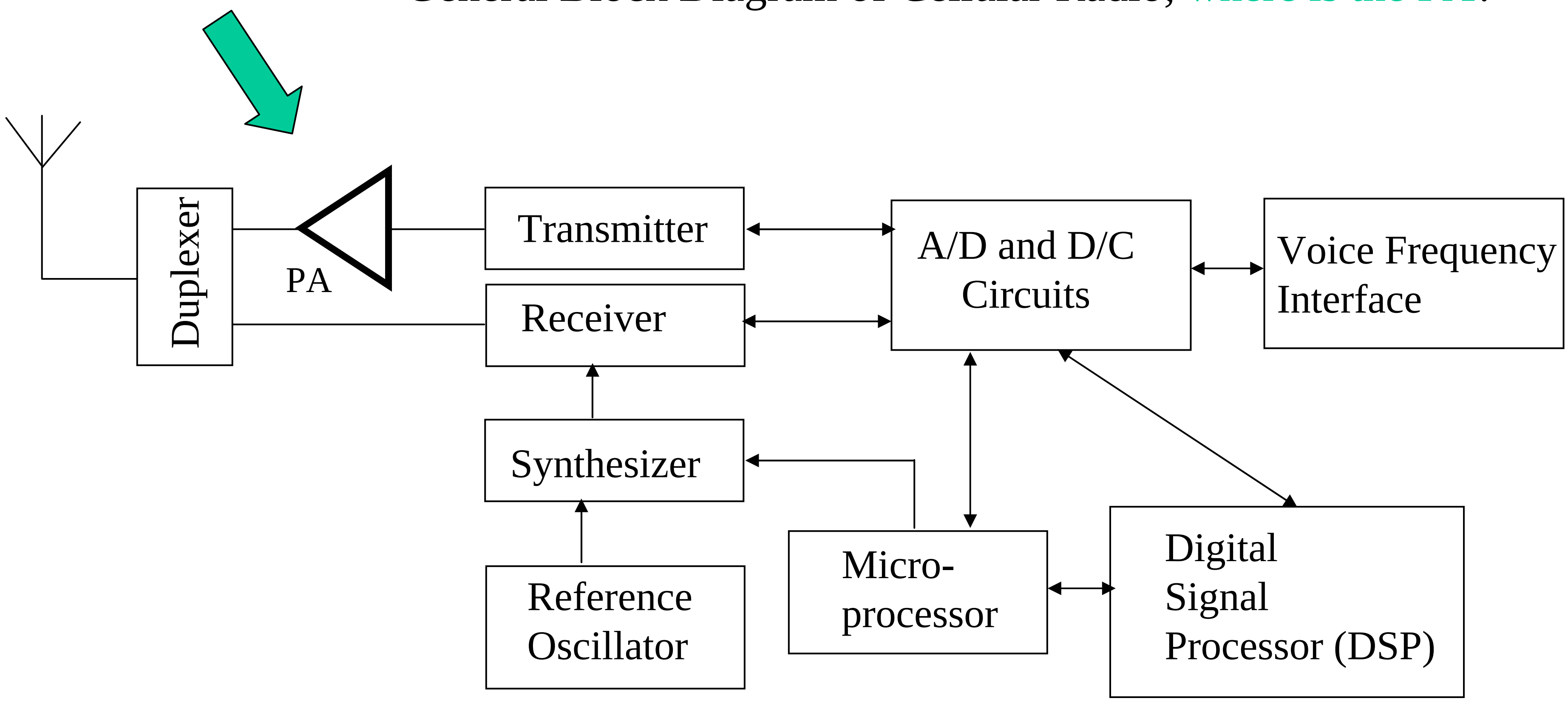
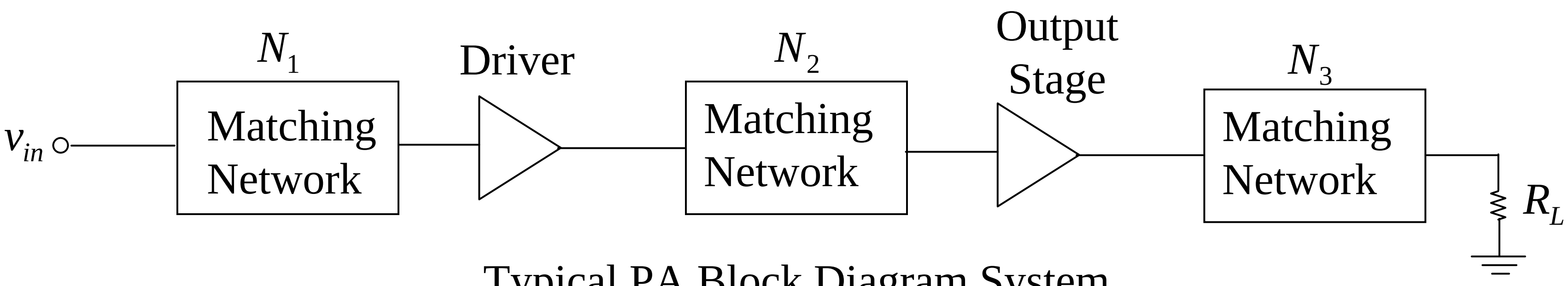
POWER AMPLIFIERS

General Block Diagram of Cellular Radio; where is the PA?

