ON ROBUST SUPPRESSION OF THIRD-ORDER INTERMODULATION TERMS IN SMALL-SIGNAL BIPOLAR AMPLIFIERS

Gleb V. Klimovitch

Watkins-Johnson Company, Stanford Research Park
3333 Hillview Ave., Palo Alto, CA, 94304-1223
Gleb.Klimovitch@WJ.com, Gleb@LAAS.Stanford.edu

ABSTRACT

A new method to cancel third-order intermodulation products in bipolar amplifiers is proposed and analyzed. Unlike many existing cancellation techniques, this intermodulation suppression is robust with respect to variations in temperature and component values.

1. INTRODUCTION

Third-order intermodulation (IM3) distortion is one of the main factors limiting spurious-free dynamic range [1] of small-signal bipolar amplifiers.

Different linearization methods by predistortion, feedback, and feedforward have been considered for small-signal and power amplifiers [1, 2]. In most feedforward techniques, third-order intermodulation terms cancel each other under certain numerical conditions on system parameters. These conditions are often quite sensitive to temperature and component variations, which necessitates component tuning or adaptive control.

In the proposed technique, the third-order intermodulation cancellation occurs when the product of the small-signal transconductance of the amplifying transistor g_m and the value of a circuit resistor equals a certain number. Using the fundamental relation between g_m and collector current, it is easy to automatically maintain the cancellation condition over a broad range of temperature changes and component variations by means of very simple adaptive control at DC.

The summary is organized as follows. In section 2, we concentrate on the analysis of intermodulation cancellation in memoryless versions of common-emitter and common-base amplifier. Extention of this technique to bipolar amplifiers with memory is also mentioned. Results and conclusions are given in section 3.

2. ANALYSIS

Fig. 1 shows a simplified circuit diagram of the common-emitter amplifier with resistive emitter degeneration R_E driven by voltage source $V_{IN} = v + V_0$, where V_0 and v are the DC and time-varying components of V_b respectively. To simplify the intermodulation analysis, the transistor is modeled an ideal current source with exponential current-voltage characteristics, and the effects of its reactive parasitics and base current on intermodulation are neglected. A more general model is considered in [3]. The collector current is written as $I_c = I_0 + i$, where I_0 is the DC collector current for $V_{IN} = v + V_0$. The voltage deviation

v in terms of current i is:

$$v = R_E i + V_T ln\left(1 + \frac{i}{I_0}\right) \ , \quad (1)$$

where $V_T = k_B T/q$ is the thermal voltage: $V_T \approx 26mV$ at room temperature. Let g_m be he transistor transconductance at $I_c = I_0$: $g_m = I_0/V_T$.

For technical convenience, we will often use V_T and I_0 as the units of voltage and current respectively: $V_T = 1$ and $I_0 = 1$. If the amplifier is considered as a system with series-series feedback and the small-signal loop gain is introduced [4]: $L = g_m R_E$, then Eq. (1) can be rewritten in a simplified dimensionless form as:

$$v = Li + ln\left(1 + i\right) = (1+L)i - \frac{i^2}{2} + \frac{i^3}{3} + ... \ , \quad (2)$$

where the terms higher than the third order in i are omitted. Series reversion [5] of Eq. (2) yields

$$\begin{aligned} i_3 &= \frac{v}{1+L} + \frac{v^2}{2(1+L)^3} + \frac{v^3}{(1+L)^4} \quad (3) \\ &+ \left(\frac{1}{2(1+L)} - \frac{1}{3}\right) \equiv \beta_1 v + \beta_2 v^2 + \beta_3 v^3 \ , \end{aligned}$$

where the terms higher than the third order in v are omitted. It is the cubic term $\beta_3 v^3$ that leads to IM3 distortion. The first and second constituents β_3 are due to quadratic and cubic nonlinearities in Eq. (2) respectively. The contributions of these nonlinearities to output IM3 perfectly cancel each other for $R_E = 1/2$ in dimensionless form, i.e. when the small-signal loop gain is:

$$L = 1/2 \ . \quad (4)$$

There are many possible biasing circuits which achieve and maintain condition (4) automatically regardless of changes in temperature and component values, since it is straightforward to build a DC voltage source with output proportional to the thermal voltage V_T [4]. If the low-frequency adaptive control is used to maintain the voltage across R_E equal $V_T/2$, then the bandwidth of the control loop $\omega_{control}$ should be small in comparison with the signal bandwidth, since IM3 is not canceled if the two tones are separated by less than $\sim \omega_{control}$. With good matching of transistors and resistors, the biasing circuit can be based on Fig. 2, where the FETs are matched, the biplar scaling factor is $M = \sqrt{e} \approx 1.65$, and the resistance of the RF choke is neglected. With this biasing circuit, IM3 is canceled for the arbitrarily small frequency difference between the two tones.

