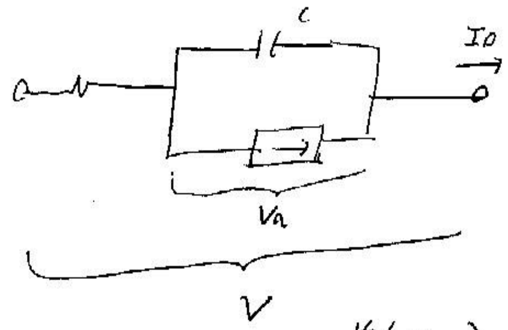
Diode Distortion

$$I_D = I_S\left(e^{qV/KT} - 1\right)$$

$$V_{th} \triangleq \frac{KT}{q}$$

Large signal model



$$I_D = I_S\left(e^{V_a/nV_T} - 1\right)$$

$$e^x = 1 + x + \frac{x^2}{2!} + \frac{x^3}{3!} + \frac{x^4}{4!} \qquad x = \frac{V}{V_T}$$

Grey + Meyer.

Let $V = A_1 \cos \omega_1 t + A_2 \cos \omega_2 t$

$$I_D = I_S\left[\left(\frac{V}{V_T}\right) + \frac{1}{2}\left(\frac{V}{V_T}\right)^2 + \frac{1}{6}\left(\frac{V}{T_T}\right)^3\right]$$

$$= a_1 V + a_2 V^2 + A_3 V^3$$

$$a_1 = \frac{I_S}{V_T} \qquad a_2 = \frac{I_S}{2V_T{}^2} \qquad a_3 = \frac{I_S}{6V_T{}^3}$$

Harmonic distortion.

Let $v = A \sin \omega t$.

$$V_0 = a_1 A \sin\omega t + a_2 A^2 \sin^2 \omega t + a_3 A^3 \sin^3 \omega t + \cdots$$

$$= a_1 A \sin\omega t + \frac{a_2 A^2}{2}\left(1 - \cos 2\omega t\right)$$

$$+ \frac{a_3 A^3}{4}\left(3 \sin\omega t - \sin 3\omega t\right) + \cdots$$

regroup terms.

second harmonic distortion $\triangleq \dfrac{2^{nd} \text{ amplitude}}{\text{fundamental}}$

small distortion. $\frac{3}{4} a_3 A^3 \ll a_1 A$

$$HD_2 = \frac{a_2 A^2}{2} \times \frac{1}{a_1 A} = \frac{a_2}{2 a_1} A$$

$$= \frac{1}{2} \frac{I_S}{2 V_T^2} \frac{V_T}{I_S} A = \boxed{\frac{A}{4 V_T} = HD_2}$$

$\Rightarrow 10\%$ $HD_2 = .1$

$A = .1 \times 4 \times 26mV \cong 10mV$

for third harmonic distortion;

$$HD3 = \frac{a_3 A^3}{4} \frac{1}{a_1 A} = \frac{1}{4} \frac{a_3}{A_1} A^2$$

$$= \frac{1}{24}\left(\frac{A}{V_{th}}\right)^2$$

Intermodulation Distortion in Diodes

HD3 ⇒ for $A = 10\,mV$ ⇒ $HD_3 = .62\%$

now Let $V = A_1 \cos \omega_1 t + A_2 \cos(\omega_2 t)$

Linear term $= a_1 [A_1 \cos \omega_1 t + A_2 \cos(\omega_2 t)$

from the squared term.

$$a_2 (A_1 \cos \omega_1 t + A_2 \cos \omega_2 t)^2$$

$$\left.\begin{array}{l} \frac{a_2}{2} A_1^2 \cos 2\omega_1 t \\ \frac{a_2}{2} A_2^2 \cos 2\omega_2 t \\ a_2 A_1 A_2 \cos[\omega_1 t - \omega_2 t] \\ a_2 A_1 A_2 \cos[(\omega_1 + \omega_2)t] \end{array}\right\} 2^{nd} \text{ order terms}$$

$$a_3 A_1 A_2 [\cos(\omega_1 + \omega_2)t + \cos(\omega_1 - \omega_2)t] [\underbrace{A_1 \cos \omega_1 t}_{①} + \underbrace{A_2 \cos \omega_2 t}_{②}$$

① $\frac{a_3 A_1^2 A_2}{2} [\cos \omega_2 t + \cos(2\omega_1 + \omega_2)t + \cos(-\omega_2 t) + \cos(2\omega_1 - \omega_2)t]$

① $\frac{a_3 A_1^2 A_2}{2} [2 \cos \omega_2 t + \cos(2\omega_1 + \omega_2)t + \cos(2\omega_1 - \omega_2)t]$

② $\frac{a_3 A_1 A_2^2}{2} [\cos \omega_1 t + \cos(\omega_1 + 2\omega_2)t + \cos \omega_1 t + \cos(\omega_1 - 2\omega_2)t]$

② $\frac{a_3 A_1 A_2^2}{2} [2 \cos \omega_2 t + \cos(\omega_1 + 2\omega_2)t + \cos(\omega_1 - 2\omega_2)t]$