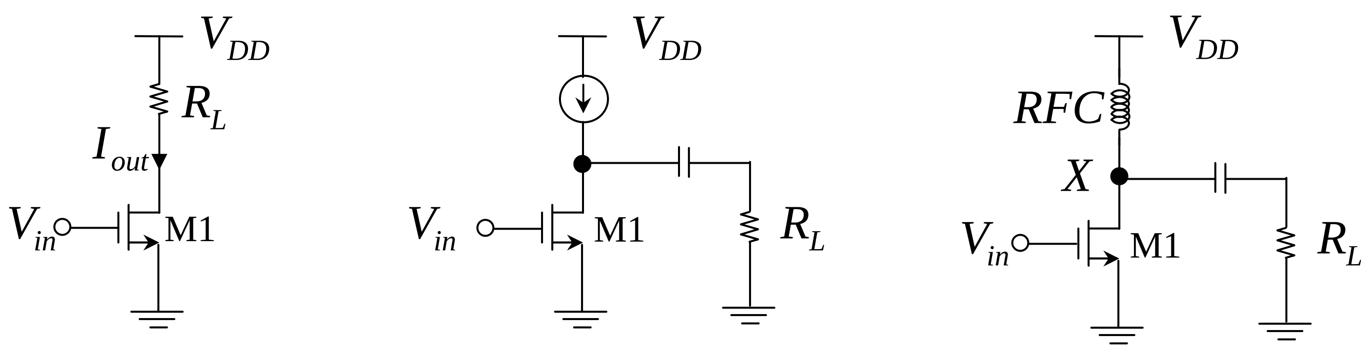


Power Amplifier System Level Considerations



Typical PA Block Diagram System

- Why can designers use only an amplifier? like this:



Common-Source Stages

To deliver for instance a 1W of power to a 50 Ω antenna with common-source is not feasible.

- Power-Amplifier Metrics
  - Power Efficiency η_{PA}

$$\eta_{PA} = \frac{P_{load}}{P_{supply}} = \frac{P_{RF-out}}{P_{DC-in}}$$

Ideal $\eta_{PA} = 100\%$

  - Power-Added Efficiency (Linear concept)

$$PAE = \frac{P_{load} - P_{input}}{P_{supply}} = \frac{P_{load}}{P_{supply}} - \frac{P_{input}}{P_{supply}}$$

$$PAE = \frac{P_{load}}{P_{supply}}\left(1 - \frac{P_{input}}{P_{load}}\right) = \eta_{PA}\left(1 - \frac{1}{G}\right)$$

where $G = \frac{P_{RF-out}}{P_{RF-in}}$

  - Output-Referred Compression $P_{o\text{-}1dB}$ (non-linear concept)

$$P_{o-1dB} \triangleq G\ P_{in-1dB}$$

$$P_{o-1dB} = 10\log G + P_{in-1dB}$$

Power Amplifier for GSM Handset

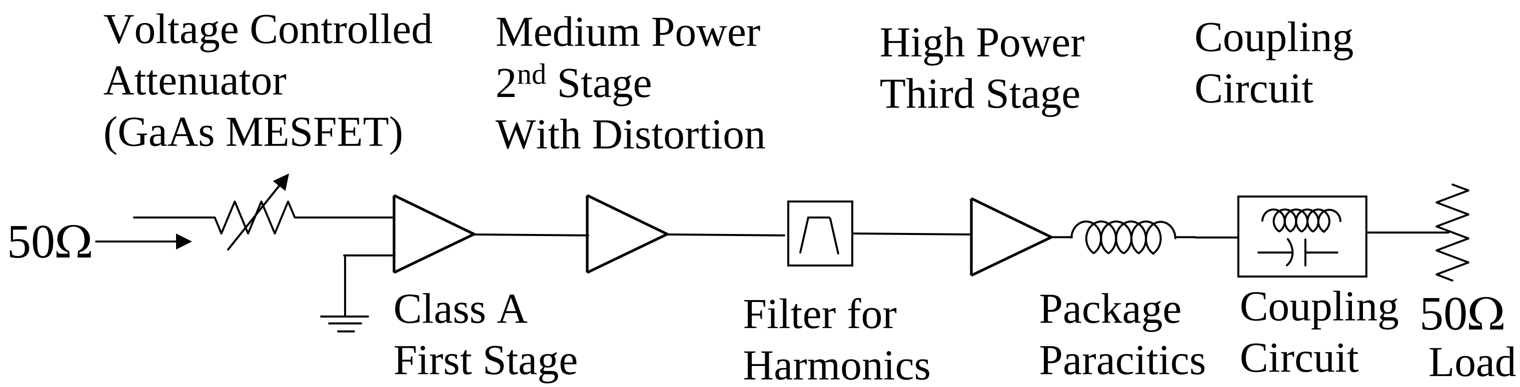
Typical Requirements of Power Amplifier:

- Positive supply only
- 890 to 915 MHz transmit frequency (GSM 900)
- ≈ 3 Watts power out of amplifier (≈ 1.2 W allowed but not popular)
- Efficiency > 40% (Market place issue)
- < 4 dB compression (Affects efficiency, limited by spurious signal accentuation)
- Gain Control capability > 34 dB (Allows for variation in other parts of transmit path)
- Supply voltage: 5 V (min.) to 8.5 V (during battery charge, in operation)
  Because of inductive loads, output device breakdown requirement may be > 17 V.

