STABILITY OF CAPACITIVELY-LOADED EMITTER FOLLOWERS — A SIMPLIFIED APPROACH

Emitter followers with a capacitive load often oscillate when driven from an inductive source. The following analysis shows that simple adjustment in bias current will often stabilize the circuit. The analysis is rigorous, yet provides a much simpler result than is known to have been published.

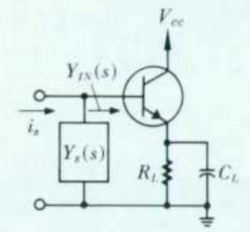

Fig. 1. *Emitter follower with a capacitive load.*

Common-collector transistor amplifiers, or emitter followers, are useful in electronic systems because they have both high input impedance and low output impedance. However, self-sustained oscillations may occur if the load on an emitter follower is capacitive and the impedance of the source which is driving the emitter follower is inductive.

In this article, stability criteria are derived which are useful in the design of capacitively-loaded emitter followers. The approach taken is that of examining the even or real part of the driving-point admittance and determining conditions for stable operation. Stability criteria for the capacitively-loaded emitter follower have, of course, been published before, but the analyses have not resulted in simple, easily-applied results. The following derivation is rigorous, yet gives a much more easily-applied result than is known to have been presented previously. The result shows, in fact, that a simple adjustment in dc emitter current can eliminate emitter-follower instability in many cases.

EQUIVALENT CIRCUIT

An emitter follower with a capacitive load is shown schematically in Fig. 1. Fig. 2 shows a very complete equivalent circuit for the transistor.

Fig. 3 is a transistor equivalent circuit which is valid for frequencies up to $\omega_T/3$, where ω_T is the transistor short-circuit current gain-bandwidth product in the common-emitter configuration. This circuit is a good compromise between simplicity and accuracy, and is adequate for most design work.[1]

Using the transistor equivalent circuit of Fig. 3, the circuit of Fig. 1 reduces to that shown in Fig. 4.

EMITTER FOLLOWER STABILITY

Conditions for the stability of the circuit of Fig. 4 can be found by examining the real part of the driving-point admittance $Y_{IN}(s)$. For this analysis the transverse base resistance $r_{b'}$ can be neglected, so the real part of $Y_{IN}(s)$ can be considered to be the same as the real part of the admittance $Y'_{IN}(s)$ in Fig. 4.

Solving for $Y'_{IN}(s)$ gives

$$Y'_{IN}(s) = \frac{i(s)}{v_{b'}(s)} = \frac{1}{Z_E + Z_L\left(1 + \frac{\alpha_o Z_E}{r_\epsilon}\right)} \qquad (1)$$

where $$Z_E = \frac{r_{b'e}}{r_{b'e}C_E s + 1}$$

and $$Z_L = \frac{R_L}{R_L C_L\, s + 1}$$

(concluded inside on p. 15)

[1]For a thorough discussion of transistor equivalent circuits, see M. S. Ghausi, "Principles and Design of Linear Active Circuits," McGraw-Hill Book Co., Inc., New York, N.Y., 1965.

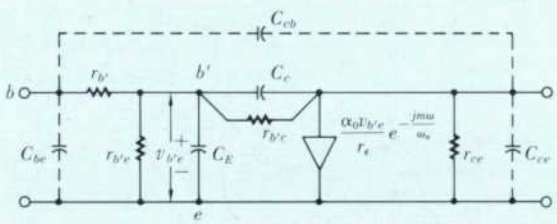

Fig. 2. *Transistor hybrid-pi equivalent circuit.*

$r_{b'c}$, r_{ce} — account for base-width modulation effects
$r_{b'}$ — transverse base resistance
r_ϵ — dynamic emitter resistance
$r_{b'e} = \frac{r_\epsilon}{1-\alpha_0} = r_\epsilon\,(\beta_0 + 1)$
α_0 — low frequency transistor current transfer ratio for common base configuration
β_0 — low frequency transistor current transfer ratio for common emitter configuration
C_{be}, C_{cb}, C_{ce} — header capacitances
C_c — collector-base junction capacitance
$C_E \cong \frac{1}{\omega_T r_\epsilon} - C_{ob}$
ω_T — transistor short-circuit current gain-bandwidth product in common emitter configuration
$C_{ob} \cong C_c + C_{cb} + C_{ce}$
$e^{-\frac{jm\omega}{\omega_\alpha}}$ — excess phase shift factor

EMITTER FOLLOWER STABILITY *(cont'd from back cover)*

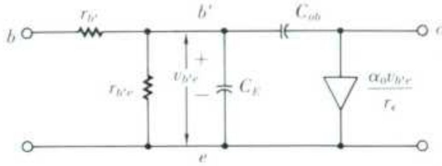

Fig. 3. *A more useful transistor equivalent circuit, valid for frequencies*
$$\omega \leq \frac{\omega_T}{3}$$

