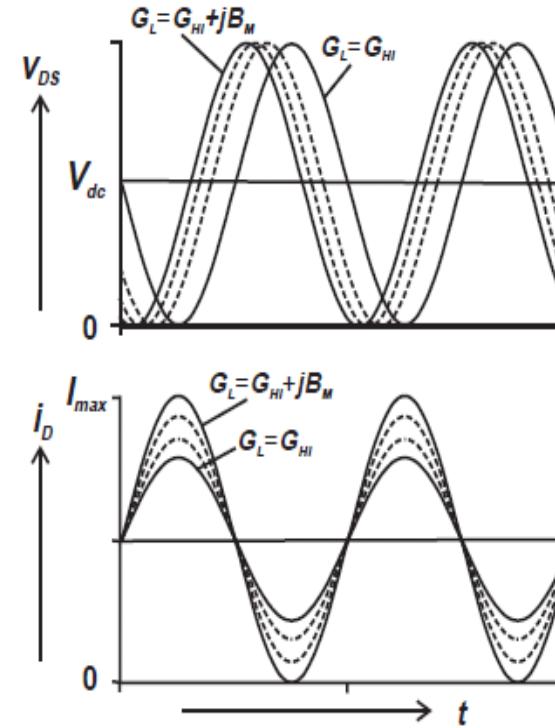
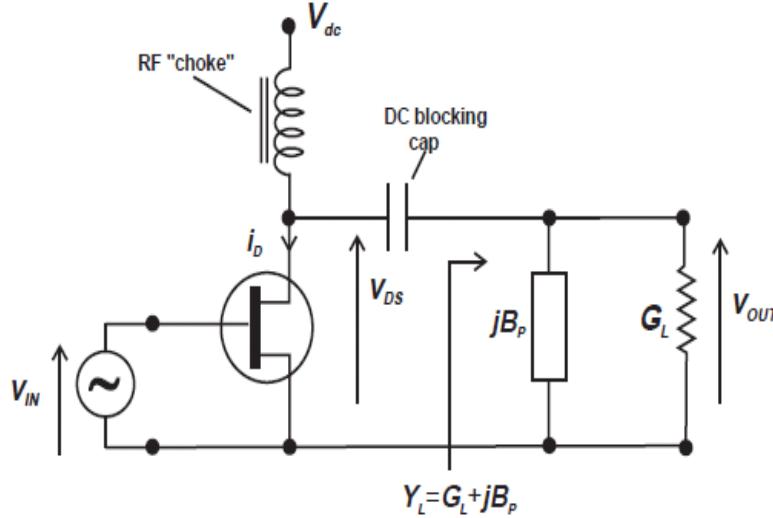
where $(R_{LO}^2 + X_M^2)^{1/2} = R_{OPT}$

Class A Amplifier

- Loadline Theory

C) Shunt reactance

Constant voltage swing and variable current excursion



$$G_{HI} = 1/p \cdot R_{OPT}$$

$$P_{RF} = V_{DC}^2 \cdot G_L / 2 \text{ for } -B_M < B_p < B_M$$

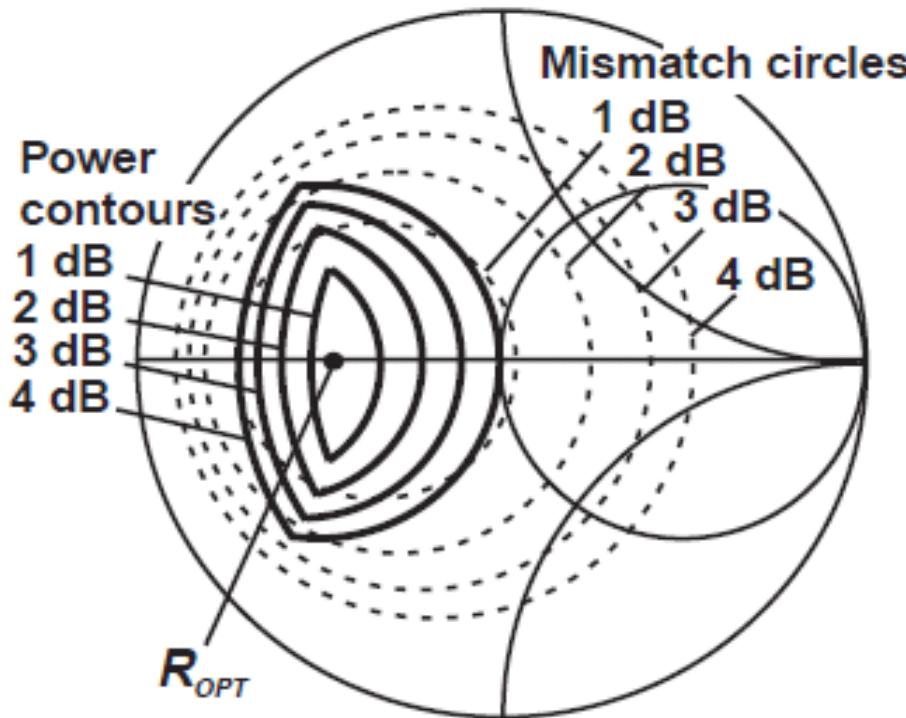
$$P_{RF} = P_{OPT} / p \text{ for } -B_M < B_p < B_M$$