System Control Requirements:

- Strict turn-on time template with limits on spurious outputs during turn-on
- Control Range; 28 dB in 2 dB steps
- Absolute tolerance; ± 2.5 dB for highest power and ± 6 dB for lowest power
- Flatness control of output pulse (± 1.0 dB)

Power Amplifier for GSM Handset, Three Stages with Power Control



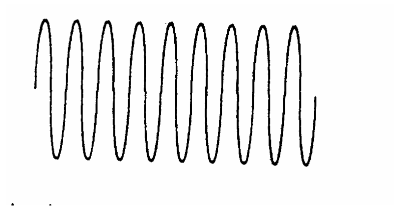
Key Questions:

- Q. Is the inter-stage band-pass filter necessary?
  A. Yes. The distortion requirements are severe. Desired efficiency cannot be met without some compression, even on the second stage.

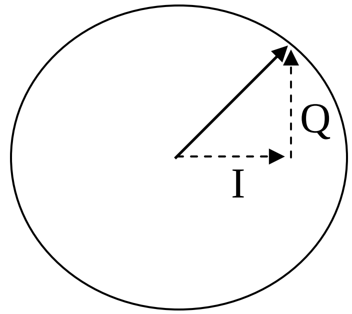
- Q. Why not include output coupling circuit inside package?
  A. Only low Q inductors are obtainable on-chip, which affects efficiency.

- Q. Why not a 2 or 3 Volt supply?
  A. Impedance level is so low (< 3 Ω) that package inductance and Q still present problems.

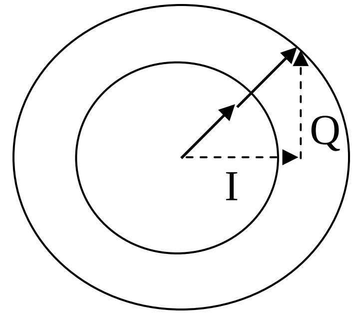
Effect of PA Clipping on Phase Error



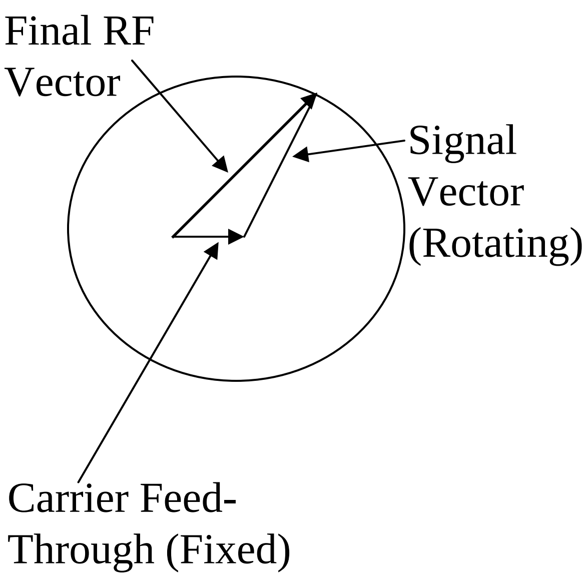
Ideal GSM (GPSK) Signal
Has No AM (Flat Envelope)

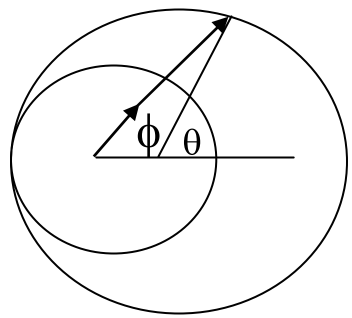


Signal Vector Only

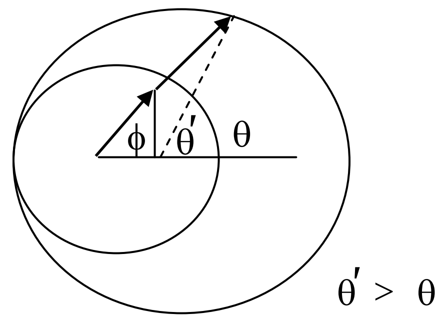


Clipping Causes No
Change in Phase



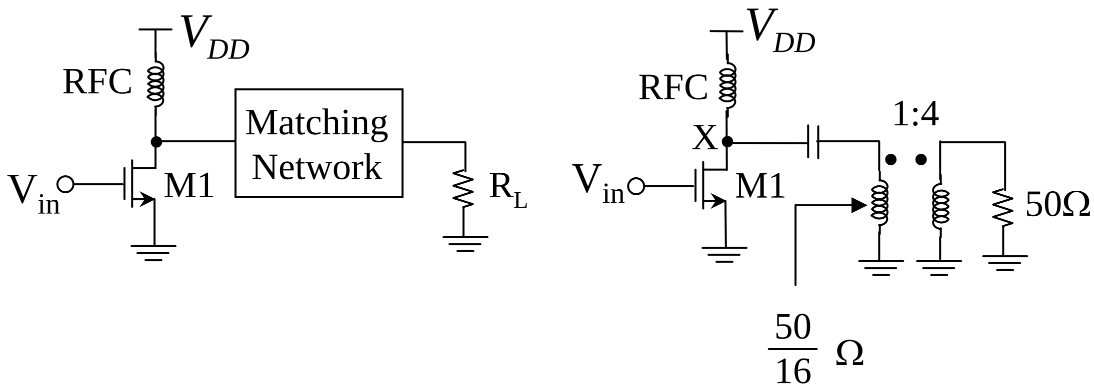


Clipping Cuts the
Final RF Vector to a
Constant Amplitude



Crossover Phase (φ) is Unaffected
But Signal Phase (θ) is Modified

How to relax power dissipation across the transistor?
One solution to alleviate this situation follows:



(i) (ii)

In general, for practical PA designs it is required matching networks to increase (and distribute) power gain.