- 3rd order terms $\propto A^3$
  $A_1 = A_2$ ⇒

Harmonic Distortion in a BJT

apply input signal voltage $V_s = V_i - V_{BE1}$

$$V_o = -R_L \left(I_s \, e^{\frac{V_s + V_{BE1}}{V_T}} - I_Q \right)$$

$$= -R_L \left(I_s \, e^{\frac{V_{BE1}}{V_T}} \, e^{\frac{V_s}{V_T}} - I_Q \right)$$

but $I_Q = I_s \, e^{\frac{V_{BE1}}{V_T}}$

so $V_o = -R_L I_Q \left(e^{\frac{V_s}{V_T}} - 1 \right)$

$$V_o = -R_L I_Q \left[\frac{V_s}{V_T} + \frac{1}{2}\left(\frac{V_s}{V_T}\right)^2 + \frac{1}{6}\left(\frac{V_s}{V_T}\right)^2 + \cdots \right]$$

$$= a_1 V_s + a_2 V_s^2 + a_3 V_s^3$$

$$a_1 = \frac{-R_L I_Q}{V_T}$$

$$a_2 = \frac{-R_L I_Q}{2 V_T^2}$$

$$a_3 = \frac{-R_L I_Q}{6 V_T^3}$$

$$\frac{V_s}{V_T} \ll 1$$

Let $V_s = \hat{V}_s \sin \omega t$

$$V_o = a_1 \hat{V}_s \sin \omega t + a_2 \hat{V}_s^2 \sin^2 \omega t + a_3 \hat{V}_s^3 \sin^3 \omega t + \cdots$$

$$= a_1 \hat{V}_s \sin \omega t + \frac{a_2 \hat{V}_s^2}{2} (1 - \cos 2\omega t) + \frac{a_3 \hat{V}_s^3}{4} (3 \sin \omega t - \sin 3\omega t)$$

$$HD_2 = \frac{a_2 \hat{V}_s^2}{2} \frac{1}{a_1 \hat{V}_s} = \frac{1}{2} \frac{a_2}{a_1} \hat{V}_s = \frac{1}{4} \frac{\hat{V}_s}{V_{th}}$$

$$HD_3 = \frac{a_3 \hat{V}_s^3}{4} \frac{1}{a_1 \hat{V}_s} = \frac{1}{4} \frac{a_3}{a_1} \hat{V}_s^2 = \frac{1}{24} \left(\frac{\hat{V}_s}{V_T}\right)^2$$

example Let $\hat{V}_o = .6\,V$

$I_Q = 1.86\,ma.$

$R_L = 1K\Omega$

$$A_V = g_m R_L = \frac{q I_C}{KT} R_L$$

Quessient gain. $$A_{VQ} = \frac{1.86}{26} \times 1000 = 70.6$$

$$\Rightarrow \hat{V}_s = \frac{V_o}{70.6} = \frac{600}{70.6} mV = 8.5\,mV$$

$$HD_2 = \frac{1}{4} \frac{8.5}{26} = .082$$

$$HD_3 = \frac{1}{24} \left(\frac{8.5}{26}\right)^2 = .0045$$

positive gain about Q point. peak output current.

$$I_Q + \hat{I}_o = (1.86 + .6)\,ma = 2.46\,ma.$$

$$A_v' = \frac{2.46}{26} \times 1000 = 94.6$$ ⇒ 34 % increase over A_Q

negative gain

$$I_Q - \hat{I}_o = 1.86 - .6 = 1.26\,ma.$$

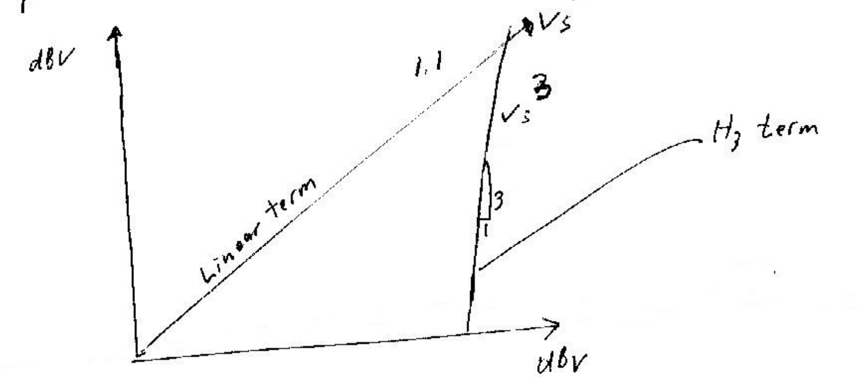
$$A_v'' = \frac{1.26}{26} \times 1000 = 48.5$$ decrease 31 % over A_Q

nonlinear gain.

$$HD_2 = \frac{a_2 \hat{V}_s^2}{2} \frac{1}{a_1 \hat{V}_s} = \frac{1}{2} \frac{a_2}{a_1} \hat{V}_s = \frac{1}{4} \frac{\hat{V}_s}{V_T}$$

$$HD_3 = \frac{a_3 \hat{V}_s^3}{4} \frac{1}{a_1 \hat{V}_s} = \frac{1}{4} \frac{a_3}{a_1} \hat{V}_s^2 = \frac{1}{24} \left(\frac{\hat{V}_s}{V_T} \right)^2.$$

plot Log Vin vs Log Vout



See the following for discussions on the Third Order Intercept Point.

16.5 SIGNAL DISTORTION DUE TO INTERMODULATION PRODUCTS

Operating an amplifier under large-signal conditions causes distortions in the output signal. This distortion is primarily caused by deviation from linear operation, which causes new frequencies to appear at the output port, usually referred to as "intermodulation products."