Cancellation condition (4) also holds for other implementations of linear feedback between collector current and base-emitter voltage in memoryless circuits. For example, a similar analysis applies to common-emitter amplifiers with other types of feedback [3] as well as to common-base amplifiers.

More flexibility in choosing the amplifier gain and input/output impedances, while cancelling IM3 products, is achieved by extending the presented treatment to circuits where impedances at baseband and carrier frequencies are different [3]. As an example, consider a simplified common-base amplifier shown in Fig. 3. We assume that the capacitor is well approximated by open and short circuit at baseband ω_{bb} and carrier ω_c frequencies respectively: $\omega_{bb} C R_S \gg 1$, $\omega_c C R_S \ll 1$, and the opposite is true for the inductor: $\omega_{bb} L \ll R_S$, $\omega_c L \gg R_S$. Then the small-signal loop gain is given by $L_{LF} = g_m R_E$ and $L_{HF} = g_m R_S$ at ω_{bb} and ω_c respectively. The general condition of IM3 cancellation is derived in [3]:

$$2L_{LF}\frac{1 + L_{HF}}{1 + L_{LF}} = 1 \ . \quad (5)$$

For memoryless systems with $L_{HF} = L_{LF} \equiv L$, Eq. (5) reduces to Eq. (4).

Effects of reactive parasitics on IM3 suppression are considered in [3]. In particular, they results in a finite IP3 at high frequencies even when the cancellation condition (4) or (5) is satisfied. However, the first-order effect of reactive parasitics on IP3 can be reduced by the proper cancellation techniques [3].

3. RESULTS AND DISCUSION

In Fig. 4, output-referred third-order intercept (IP3) for the common-emitter circuit of Fig. 1 is plotted as a function of emitter degeneration R_E with respect to IP3 at $R_E = 0$. R_E is swept under constant parameters - R_C and g_m. More than 10dB improvement in IP3 is observed when $R_E g_m = 0.5 \pm 0.05$. Since part of this improvement results from general linearization properties of negative feedback due to resistive emitter degeneration [6] rather than from IM3 cancellation, we also plotted increase in output-referred IP3 due to feedback only in a system with purely cubic nonlinearity for fair comparison. The latter equals $(1 + g_m R_E)^3$ in this case [3], as can be derived from Eq. (3).

In Fig. 5, input-referred IP3 for common-base circuit of Fig. 3 is plotted as a function of transistor transconductance g_m at constant $R_E = R_S = 50$Ohm for five values of emitter parasitic reactance $X_E = \omega_c L_E$ at the carrier frequency. Strong increase in IP3 is observed around $R_S g_m = 1/2$ for small L_E, as expected from condition (4). The impedance mismatch then results in return loss about 10dB. Increase in reactance reduces the peak IP3. At $R_S g_m = 1$, the calculated IP3 is approximately -6.7dBm in agreement with [1]. Both IM3 cancellation and impedance matching is achieved for $R_S g_m = 1$, $R_E g_m = 1/3$, according to Eq. (5).

The experimental measurements were performed on circuit based on Fig. 1 built with Siemens bipolar RF transistor BFP420 at frequency 50MHz. As the collector current was swept at two constant values of emitter degeneration resistor $R_E = 0.99$Ohm and $R_E = 2.00$Ohm, IM3 was suppressed by more than 15dB in the vicinity of $I_c = 6.7$mA and $I_c = 13.2$mA respectively. Both values of I_c are very close to the theoretical prediction $I_c = V_T/(2R_E)$, if two counteracting second-order effects are neglected: internal parasitic emitter resistance and transistor heating. As the fundamental frequency increases above 100MHz , the IM3 suppression starts degrading due to reactive parasitics in the circuit. Since the degradation is more pronounced for the smaller R_E and higher g_m values, it is probably caused by emitter inductance originating from both transistor and circuit layout parasitics.