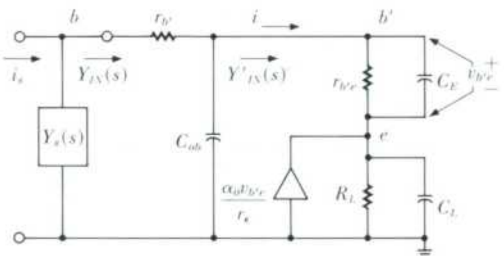

Fig. 4. *Equivalent circuit for circuit shown in Fig. 1.*

Substituting Z_E and Z_L into Equation 1 and rearranging gives

$$Y'_{IN}(s) = \frac{1}{(\beta_o + 1)\ (r_\epsilon + R_L)} \cdot \frac{r_{b'e}R_L C_E C_L s^2 + (r_{b'e}C_E + R_L C_L)\ s + 1}{\frac{r_\epsilon R_L}{r_\epsilon + R_L}\ (C_L + C_E)\ s + 1} \qquad (2)$$

The numerator and denominator polynomials can be separated into even and odd parts such that[2]

$$Y'_{IN}(s) = \frac{m_1(s) + n_1(s)}{m_2(s) + n_2(s)} \qquad (3)$$

The even part of $Y'_{IN}(s)$ is given by

$$Ev\ Y'_{IN}(s) = \frac{m_1 m_2 - n_1 n_2}{m_2^2 - n_2^2} \qquad (4)$$

with the real part resulting when $s = j\omega$. Thus

$$Re\ Y'_{IN}(j\omega) = Ev\ Y'_{IN}(s) \Big|_{s = j\omega} \qquad (5)$$

Using Equations 2 through 5, the real part of the driving-point admittance can be found to be

[2] M. E. Van Valkenburg,: "Introduction to Modern Network Synthesis", John Wiley & Sons, Inc., New York, 1960, Chapters 4 and 8.

$$Re\ Y'_{IN}(j\omega) = \frac{1}{(\beta_0 + 1)\ (r_\epsilon + R_L)} \cdot \frac{1 - \omega^2 \left[r_{b'e}R_L C_E C_L - \frac{r_\epsilon R_L}{r_\epsilon + R_L}\left(C_L + C_E\right)\left(r_{b'e}C_E + R_L C_L\right)\right]}{1 + \omega^2\left[\frac{r_\epsilon R_L}{r_\epsilon + R_L}\left(C_L + C_E\right)\right]^2} \qquad (6)$$

For the case where the imaginary part of the source impedance $Im\ Y_s(j\omega)$ is inductive, self-sustained oscillations will occur at a frequency determined by the source inductance and the transistor input capacitance for the condition

$$Re\ Y_s(j\omega) + Re\ Y'_{IN}(j\omega) \leq 0 \qquad (7)$$

Self-sustained oscillations will not exist so long as

$$Re\ Y_s(j\omega) + Re\ Y'_{IN}(j\omega) > 0 \qquad (8)$$

Therefore, oscillations can be prevented by controlling either $Re\ Y_S(j\omega)$ or $Re\ Y'_{IN}(j\omega)$, or both.

If $Re\ Y_S(j\omega) > 0$, as is usually the case, stability can be assured by making $Re\ Y'_{IN}(j\omega) > 0$. From Equation 6 it can be determined that $Re\ Y_{IN}(j\omega) > 0$ so long as

$$r_{b'e}R_L C_E C_L < \frac{r_\epsilon R_L}{r_\epsilon + R_L}\ (C_L + C_E)\ (r_{b'e}C_E + R_L C_L) \qquad (9)$$

or

$$r_\epsilon > \frac{\beta_0 r_\epsilon C_E}{C_L} - \frac{(\beta_0 + 1)\ r_\epsilon^2 C_E^2}{C_L^2\ R_L} \qquad (10)$$

where $r_\epsilon C_E$ = constant. For

$$\omega_T \cong \frac{1}{r_\epsilon C_E},$$

Equation 10 can be rewritten as

$$r_\epsilon > \frac{\beta_0}{\omega_T C_L} - \frac{\beta_0 + 1}{\omega_T^2 C_L^2 R_L} \qquad (11)$$

DESIGN FOR STABILITY

Equation 10 or 11 represents a very useful design criterion for assuring the stability of a capacitively-loaded emitter follower with an inductive source. The dynamic emitter resistance r_ϵ is inversely proportional to the dc emitter current I_E:

$$r_\epsilon = \frac{kT}{q\,I_E}$$

where k is the Boltzmann constant, T is the absolute temperature, and q is the electronic charge. At room temperature,

$$r_\epsilon = \frac{0.026}{I_E}.$$

This means that Equation 10 or 11, and therefore the condition $Re\ Y'_{IN}(j\omega) > 0$, can be satisfied simply by adjusting the dc emitter current I_E. However, if the emitter current needed to meet this condition is contradictory to other design conditions (e.g., ac signal requirements), then the more general case given by Equation 8 should be considered.

—Glenn B. DeBella

GLENN B. DeBELLA

Glenn DeBella joined -hp- in 1965 as a development engineer in the Frequency and Time Division. He has participated in the development of the 5260A Frequency Divider, and has given a series of lectures on transistors and linear circuits for engineers of the F and T Division.

Glenn holds BSEE and MSEE degrees from San Jose State College, and is a member of IEEE and Phi Kappa Phi. Before joining -hp-, he worked for three years as a circuit design engineer on magnetic tape recording equipment. He has also served for two years with the U.S. Army as a radar and radio technician.