(i) Matching network as voltage amplifier, (ii) use of a transformer as a matching network.

Class A Amplifiers

$$i_D = I_{DC} + i_{RF} \sin \omega_o t$$

$$v_{o,RF} = -i_{RF} R \sin_o t$$

$$P_{RF} = \frac{i_{RF}^2 R}{2}$$

$$P_{RF} = \frac{v_{o,RF}^2}{2R}$$

$$P_{sup\,ply} = \frac{v_{oi,RF} V_{DD}}{R}$$

Thus, the power efficiency for a class A amplifier yields:

$$\eta_A = \frac{v_{o,RF}}{2V_{DD}}$$

Observe that in order to keep the transistor in saturation, $v_{DS} > V_{DSAT} = V_{GS} - V_T$ which becomes

$$V_{DST} = \frac{I_D + \sqrt{I_D^2 + 2kI_D L^2 E_{SAT}^2}}{kLE_{SAT}} \quad ; \quad |v_{o,RF}| \le |V_{DD} - V_{DSAT}|$$

where

$$k = \mu_o C_{ox} \frac{W}{L}, \; I_D = \frac{k}{2} \frac{(V_{GS} - V_T)^2}{1+\theta} \quad ; \quad \theta = \frac{V_{GS} - V_t}{LE_{SAT}}$$

E_{SAT} is the field strength in V/m at which the carrier has reached velocity saturation. Furthermore, remember that:

$$f_T \cong \frac{g_m}{2\pi C_{gs}} = \begin{cases} \dfrac{3}{4} \dfrac{\mu(V_{GS} - V_T)}{L^2} & \text{for long channel} \\ \dfrac{1}{4\pi} \dfrac{\mu E_{SAT}}{L} & \text{for short channel} \end{cases}$$

Similarly for the BJT

$$f_{T,BIP} \cong \frac{1}{2\pi\tau_F} = \begin{cases} \dfrac{\mu kT}{\pi q \omega_B^2} & , \quad \text{low - level injection} \\ \dfrac{\mu E_{SAT}}{4\pi\omega_B} & , \quad \text{velocity saturation} \end{cases}$$

Typical values are

$$E_{SAT} \cong 4\times10^6\, v/m \;, \mu_{channel} = 1.5\times10^{-2}\, m^2/V_S \;, \mu_{BULK} = 5.5\times10^{-2}\, m2/V_S$$

$$f_{T,NMOS} \sim 20-25\,\text{GHz}\,, 0.25\mu_m \;, V_{GS} - V_t = 0.8V$$

$$f_{T,BJT} \sim 30GHz \quad , \quad \omega_B = 0.05\mu_m \quad , \quad V_{be} = 0.8V$$

Example. Assume V_{DD}=3.V and V_{DSAT}=0.42

$$\eta_A = \frac{v_{o,RF}}{2V_{DD}} = \frac{3.3-0.5}{2\times3.3} = 0.42$$

$$A_V = -g_m R_{eq}$$

$$G_A = g_m^2 R_{eg} R_s$$

R_s is the source resistance, and $R_{eg} = \dfrac{v_{o,RF}^2}{2P_{LOAD}}$. Typically G ~ 10dB, then the PAE is at least 90% of the power efficiency.

Let us now explore the output 1-dB compression point.

Recall that

$$g_m = \frac{\mu C_{ox} W}{L}\left(V_{GS} - V_T\right)\frac{1 + 0.5\theta}{\left(1+\theta\right)^2}$$

Then

$$\frac{G_A}{R_{eq}^2} P_{1-dB} = \left|g_m\right|^2 R_s\, P_{1-dB} / R_{eq}$$

$$P_{1-dB} = \frac{2}{9}\left|\frac{a_1}{a_3}\right|\frac{10^{0.05} - 1}{R_s}$$

a_1 and a_3 are the coefficients of the non-linear polynomial of g_m.

Then the output compression point becomes:

$$|a_1|^2 \frac{R_s P_{1-dB}}{R_{eq}} = 10\,log\left[4.3384\times 10^2 \frac{(1+0.5\theta)^3}{\theta} P_{load}\right]$$

Example.- V_{DD} = 3.3V, V_{DSAT} = 0.5 (in a $0.25\mu_m$ CMOS technology), and a transmit power of 600 mW. (27.8dBm).

Then

$$10\,log\left[4.3384\times 10^2 \frac{(1+0.5\theta)^3}{\theta} P_{load}\right]_{\substack{P_{load}=0.6\\ \theta=0.5}} = 30.1dBm$$

Note that 30.1dBm is about 2.3dB higher thant the $P_{o\text{-}1dB}$

$$\eta_A = \frac{v_{o,RF}}{2V_{DD}} = \frac{V_{DD}-V_{DSAT}}{2V_{DD}} = \frac{3.3-0.5}{2\times 3.3} = 42\%$$

$$R_{eq} = \frac{(V_{DD}-V_{DSAT})^2}{2P_{load}} = \frac{(3.3-0.5)^2}{2\times 0.6} = 6.5\Omega$$