$$\text{where } (G_{HI}^2 + B_M^2)^{1/2} = 1/R_{OPT}$$

Class A Amplifier

12/2

Load-Pull Contours



- The point gives the impedance value which provide the maximum output power (optimum resistance).
- The contours around the optimum impedance value show boundaries of specific output power levels, and they give other impedance values which provide a power X dBs below the maximum value.
- Mismatch circles also are represented.
- Higher values than 4dB have no sense, since the resulting design will be the same as the one for MTP (which is easier).

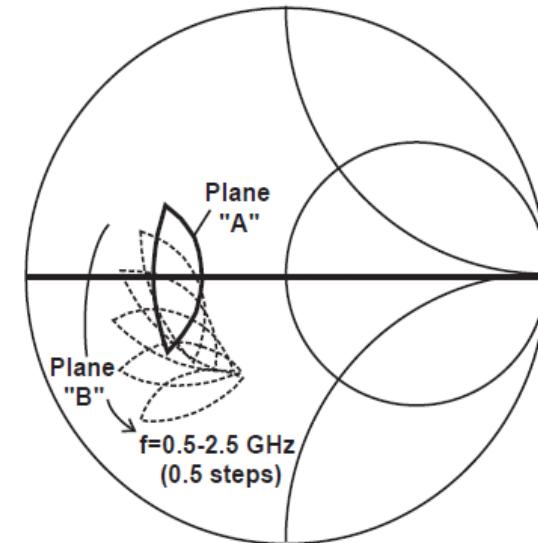
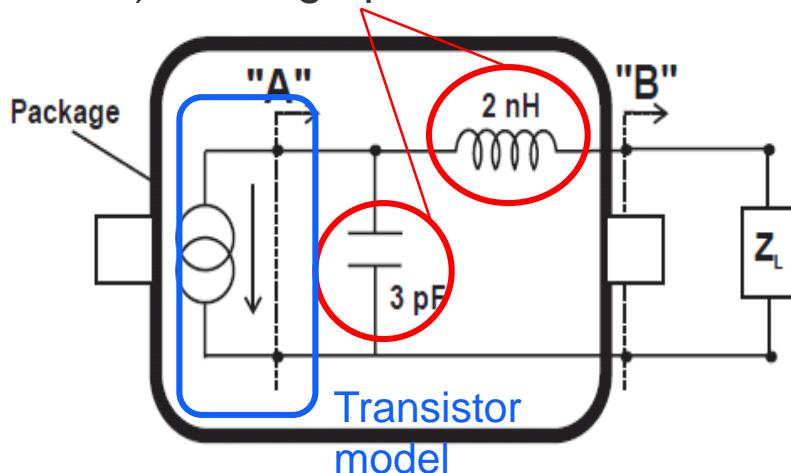
Class A Amplifier

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■ Additional considerations

- Several issues must be taken into account for reaching a good matching between theoretical and measured *load-pull* contours
- Measured contours at drain terminal do not depend on frequency. However, in practice the access point is out of the package device. At this point parasitics introduce the dependence with frequency (due to the reactances) this produces the contours movement (Cripps pages 27-31).

