RCA Application Note

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Cross-Modulation Effects in Single-Gate and Dual-Gate MOS Field-Effect Transistors

by

E.F. McKeon

This Note describes the cross-modulation performance of typical single-gate and dual-gate MOS transistors, and shows the inherent superiority of the dual-gate type for automatic-gain-controlled stages. It is shown that both single-gate and dual-gate MOS devices meet the general design requirements for receiver front-end stages.

Introduction

MOS field-effect transistors are now available that have gain and noise performance equal to or better than that of bipolar transistors in the vhf and uhf bands. This level of performance makes it possible for circuit designers to take advantage of the low cross-modulation distortion and large dynamic range of the MOS transistor, characteristics that are very desirable in front-end receiver stages.

MOS field-effect transistors have a transfer characteristic (drain current I_D as a function of gate voltage V_G) that closely resembles a quadratic curve. If an expression is written for MOS transistor I_D as a function of V_G, the third-order components are found to be negligible over a wide range in comparison with bipolar-transistor third-order components.

Third-order effects are most significant in their contribution to cross-modulation distortion. Cross-modulation may be defined as the transfer of information from an undesired carrier to a desired carrier. The first and second stages of a receiver are the most important in consideration of cross-modulation distortion because the amplitude of the undesired carrier is insignificant in later stages as a result of the selectivity of the initial stages.

Explanation of Cross-Modulation

Cross-modulation may be explained as follows: If a device has a nonlinear transfer (output-current/signal-voltage) characteristic, and has both a desired signal V_1 and an interfering signal V_2 applied at its input, the amplification α_1 of the desired signal V_1 can be expressed as a function of the undesired signal voltage V_2. The gain α_1 is independent of the input-voltage amplitude if the transfer characteristic is either linear or quadratic. If such is the case, α_1 is also independent of the amplitude of the interfering signal V_2. However, if the transfer characteristic deviates from either the linear or quadratic form, the amplification of the desired signal depends on the magnitude of V_2. Thus, if V_2 is amplitude-modulated, α_1 varies as the modulation on V_2 and produces cross-modulation when the AM information is transferred from V_2 to V_1 through the gain fluctuations of α_1. A more detailed treatment of this phenomenon is given in the Appendix.

Appendix Eq.(10) shows that the cross-modulation factor K is independent of the magnitude of the desired signal V_1 and is directly proportional to the square of the interfering signal. Because the interference or undesired information occurs as a modulation of the desired signal, when cross-modulation is present no amount of selectivity after the device causing the cross-

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modulation can remove the distortion. The undesired carrier frequency must either be suppressed by selective circuits before the active device, or the device itself must have a low value of S_3 (third-order effects of I_D as a function of V_G). If the transfer characteristic is quadratic, as is very nearly true in the case of the MOS transistor, then S_3 is minimized along with K. For a given value of S_3, K is minimized if S_1 (transconductance) is high.

Modulation distortion is defined as the production of harmonics of the modulation frequency of the desired carrier. Cross-modulation is related to modulation distortion by the following expression:

$$\frac{D_2}{V_1^2}:\frac{K}{V_2^2}=\frac{3}{8}m_1$$

where D_2 is the per-cent distortion of the desired carrier modulation (i.e., the modulation distortion) and m_1 is the per-cent modulation of the desired carrier. Because modulation distortion is related to cross-modulation (both are caused by third-order effects of the transfer characteristic), a device that minimizes one effect also minimizes the other. A cross-modulation distortion of one per cent (transferred modulation $m_k = 1\%$ of desired-carrier modulation m_1) is generally regarded as satisfactory; although this percentage is arbitrary, most published data are based on it.

Cross-Modulation of RCA 3N128 and 3N140 MOS Transistors

Cross-modulation in an rf device designed for use in front-end receiver circuits has meaning only when it is considered in conjunction with the agc action of the device. Fig.1 shows the transfer characteristic of a typical single-gate MOS transistor, the RCA-3N128.

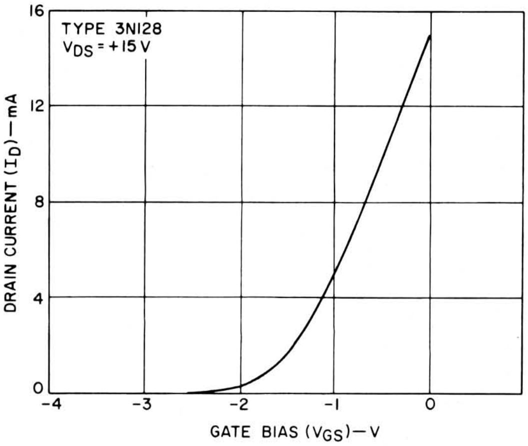

Fig.1 - Transfer characteristics of the RCA-3N128 MOS field-effect transistor.