The corresponding bias current is

$$I_D = \frac{V_{DD} - V_{DSAT}}{R_{eq}} = \frac{3.3 - 0.5}{6.5} = 4.30.8mA$$

From the expression of I_D one can determine W

$$W = \frac{2LI(1+0)}{\mu C_{ox}(V_{GS} - V_t)^2} = 2I_D \frac{LE_{SAT} + V_{DSAT}}{\mu C_{ox} V_{DSAT}^2 E_{SAT}}$$

$$W = 2 \times 0.4308 \frac{0.25 \times 10^{-6} \cdot 4 \times 10^6 + 0.5}{0.015 \cdot 0.0035 (0.5)^2 4 \times 10^6} = 24.6mm$$

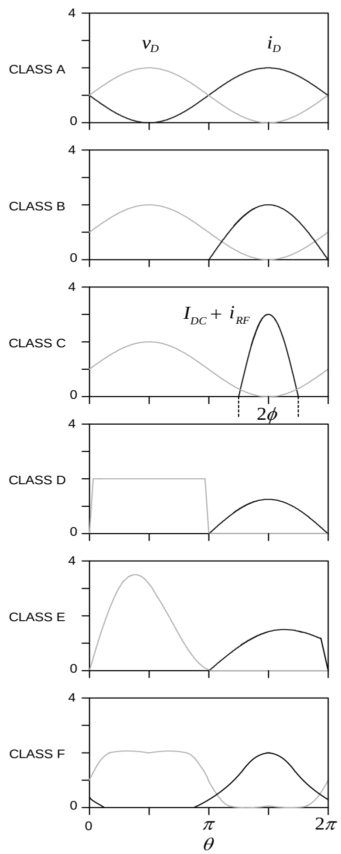
Its corresponding transconductance becomes:

$$g_m = \sqrt{\frac{2\mu C_{OX} W I_D}{L}} \frac{1 + 0.5\theta}{(1+\theta)^{1.5}} = \sqrt{\frac{2 \times 0.015 \times 0.0035 \times 0.0246 \times 0.4308}{0.25 \times 10^{-6}}} \frac{1 + 0.5^2}{(1+0.5)^{3/2}}$$

$$g_m = 1.44 \ \mho$$

$$G = g_m^2 R_{eq} R_s \Big|_{R_S = 50} = 1.44 \times 6.5 \times 50 = 2.8.3dB \Rightarrow P_{in,2dB} = 1.8dB$$

The efficiency for class-A power amplifier is rather modest, thus other efficient amplifier are investigated.



Waveforms for Ideal PAs.

45% <

$$\eta_A = \frac{v_{o,RF}}{2V_{DD}}$$

78% <

$$\eta_B = \frac{\pi v_{o,RF}}{4V_{DD}} = 0.5\pi\eta_A \quad , \quad G_B = \frac{G_A}{4}$$

Distortion ↑

$$\eta_C = \frac{2\phi - sin\,2\phi}{4(sin\,\phi - \phi\,cos\,\phi)}$$

Efficiency ↑

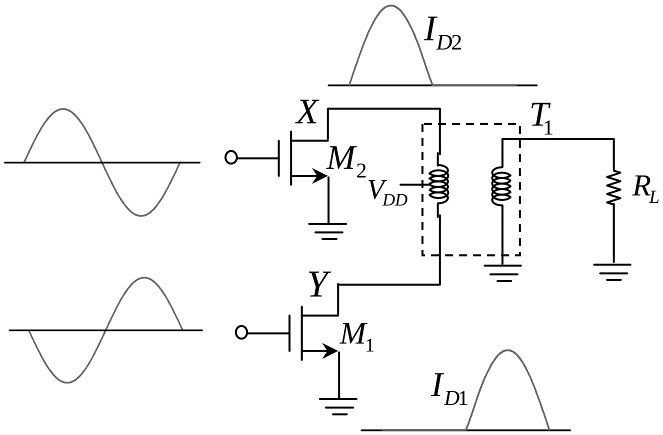
$$\eta_{D,IDEAL} = 100\% \quad \text{with ideal switches}$$

$$\eta_{E,IDEAL} = 100\%\ , max\,P_{LOAD} = \frac{2}{1+\pi 2/4}\frac{V_{DD}^2}{R} \cong 0.577\frac{V_{DD}^2}{R}$$

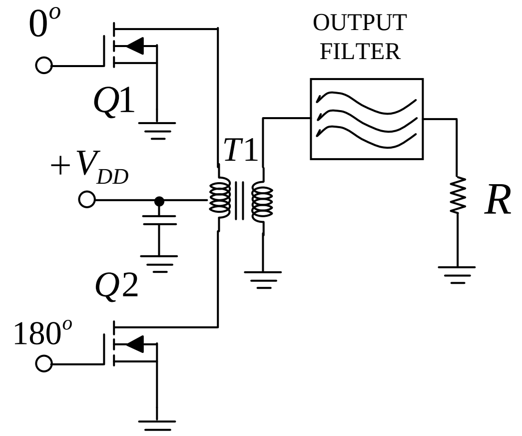
$$\eta_{F,IDEAL} = 100\%$$

Observe that class C, D, E and F amplifiers are essentially constant-envelope amplifiers.