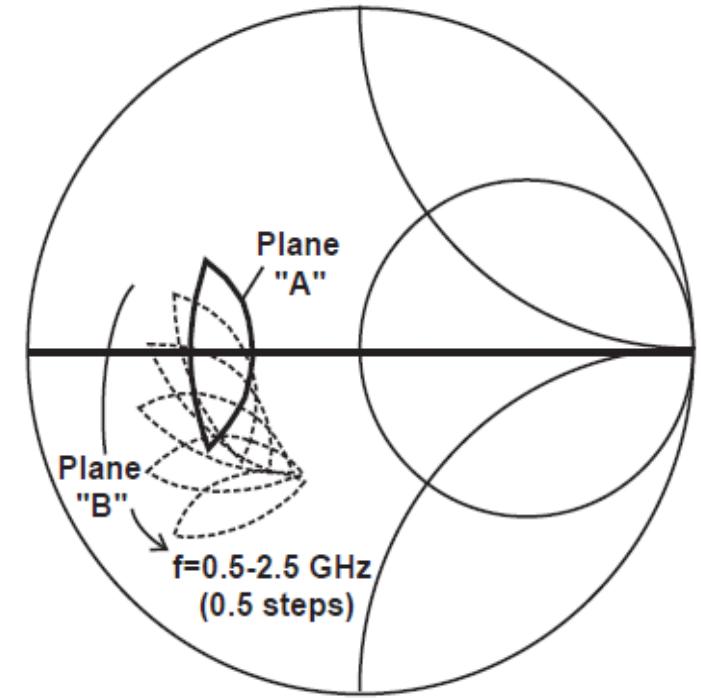
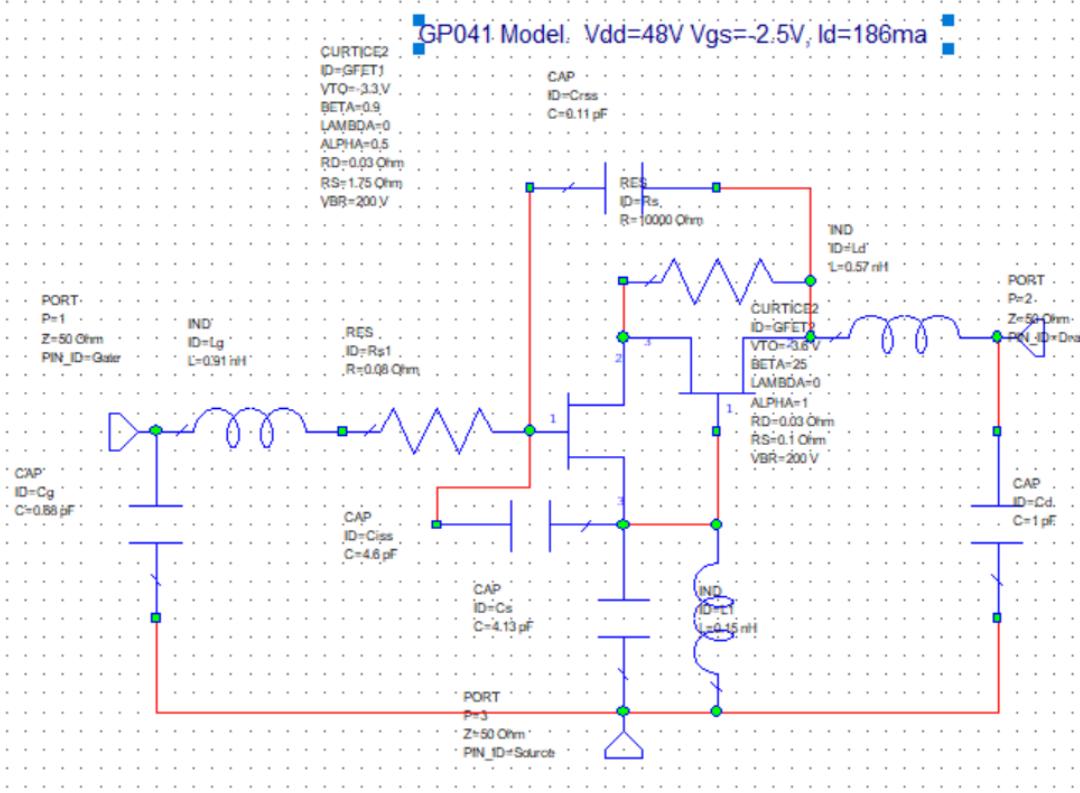
a) Package parasitics



Class A Amplifier

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Additional considerations



Class A Amplifier

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■ Design Procedure