To summarize, the proposed cancellation technique improves third-order intercept in small-signal bipolar amplifiers by more than 10 dB in theory and more than 7 dB in experiment. It is especially efficient at sufficiently low frequencies when reactive parasitics hardly affect cancellation precision. An important advantage of the new technique is its insensitivity to large temperature and component variations.

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4. REFERENCES

[1] B. Razavi, *RF Microelectronics*, Upper Saddle River, NJ : Prentice Hall PTR, 1998, pp. 49, 313-320, 177.

[2] Y. Hu, J. C. Mollier, and J. Obregon, A new method of third order intermodulation reduction in nonlinear microwave systems, *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 499-503, 1988.

[3] G. V. Klimovitch, Robust suppression of third-order intermodulation terms at high frequencies, to be submitted to *IEEE Trans. Mi-*

crowave Theory Tech., Special Issue on IMS-2000.

[4] P.R. Gray and R.G. Meyer, *Analysis and Design of Analog Integrated Circuits* New York: Wiley, 1993, pp. 338-346.

[5] M. Abramowitz and I. A. Stegun, *Handbook of mathematical functions*, New York: Dover, 1970.

[6] K.K. Clarke and D.T. Hess, *Communication circuits: analysis and design*, Reading, Mass: Addison-Wesley Pub. Co., 1971, pp. 123-129.

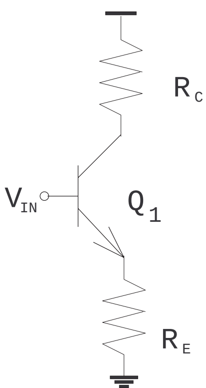

Figure 1: Simplified common-emitter amplifier with emitter degeneration.

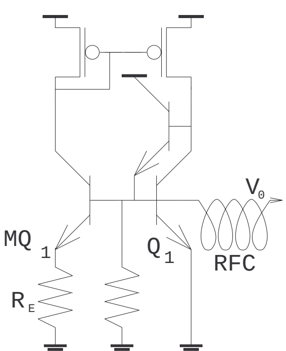

Figure 2: A possible biasing circuit for the common-emitter amplifier

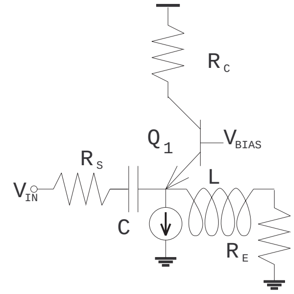

Figure 3: Simplified common-base amplifier.

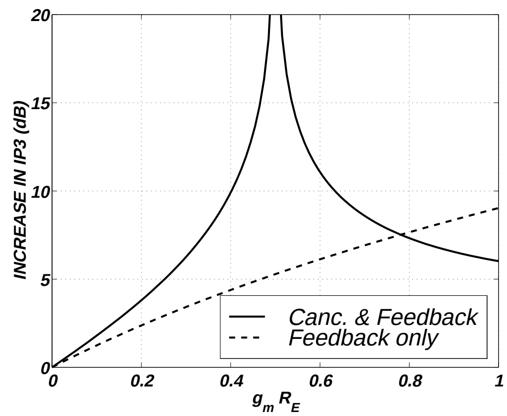

Figure 4: Increase in output-referred IP3 in common-emitter amplifier.

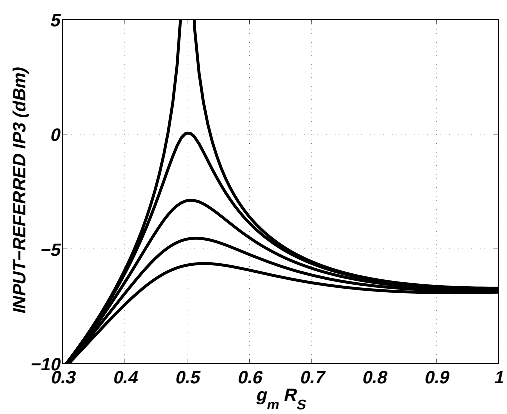

Figure 5: Input-referred IP3 in common-base amplifier for $R_E = R_S$ and emitter reactance $L_E \omega_c$ =0,0.05,0.1,0.15,0.2R_S.