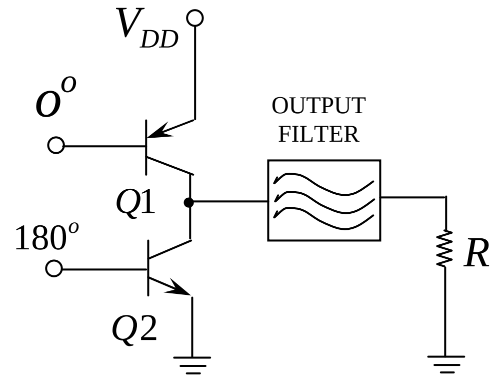
CLASS-B POWER AMPLIFIER



Class B stage using a transformer
For Broadband Operations.



Transformer-coupled push-pull PA. For audio applications.



Complementary PA.

- 50% duty cycle

$$i = \frac{2}{T}\int_{o}^{T/2} i_{RF}\, sin\,\omega_o t \times sin\,\omega_o t\; dt = \frac{i_{RF}}{2}\;;\; max\; i_{RF} = \frac{2V_{DD}}{R}$$

$$v_{o,RF} \cong \frac{i_{RF}}{2} R\, sin\,\omega_o t$$

$$P_o = \frac{v_{0,RF}^2}{2R} \quad ; \quad P_{o,max} = \frac{V_{DD}^2}{2R}$$

$$i_{D,average} = \frac{1}{T}\int_{o}^{T/2} \frac{2V_{DD}}{R} sin\,\omega_o t\; dt \frac{2V_{DD}}{\pi R}$$

$$P_{DC} = \frac{2V_{DD}^2}{\pi R}$$

thus

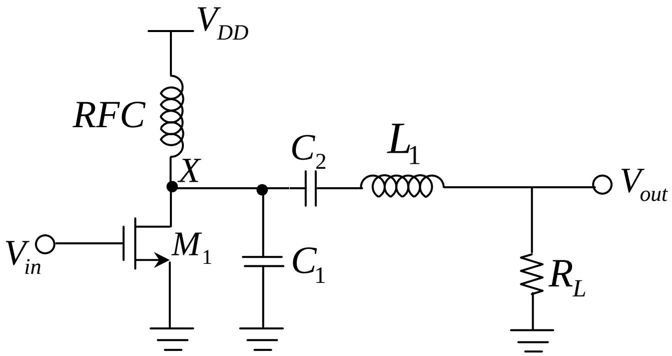
$$max\,\eta_B = \frac{P_{o,max}}{P_{DC}} = \frac{\pi}{4} \cong 0.785$$

- Class B scarifies linearity but gains in terms of efficiency.

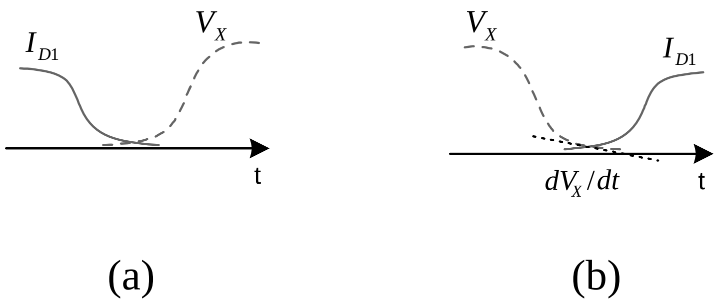
- For practical implementations

$$I_D \cong \frac{i_{peak}}{10}$$

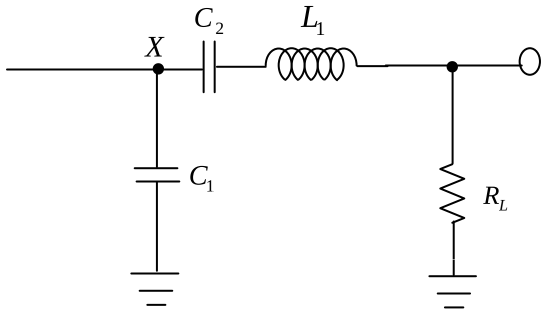
CLASS E AMPLIFIER

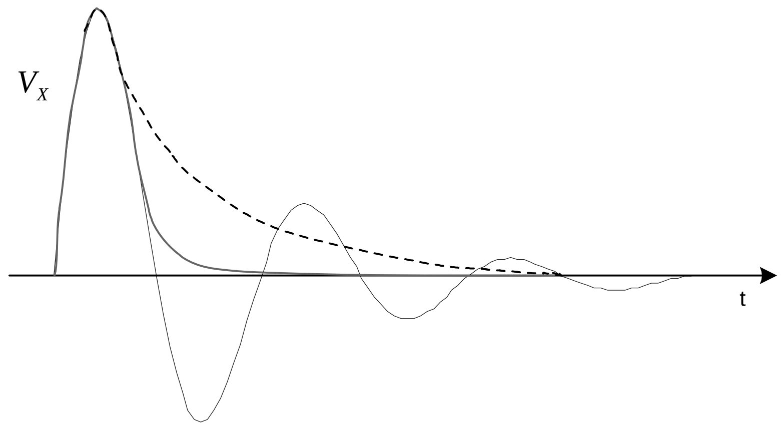


Class E stage.



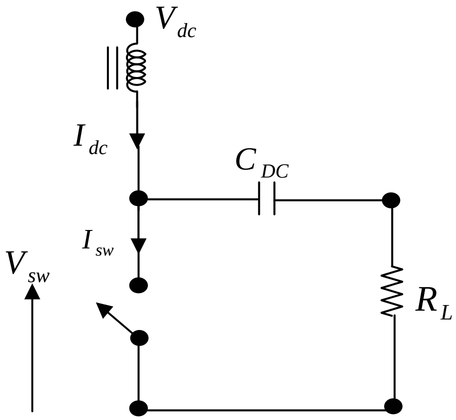
Voltage and current waveforms in a class E stage.