DEFINITION-INTERMODULATION PRODUCTS: *The additional frequencies at the output of a nonlinear amplifier (or in general any nonlinear network) when two or more sinusoidal signals are applied at the input.*

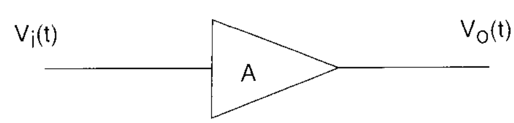
To illustrate this concept, let's consider the following input signal consisting of two frequencies, each with unity amplitude:

$$V_i(t) = \cos(2\pi f_1 t) + \cos(2\pi f_2 t) \tag{16.22}$$

If $V_i(t)$ is applied to a nonlinear amplifier (see Figure 16.22) with the output/input voltage characteristic of:

$$V_o(t) = AV_i(t) + BV_i^2(t) + CV_i^3(t) \tag{16.23}$$

FIGURE 16.22 Nonlinear amplifier.



Then, the output signal $V_o(t)$ will contain not only the original frequencies f_1 and f_2, but also the following intermodulation products: DC, $2f_1$, $2f_2$, $3f_1$, $3f_2$, $f_1 \pm f_2$, $2f_1 \pm f_2$, and $2f_2 \pm f_1$.

We may classify these intermodulation products as follows:

Second harmonics: $2f_1, 2f_2$ (caused by V_i^2 term).

Third harmonics: $3f_1, 3f_2$ (caused by V_i^3 term).
Second-order intermodulation products: $f_1 \pm f_2$ (caused by V_i^2 term).
Third-order intermodulation products: $2f_1 \pm f_2, 2f_2 \pm f_1$ (caused by V_i^3 term).

These are plotted on the frequency scale along with the original frequencies, as shown in Figure 16.23. From this figure, we can see that all additional frequencies can be filtered out except the intermodulation products $2f_1 - f_2$ and $2f_2 - f_1$, which are very close to f_1 and f_2 and fall within the amplifier bandwidth and cannot be filtered out. Thus, they are capable of signal distortions at the output.

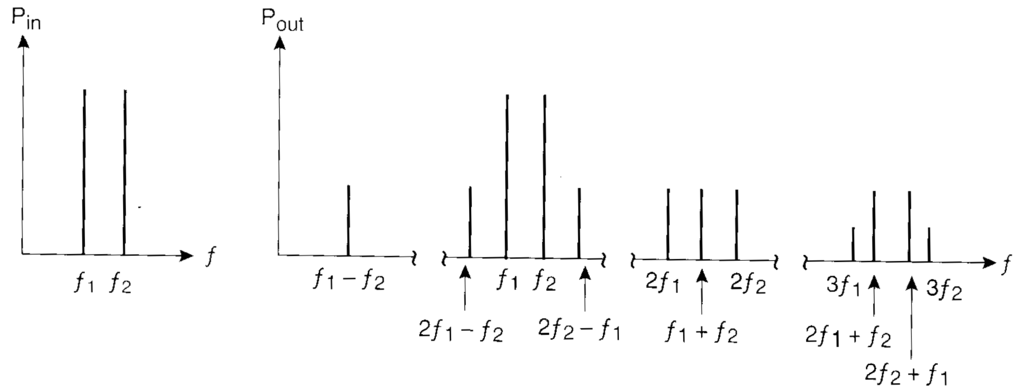


FIGURE 16.23 Input and output power spectrum.

Third-order two-tone intermodulation products $(2f_1 - f_2)$ and $(2f_2 - f_1)$ have special importance because they set the upper limit on the dynamic range or bandwidth of the amplifier.

A measure of the second- or third-order intermodulation distortion is given by two theoretical intercept points, as shown in Figure 16.24. As can be seen from Figure 16.24, the third-order product has a lower intercept point than the second-order product and thus is more significant in distortion analysis.

If the third-order product output power is measured versus the input power, then the third-order intercept (TOI) point can be theoretically obtained, as shown in Figure 16.25. The higher the value of power at TOI (P_{TOI} or P_{IP}), the larger the dynamic range of the amplifier will be.

The power at the third-order intercept point can be theoretically and experimentally obtained to be approximately given by:

$$P_{IP}(\text{dBm}) \approx P_{1\text{dB}}(\text{dBm}) + 10(\text{dB}) \qquad (16.24)$$

Furthermore, the difference between the two curves $(P_{f1} - P_{2f1-f2})$ is a variable quantity and is maximum at $P_{o,mds}$ and zero at P_{IP}. It can be shown that:

$$P_{f_1} - P_{2f_1 - f_2} = \frac{2}{3}(P_{IP} - P_{2f_1 - f_2}) \ \ (\text{dBm}) \qquad (16.25)$$

We now define the "spurious free dynamic range" (DR_f) to be the difference between two powers $P_{f_1} - P_{2f_1 - f_2}$ when the third-order intermodulation product is equal to the minimum detectable signal. That is:

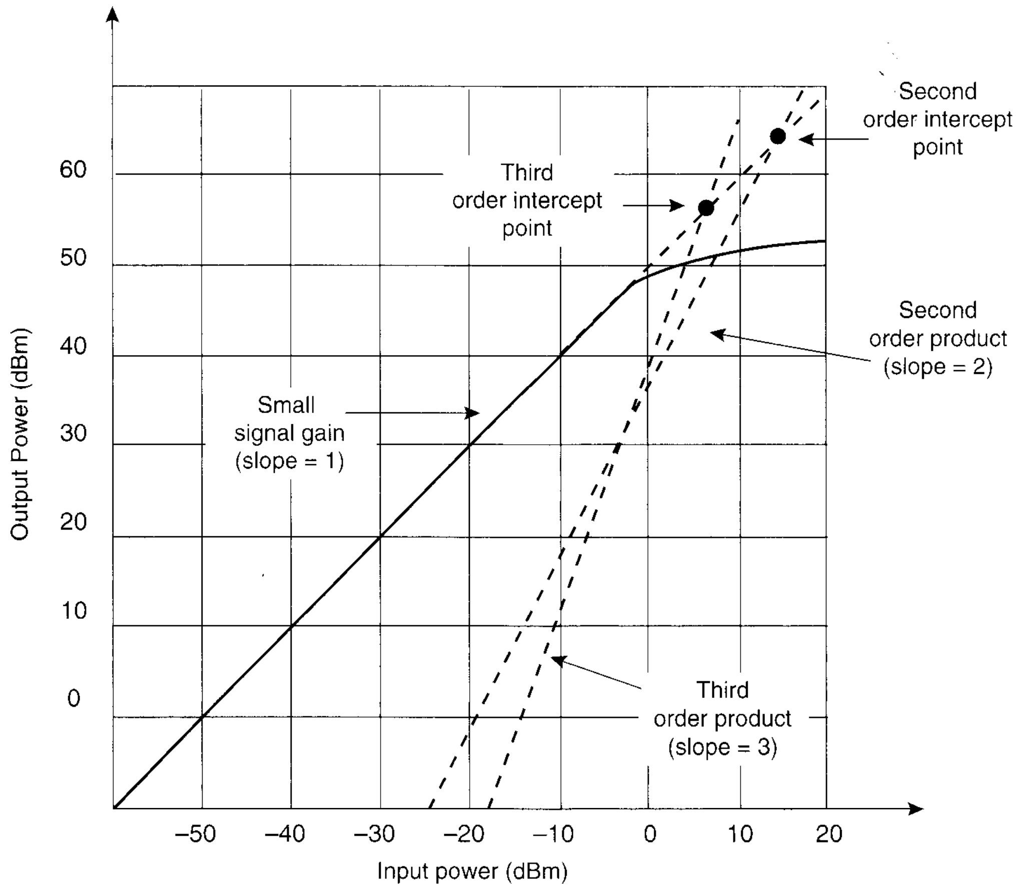


FIGURE 16.24 Second and third order intercept points.

$$DR_f = (P_{f_1} - P_{2f_1-f_2}) \quad \text{when} \quad (P_{2f_1-f_2} = P_{o,mds})$$

Thus, we can write DR_f as:

$$DR_f = (P_{f_1} - P_{o,mds}) = \frac{2}{3}(P_{IP} - P_{o,mds}) \quad \text{(dBm)} \tag{16.26}$$

where from Equation 16.10,

$$P_{o,mds}(\text{dBm}) = -174 \text{ dBm} + 10\log_{10}\text{B} + G_A(\text{dB}) + F(\text{dB}) + X(\text{dB})$$

EXAMPLE 16.7

Calculate dynamic range (DR) and spurious free dynamic range (DR_f) for a microwave high-power/broadband amplifier that has a gain of 20 dB, a noise figure of 5 dB, a bandwidth of 250 MHz and can deliver a power of P_{1dB} = 30 dBm (assume X = 3 dB).

Solution:

$$P_{o,mds}(\text{dBm}) = -174 \text{ dBm} + 10\log_{10}B + G_A(\text{dB}) + F(\text{dB}) + X(\text{dB})$$

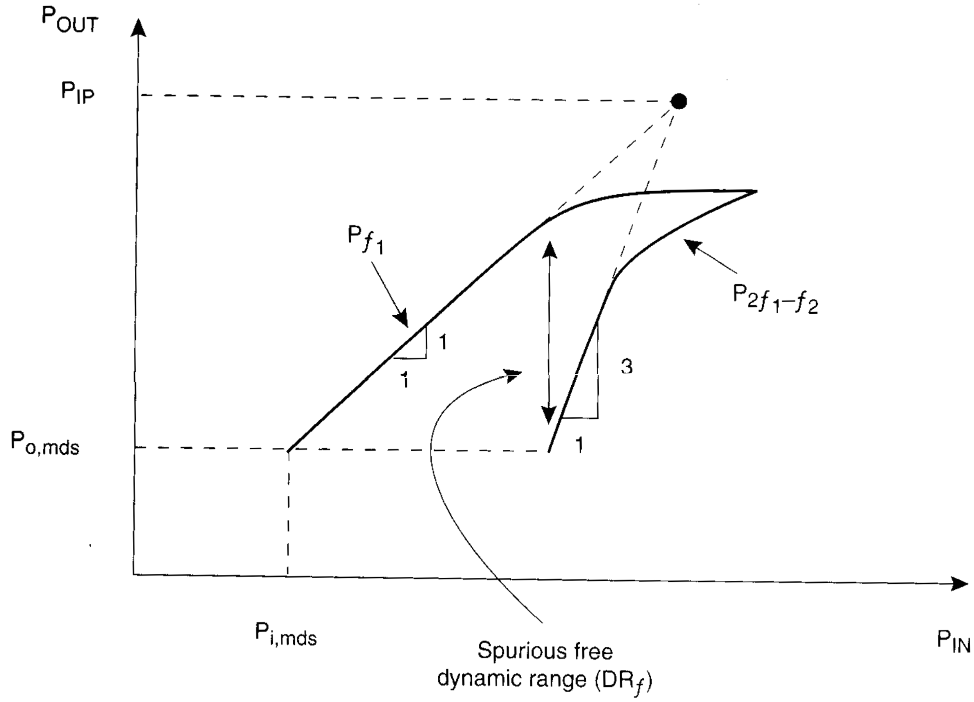


FIGURE 16.25 Third-order intercept point.