Fig.2 shows the family of transfer characteristics for a dual-gate MOS type, RCA-3N140. The curves in Fig.2 represent different values of bias for gate No.2 of the 3N140 dual-gate unit. The transconductance is determined from the slope of the transfer curve at any given point. Forward agc is not recommended for the 3N128; reverse agc offers excellent gain control, but nonlinearities are encountered near "pinch-off". The recommended method of gain control for the 3N140 involves the use of reverse agc on gate No.2.

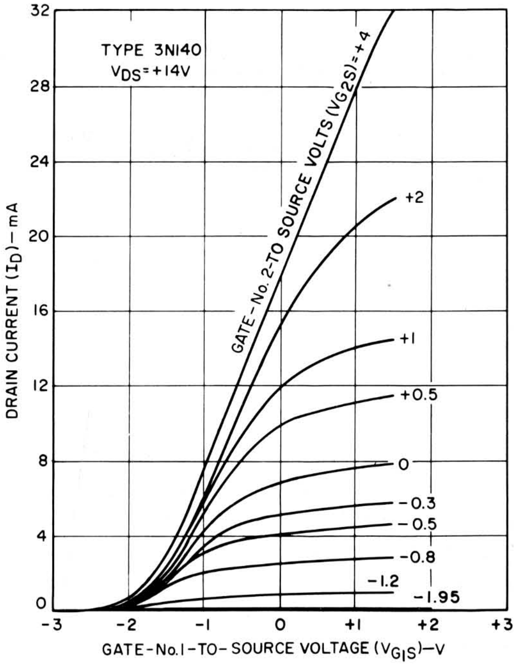

Fig.2 - Transfer characteristics of the RCA-3N140 dual-gate MOS field-effect transistor.

The first step in the complex problem of true evaluation of the available cross-modulation characteristics of a device is to measure the cross-modulation as a function of the gain reduction caused by agc. This measurement should be performed in a test fixture that uses an untuned input circuit. The untuned input circuit eliminates any attenuation of the interfering signal before it reaches the gate of the MOS and provides a measure of the inherent susceptibility of the device itself to cross-modulation distortion. Fig.3 shows a

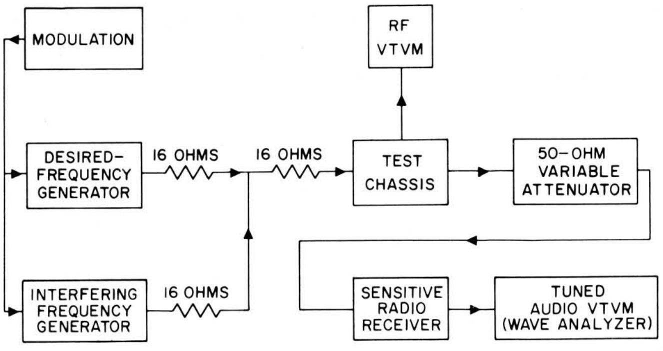

Fig.3 - Block diagram of the cross-modulation test circuit.

block diagram of the cross-modulation test circuit. The desired signal used for the test had a frequency of 200 MHz and an amplitude of 10 millivolts; the interfering-signal frequency was 150 MHz. The amplitude modulation (1 kHz) was 30 per cent for both the desired and the undesired signals.

3N128 Performance

Fig.4 shows the results of the cross-modulation measurements performed on the 3N128 for two methods of gain reduction: curve A represents gain reduction obtained by reverse bias applied to the gate; curve B was obtained by reduction of the drain-to-source voltage.

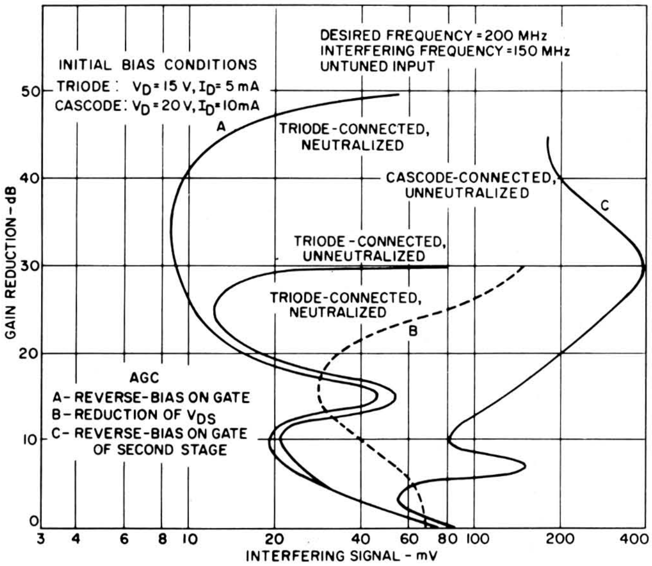

Fig.4 - Cross-modulation performance of the 3N128.