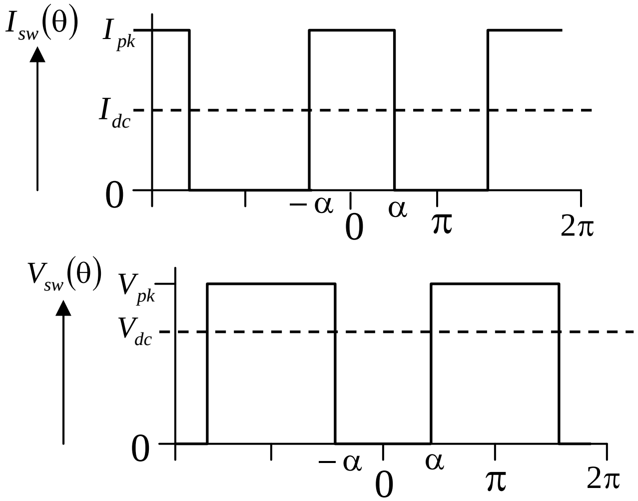


Response of class E stage when the transistor turns off.

BASIC RF SWITCHING AMPLIFIER



Basic RF switching amplifier.



Basic RF switch waveforms.

$$V_{dc} = \frac{1}{2\pi}\int_{-\pi}^{\pi} v_{sw}(\theta)d\theta$$

$$= \frac{1}{\pi}\int_{\alpha}^{\pi} V_{pk}d\theta \qquad (1)$$

$$\frac{V_{dc}}{V_{pk}} = \frac{(\pi - \alpha)}{\pi}$$

The voltage waveform can be considered to be an alternating voltage with zero mean value if it is offset by V_{dc}. So the peak-to-peak current swing will be V_{pk}/R_L and

$$I_{pk} = V_{pk} / R_L \qquad (2)$$

$$I_{dc} = \frac{\alpha}{\pi} I_{pk} \qquad (3)$$

The fundamental even Fourier coefficient of current, I_p is given by

$$I_1 = \frac{1}{\pi} \int_{-\pi}^{\pi} i_{sw}(\theta) cos(\theta) d\theta \qquad (4)$$

where

$$\begin{aligned} i_{sw}(\theta) &= I_{pk}, -\alpha < \theta < \alpha \\ &= 0, -\pi < \alpha, \alpha < \theta < \pi \\ &= \frac{2}{\pi} \int_{o}^{\alpha} I_{pk} cos(\theta) d\theta \qquad (5) \\ \frac{I_1}{I_{pk}} &= \frac{2 sin(\alpha)}{\pi} \end{aligned}$$

Similarly, the fundamental Fourier coefficient for the voltage waveform is

$$\frac{V_1}{V_{pk}} = -\frac{2\,sin(\alpha)}{\pi} \tag{6}$$

Combining (6.1) through (6.6), the RF power, P_{rf} can be expressed in terms of the dc supply terms, V_{dc} and I_{dc}.

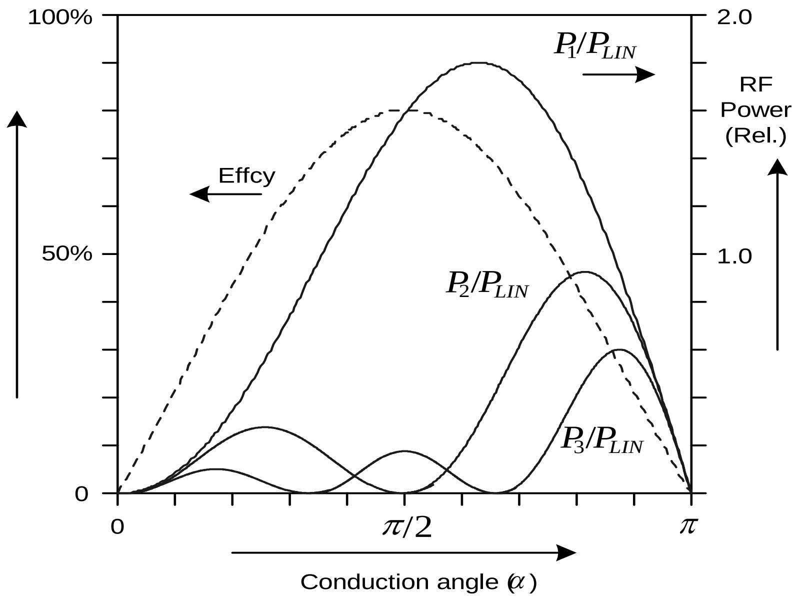
$$\frac{V_1}{V_{dc}} = \frac{2\,sin\,\alpha}{\pi - \alpha}$$

$$\frac{I_1}{I_{dc}} = \frac{2\,sin\,\alpha}{\alpha} \tag{7}$$

$$P_{rf} = V_{dc} I_{dc} \frac{2\,sin^2\alpha}{\alpha(\pi - \alpha)}$$

So that the output efficiency, η, is given by

$$\eta = \frac{2\,sin^2\alpha}{\alpha(\pi - \alpha)} \tag{8}$$



RF output power and efficiency of basic RF switch. Fundamental and harmonic power are expressed relative to P_{lin} the class A RF power having the same peak RF current at the same dc supply voltage.