$$= -174 + 10\log_{10}(250 \times 10^6) + 5 + 20 + 3 = -62 \text{ dBm}$$

$$DR = 30 - (-62) = 92 \text{ dB}$$

$$DR_f = (2/3)(30 + 10 + 62) = 68 \text{ dB}$$

Two-Tone Measurement Technique. Third-order intercept power (P_{IP}) is a figure of merit for intermodulation product suppression. A high intercept point is a good indicator and signifies a high suppression of undesired intermodulation products. An experimental method for finding P_{IP} is by the use of a technique called "two-tone measurement technique."

In this technique, two signals of close but different frequencies that have equal magnitude are applied to the input of the amplifier, as shown in Figure 16.23. Using a spectrum analyzer, the outputs are examined (see Figure 16.23), and from a simple measurement of the difference in power between the main output (P_{f_1} in dBm) and the third-order intermodulation product ($P_{2f_1-f_2}$ in dBm), we can obtain P_{IP} (in dBm). To find P_{IP}, first let's define:

$$\Delta = P_{f_1} - P_{2f_1-f_2} \text{ (dB)} \tag{16.27a}$$

Then, substituting for $P_{2f_1-f_2}$ from Equation 16.27a in 16.25, we can write:

$$\Delta = \frac{2}{3}(P_{IP} - P_{2f_1 - f_2}) = \frac{2}{3}[P_{IP} - (P_{f_1} - \Delta)] \quad \text{(dB)} \tag{16.27b}$$

By rearranging terms in Equation 16.27b we obtain $(P_{IP})_3$ for the third-order harmonic as :

$$(P_{IP})_3 = P_{f_1} + \frac{\Delta}{2} \quad \text{(dBm)} \tag{16.28a}$$

Thus, by halving the difference (in dB) between the main output and one of the third-order intermodulation products and adding it to the main output, we can obtain the third-order intercept point (in dBm) as in dicated by Equation 16.28a.

EXAMPLE 16.8

If through measurement we find that P_{f_1} = 8 dBm and Δ = 40 dB for the third-order intermodulation product, what is the power at the third-order intercept point?

$(P_{IP})_3$ can easily be calculated to be:

$$(P_{IP})_3 = 8 + 40/2 = 28 \text{ dBm}$$

NOTE: *It can be shown that in general, for the n^{th} order intermodulation product ($n \neq 1$), Equation 16.28a can be generalized as:*

$$(P_{IP})_n = P_{f_1} + \frac{\Delta}{n-1} \quad \text{(dBm)} \tag{16.28b}$$

where Δ is the difference between the fundamental harmonic power and the undesired n^{th} intermodulation product power.

16.6 MULTISTAGE AMPLIFIERS: LARGE-SIGNAL DESIGN

As discussed in the last chapter, most practical transistor amplifiers usually consist of a number of stages connected in cascade forming a multistage amplifier. In a high-power amplifier, each stage should be designed for operation at maximum power such that the maximum power transfer condition is met. In the next several sections, we will present a detailed analysis of a multistage high-power amplifier.

16.6.1 Analysis

Consider a general N-stage amplifier configuration, as shown in Figure 16.26. To have a stable amplifier, the stability of the individual stages as well as overall stability must be checked.

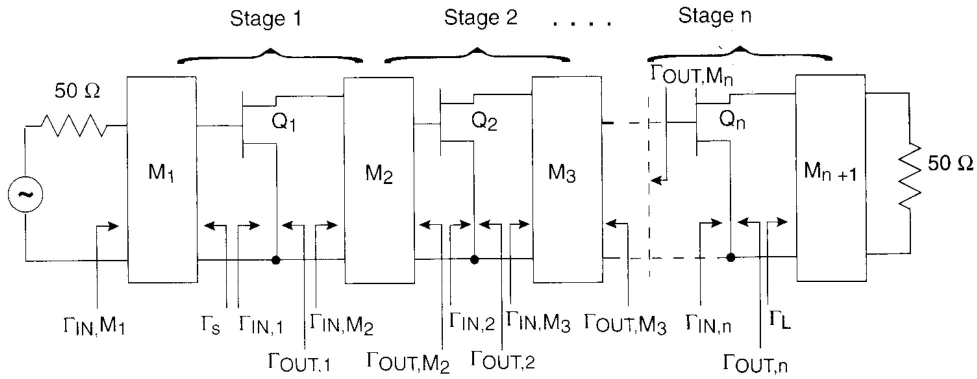


FIGURE 16.26 N-stage FET amplifier configuration.