In both cases, the zero-dB attenuation point corresponds to the initial bias setting for optimum power gain and noise performance. At the zero-dB point, the cross-modulation performance of the 3N128 (i.e., the performance of the device itself without regard to matching compromises) is approximately ten times better than that of a bipolar transistor. In selection of an operating point for bipolar and MOS transistors, gain and noise usually determine the criteria because the curvature of the transfer characteristic (typically linear or quadratic) provides little cross-modulation in the region normally used for high transconductance.

At high attenuations, however, the cross-modulation performance of the 3N128 is not as good as can be obtained when forward agc is used with some bipolar transistors. This deterioration in cross-modulation performance at high attenuation results from the fact that the 3N128 is a sharp-cutoff device. As a result, large third-order nonlinearities occur near "pinch-off" in the low-transconductance region. Beyond "pinch-off," the transadmittance depends primarily on the capacitive feedthrough (C_{rss}), a parameter that does not have large third-order nonlinearities. Therefore, cross-modulation performance at the extreme limit of attenuation is very good.

An alternative method of agc for the 3N128 involves the reduction of the drain-to-source voltage. This method, however, limits the attenuation range possible because capacitive feedthrough increases by four to six times as the drain-to-source voltage approaches zero.

In an effort to improve cross-modulation performance at high attenuation, the cascode or "driven-grounded-gate" configuration was investigated. For the purposes of this Note, "cascode" is defined as the familiar series connection of a common-source and a common-gate stage. Two 3N128 transistors driven at 200 MHz in a cascode configuration provided satisfactory gain (17 dB) and noise figure (4.2 dB), and exhibited very good cross-modulation characteristics, as shown by curve C in Fig.4. Attenuation was accomplished by reverse-biasing only the gate of the common-gate stage. Neutralization was not required because rf stability was excellent without external feedback. Even without neutralization, a large attenuation range was obtained. Because the cascode arrangement allows the rf signal and the agc voltage to be kept separate, practically no detuning occurs during application of agc. These promising results led to the evolution of a dual-gate MOS field-effect transistor, the RCA-3N140.

3N140 Performance

The 3N140 features a series arrangement of two channels, each of which has an independent and isolated control gate. When properly connected (gate No.2 ac grounded), the 3N140 can provide rf-amplifier performance equal to or better than that of two MOS triodes connected in a cascode configuration. A 3N140 driven at a frequency of 200 MHz typically provides gain per stage of 20 dB and a noise figure of 2.4 dB; neutralization is not required.

One of the advantages of the dual-gate MOS transistor is its superior cross-modulation performance as a function of agc. Normally, the rf signal is applied to gate No.1, and gate No.2 is used for gain control. However, various combinations of agc voltage on gates Nos.1 and 2 are possible and were investigated. The cross-modulation performance of the 3N140 dual-gate unit is better than that of the 3N128 because the transfer characteristic of the 3N140 covers a larger portion of the quadratic region over the attenuation range. In addition, the series arrangement of channels produces a remote-cutoff characteristic.

Fig.5 shows the cross-modulation performance of the 3N140 when reverse agc is applied to gate No.2 only; both the attenuation range and the cross-modulation performance are good when this agc method is used, and amplifier detuning is minor (a shift of 3 MHz after

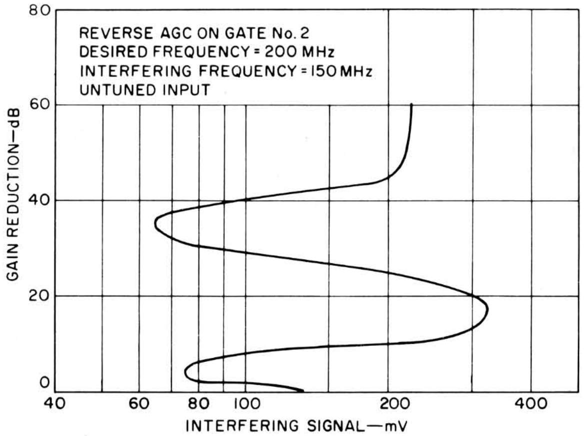

Fig.5 - Cross-modulation performance of the 3N140 when reverse agc is applied to gate No.2.