and defining the linear power, P_{lin} as

$$P_{lin} = \frac{V_{dc} I_{pk}}{4} \tag{9}$$

the relative power is

$$\frac{P_1}{P_{lin}} = \frac{8 \sin^2 \alpha}{\pi(\pi - \alpha)} \tag{10}$$

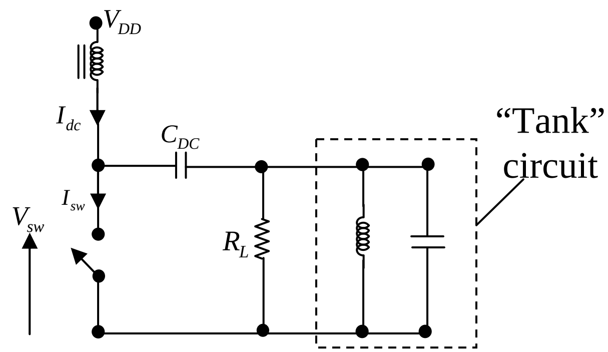
TUNED RF SWITCHING

$$\frac{I_1}{I_{pk}} = \frac{2\sin(\alpha)}{\pi}$$

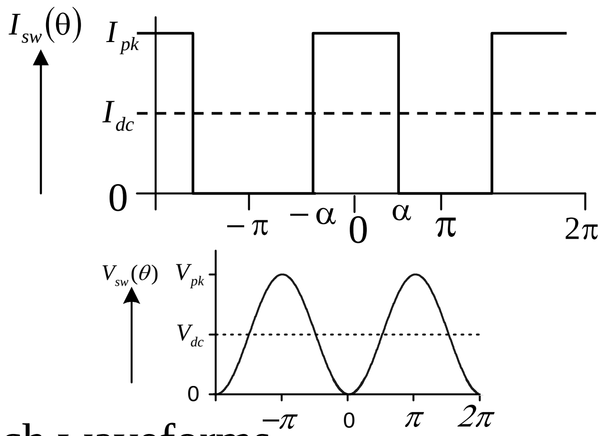
$$\frac{I_{dc}}{I_{pk}} = \frac{\alpha}{\pi}$$

The sinusoidal voltage gives the simple relationship

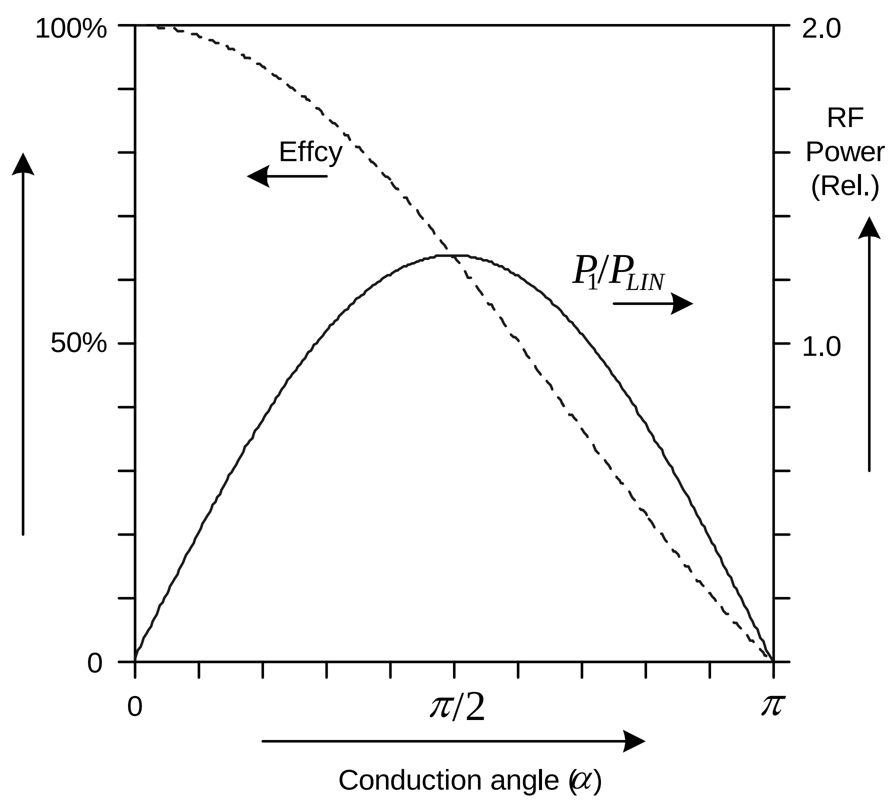
$$V_1 = V_{dc}$$



Tuned RF switching amplifier.



Tuned switch waveforms.



RF output power and efficiency of basic RF switch with harmonic short.

Thus the fundamental RF power is

$$P_1 = I_{dc} V_{dc} \frac{sin(\alpha)}{\alpha} \tag{11}$$

Yielding this output efficiency:

$$\eta = \frac{sin(\alpha)}{\alpha} \tag{12}$$

The relative power can be determined as before, by expressing P_1 (in 11), in terms of the peak current I_{pk} using the relationship of (3):

$$P_1 = I_{pk} V_{dc} \frac{sin(\alpha)}{\pi}$$

So the ratio of fundamental RF power to P_{lin} is becomes

$$\frac{P_1}{P_{lin}} = I_{pk} V_{dc} \frac{4\,sin(\alpha)}{\pi} \left(P_{lin} = \frac{I_{pk} V_{dc}}{4} \right) \tag{13}$$