In this type of amplifier, the goal is to produce the overall highest possible power. Thus, each stage must operate at or close to its 1-dB gain compression point under large-signal conditions. This means that using power contours we need to select Γ_{LP} at the point where $P_{OUT} = P_{max}$, (which is at the output port of each transistor) and then use conjugate matched condition for the input port to minimize *VSWR* and create the maximum power transfer condition, i.e.,

$$\Gamma_S = (\Gamma_{IN,1})^* \quad (16.29a)$$

$$\Gamma_{IN,M2} = \Gamma_{LP,1} \quad (16.29b)$$

$$\Gamma_{OUT,M2} = (\Gamma_{IN,2})^* \quad (16.29c)$$

$$\vdots$$

$$\vdots$$

$$\Gamma_{OUT,Mn} = (\Gamma_{IN,n})^* \quad (16.29d)$$

$$\Gamma_L = \Gamma_{LP,n} \quad (16.29e)$$

where $\Gamma_{LP,1}$, $\Gamma_{LP,2}....\Gamma_{LP,n}$ represent points on the power contours where P_{max} occurs for transistors $Q_1, Q_2, ...Q_n$, respectively.

16.6.2 Overall Third-Order Intercept Point Power

High-power amplifiers are designed not only to obtain large amounts of output power but also to have a high third-order intercept point (*TOI*). If each individual stage has a known power value at *TOI* (P_{TOI}), then assuming in-phase addition (of the P_{TOI} of each stage), overall P_{TOI} of the multistage power amplifier is given by:

$$\frac{1}{P_{TOI}} = \frac{1}{P_{TOI,n}} + \frac{1}{G_{Pn}P_{TOI,n-1}} + \dots + \frac{1}{G_{Pn}G_{Pn-1}\dots G_{P2}P_{TOI,1}} \quad (16.30)$$

where G_P is the power gain.

If all stages are identical, i.e.,

$$G_{Pk} = G_P,\ k = 1,2,\ldots,n \quad (16.31a)$$

$$P_{TOI,k} = P,\ k = 1,2,\ldots,n \quad (16.31b)$$

Then, Equation 16.22 reduces to:

$$\frac{1}{P_{TOI}} = \frac{1}{P}\left(1 + \frac{1}{G_P} + \frac{1}{G_P^2} + \ldots + \frac{1}{G_P^{n-1}}\right) \quad (16.32)$$

Using Equation 15.33, the geometric series identity, Equation 16.32 becomes:

$$\frac{1}{P_{TOI}} = \frac{1}{P}\left(\frac{1 - 1/G_P^n}{1 - 1/G_P}\right) \quad (16.33)$$

For an infinite chain of amplifier stages (i.e., $n \to \infty$), we can write Equation 16.32 as:

$$P_{TOI} = P(1 - 1/G_P) \quad (16.34)$$

In practice, n may be large (but is not infinite), thus Equation 16.34 gives a best case scenario for the amplifier's overall power at the third-order intercept point (P_{TOI}), which is a power amplifier's figure of merit—very similar to noise measure (M), which is a figure of merit for an LNA as discussed earlier.

16.6.3 Dynamic Range Considerations

As discussed earlier, the dynamic range of an amplifier is bound at the lower end by noise considerations ($P_{o,mds}$) and at the upper end by 1-dB gain compression point ($P_{1\text{dB}}$). Thus, for an n-stage amplifier, we can write:

a) Lower Limit of Dynamic Range.

$$(P_{o,mds})_{cas} = KT(\text{dBm}) + 10\log_{10}B + 10\log_{10}F_{cas} + G_{A,tot}(\text{dB}) + X(\text{dBm}) \quad (16.35)$$

where (see Equation 15.28):

$$G_{A,tot} = G_{A1}G_{A2}\ \ldots\ G_{An}, \quad (16.36)$$

$$F_{cas} = F_1 + \frac{F_2 - 1}{G_{A1}} + \frac{F_3 - 1}{G_{A1}G_{A2}} + \ldots + \frac{F_n - 1}{G_{A1}G_{A2}\ldots G_{An-1}} \quad (16.37)$$

Special Case: Identical Amplifiers For identical amplifiers and n very large ($n \to \infty$), Equations 16.36 and 16.37 simplify as:

$$G_{Ak} = G_A, \quad k = 1,2,\ldots,n \quad (16.38a)$$

$$F_k = F, \quad k = 1,2,\ldots,n \quad (16.38b)$$

$$G_{A,tot} = (G_A)^n \quad (16.39)$$

$$F_{cas} = 1 + \frac{F - 1}{1 - 1/G_A} = 1 + M \quad (16.40)$$

$$(P_{o,mds})_{cas} = KT(\text{dBm}) + 10\log_{10}B + 10\log_{10}(1 + M) + nG_A(\text{dB}) + X(\text{dB})\ \ (\text{dBm}) \quad (16.41)$$

where (see Equation 15.36):

$$M = \frac{F-1}{1-1/G_A} \tag{16.42}$$

NOTE: *$(P_{o,mds})_1$ for the first stage is given by:*

$$(P_{o,mds})_1(\text{dBm}) = KT(\text{dBm}) + 10\log_{10}B + 10\log_{10}F + G_A(\text{dB}) + X\,(\text{dB}) \tag{16.43}$$

Combining Equations 16.41 and 16.43 we can see that:

$$(P_{o,mds})_{cas}(\text{dBm}) = (P_{o,mds})_1(\text{dBm}) + \Delta P_{o,n} \tag{16.44a}$$

where

$$\Delta P_{o,n} = 10\log_{10}\left(\frac{1+M}{F}\right) + (n-1)G_A \quad (\text{dB}) \tag{16.44b}$$

Because $\Delta P_{o,n}$ is always positive, thus we can write:

$$(P_{o,mds})_{cas} \geq (P_{o,mds})_1 \tag{16.45}$$

Equation 16.45 shows an important consideration where the output minimum detectable signal for the whole cascade $(P_{o,mds})_{cas}$ is determined by and depends greatly on the minimum detectable signal of the first stage of the cascade, $(P_{o,mds})_1$. Thus, it is important to have the first stage operate at the lowest possible output noise level.

b) Upper Limit of Dynamic Range. At the upper limit, the total output power at 1-dB gain compression point ($P_{1\text{dB},cas}$) can be shown to be of similar form to Equation 16.30:

$$\frac{1}{P_{1\text{dB},cas}} = \frac{1}{P_{1\text{dB},n}} + \frac{1}{G_{Pn}P_{1\text{dB},n-1}} + \ldots + \frac{1}{G_{Pn}G_{Pn-1}\cdots G_{P2}P_{1\text{dB},1}} \tag{16.46}$$

If all stages are identical, i.e.,

$$G_{Pk} = G_P,\ k = 1,2,\ldots,n \tag{16.47a}$$

$$P_{1\text{dB},k} = P_{1\text{dB}},\ k = 1,2,\ldots,n \tag{16.47b}$$

then Equation 16.46 reduces to:

$$\frac{1}{P_{1\text{dB},cas}} = \frac{1}{P_{1\text{dB}}}\left(1 + \frac{1}{G_P} + \frac{1}{G_P^2} + \ldots + \frac{1}{G_P^{n-1}}\right) \tag{16.48}$$

Using Equation 15.33, the geometric series identity, Equation 16.48 becomes:

$$\frac{1}{P_{1\text{dB},cas}} = \frac{1}{P_{1\text{dB}}}\left(\frac{1-1/G_P^n}{1-1/G_P}\right) \tag{16.49}$$

For an infinite and idenfical chain of amplifier stages (i.e., $n\to\infty$) and very similar to Equation 16.34, we can write Equation 16.49 as:

$$P_{1\text{dB},cas} = P_{1\text{dB}}(1 - 1/G_P) \tag{16.50}$$

NOTE: *Because $G_P \geq 1$ and $P_{1dB,n} = P_{1dB}$, we can see from Equation 16.50 that at all times:*

$$P_{1\text{dB},cas} \leq P_{1\text{dB},n} \tag{16.51}$$

That is, the signal at the output of the cascade is below or at the 1 dB gain compression point of the last-stage amplifier.

As can be seen from Equation 16.51, the power output at 1 dB gain compression point for the whole cascade ($P_{1\text{dB},cas}$) is limited by the last stage 1 dB gain compression point power capability ($P_{1\text{dB},n}$). Thus, we can conclude the following:

CONCLUSION: *It is crucial to have the last stage of the cascade be designed such that it has the highest power handling capability.*

16.6.4 Wide Dynamic Range Multi-stage Amplifier Design

For an n-stage amplifier, the actual dynamic range (DR) is given by:

$$DR = P_{1\text{dB},cas}(\text{dBm}) - (P_{o,mds})_{cas}(\text{dBm}) \tag{16.52}$$

As noted from the previous section, however, for a wide dynamic range design we need to have the following considerations firmly in place:

a. From Equation 16.45 it can be concluded that the first stage sets the lower limit. Therefore, ideally we would like to have:

$$(P_{o,mds})_{cas} \approx (P_{o,mds})_1 \tag{16.53}$$

b. From Equation 16.51, we can observe that the upper limit is determined by the 1 dB gain compression point of the last stage and ideally, we would like to have:

$$P_{1\text{dB},cas} \approx P_{1\text{dB},n} \tag{16.54}$$

Thus, the maximum dynamic range or the **best estimate** of dynamic range (DR_{max}) that we can hope for, can be written as:

$$DR_{max} = P_{1\text{dB},n}(\text{dBm}) - (P_{o,mds})_1(\text{dBm}) \tag{16.55a}$$

Thus, for n identical amplifiers ($n \to \infty$):

$$DR_{max} - DR = \Delta P_{o,n} + P_{1\text{dB}}/G_P \tag{16.55b}$$

POINT OF CAUTION: *From Equations 16.44 and 16.45 we can observe that increasing the number of stages (n) and/or available gain (G_A) of each stage will increase the overall gain but will reduce the effective dynamic range by increasing $(P_{o,mds})_{cas}$. Thus, there is a trade-off between the overall gain and the dynamic range of a multistage amplifier.*

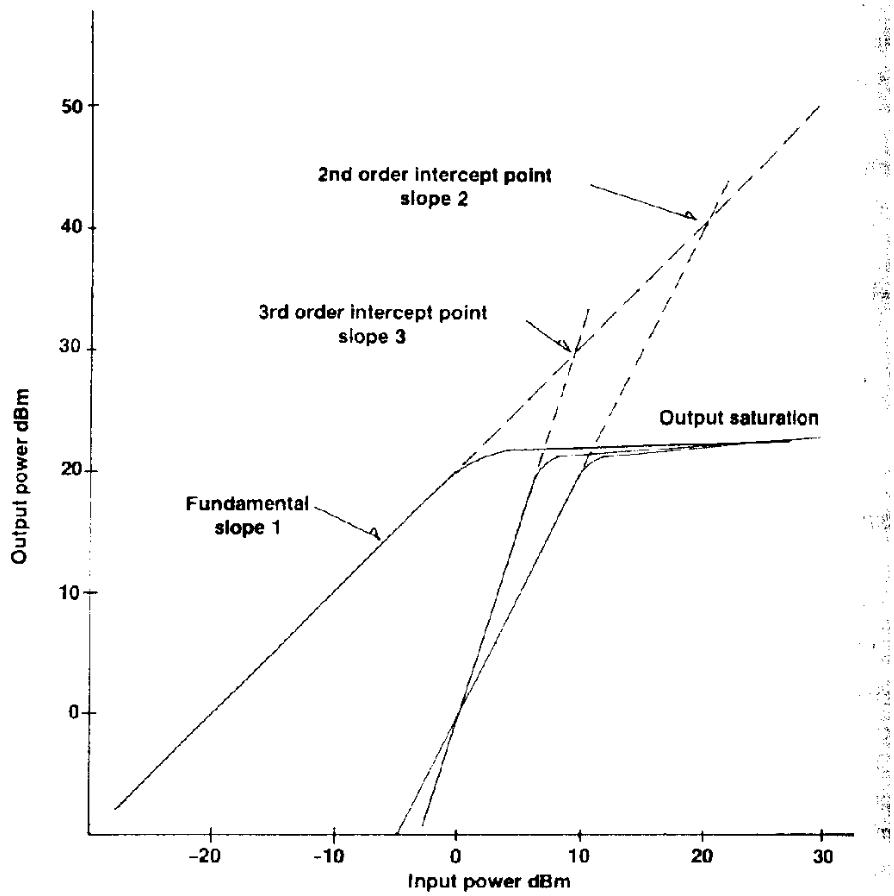


Fig. 4-21. Device outputs showing the fundamental and 2nd and 3rd order distortion products together with the extrapolated respective intercept points.

4.21 SPURIOUS FREE DYNAMIC RANGE

It is informative to determine the dynamic signal operation range of a system which is free of spurious signals resulting from third order intermodulation products. In this definition spurious free means that these spurious signals are equal to the noise level. The third order intermodulation product signal levels are related to the level of the two signals (producing them, respectively, on a three for one dB relationship). The spurious free dynamic range is related to the third order system intercept point by the following relationship:

$$SFDR\ (\text{dB}) = 0.67\ (\overset{3}{I} - kT/\text{MHz} - 10\log B - NF) \qquad (4\text{-}57)$$

where

SFDR is the spurious free dynamic range

$\overset{3}{I}_i$ is the system third order input intercept point
kT is the thermal noise level in a 1 MHz bandwidth
= −114 dBm / MHz
B is bandwidth in MHz
NF is the system noise figure (dB)

An example of the application of this relationship follows.

Example 4-11:

Given:

Third order input intercept point = 10 dBm (Note: Given the output intercept point, the input intercept point is the output intercept point minus the gain in dB notation.)
Bandwidth is 10 kHz
Noise figure is 5 dB

Find the spurious free dynamic range.

Solution:

SFDR = 0.67 (10 − (−114) − (−20) − 5)
= 92.67 dB

4.22 IMAGES

In the mixing process it was shown that when two input signals comprised of an RF signal and a local oscillator signal are applied to a mixer, intermediate frequency signals are produced at the mixer's output. More specifically:

$$F_{if} = \left| NF_r \pm MF_{lo} \right| \qquad (4\text{-}58)$$

where

F_{if} is the intermediate frequency
F_r is the receive frequency
F_{lo} is the local oscillator frequency
M and N are intergers

The primary or desired outputs result when $M = N = 1$

Then equation (4-58) results in:

$$F_{if} = \left|F_r \pm F_{lo}\right| \tag{4-59}$$

Since F_{if} is a constant, then for a given LO frequency there exist two values of F_r which satisfy the relationship.

EXAMPLE 4-13:

Let F_{lo} = 160.7 MHz
F_{if} = 10.7 MHz
then $F_r = \left|F_{if} \pm F_{lo}\right|$
or $\left|10.7 \pm 160.7\right|$ = 150 and 171.4 MHz.

Thus, the mixer is equally responsive to two frequencies, both of which are twice the IF apart. Of these, one is the desired response and the other is called the image frequency. The receiver must reject the image term to provide satisfactory performance. This is one of the reasons mixers are always preceded by preselection filtering of some form. The selectivity requirements of the preselector are governed by the frequency separation between the image frequency and the desired frequency. The amount of image frequency rejection is strictly a function of the attenuation, provided by the preselector filter alone (unless an image rejection mixer is used).

To avoid extreme selectivity requirements from the preselector, the ratio of F_r to F_{if} should not exceed 10 or 20 to 1 for a first conversion in a down conversion superheterodyne receiver.

In up-conversion systems, the high value of intermediate frequency effectively removes the image frequency out of the preselector bandwidth. Here, simple fixed tuned filters often suffice as preselector filters, eliminating tracking and tuning problems. Fig. (4-27) is an example illustrating the relative image frequency behavior between up and down conversion.