35-dB attenuation at an operating frequency of 200 MHz). Reverse agc on gate No.1 is not recommended because the cross-modulation performance degrades rapidly as cutoff is approached as a result of a deterioration of the quadratic curvature in this region. Forward agc on gate No.1 enhances the cross-modulation performance, but limits the attenuation range to less than 30 dB. A combination of reverse agc on gate No.2 and partial forward agc on gate No.1 provides improved cross-modulation, as shown in Fig.6, but limits the attenuation range to less than 35 dB because the feedback capacitance increases with the increasing forward bias of gate No.1. For the same reason (increasing feedback capacitance), reduction of gain by a decrease in the drain-to-source voltage is not recommended. Drain-to-source voltage reduction also leads to serious detuning problems.

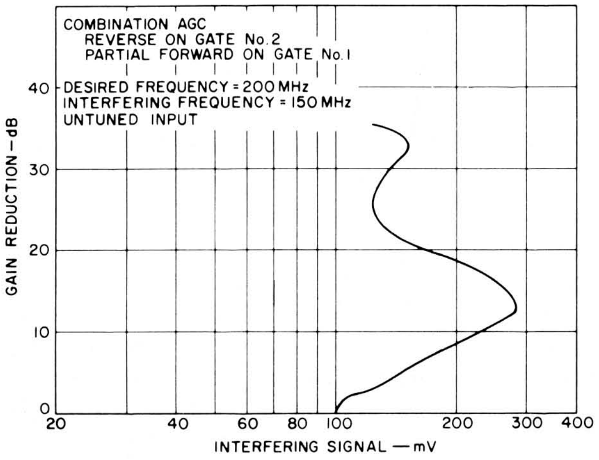

Fig.6 - Cross-modulation performance of the 3N140 with reverse agc on gate No.2 and partial forward agc on gate No.1.

Fig.7 shows curves of transconductance as a function of gate-No.1-to-source voltage V_{G1S} for several values of gate-No.2-to-source voltage V_{G2S}. These curves provide an explanation for the two dips in the cross-modulation curve of Fig.5. The circled areas A and B show the nonlinear transconductance region as the device gain is reduced by application of reverse bias to gate No.2. Third-order effects are introduced in these regions as the transfer characteristic varies with gate-No.2 bias.

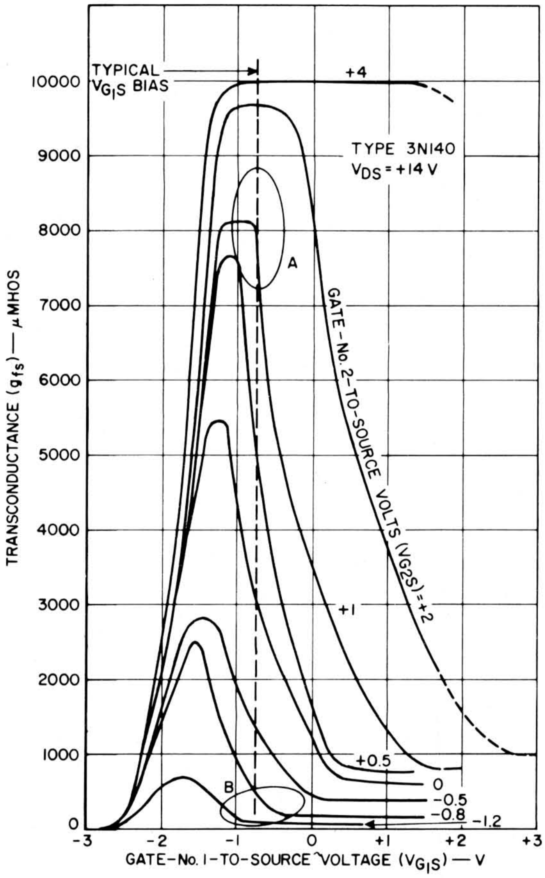

Fig.7 - Transconductance of the 3N140 as a function of gate-No.1-to-source voltage.

Front-End Design Using the 3N128 and 3N140

A front-end design must take into account the inherent cross-modulation properties of the active device used, as well as the input impedance of the device and the degree of mismatch required for best noise performance. The driving impedance that appears across the device terminals affects gain, noise, stability, and the level of interfering signal present at the input terminal after selectivity. Therefore, if a device has a very high input impedance and a complete conjugate match is achieved, the interfering signals that are not rejected by input selectivity are stepped up by the same high ratio as the desired signal. However, a high step-up ratio is not necessarily a disadvantage if the following conditions exist:

(a) the dynamic range of the device is good,

(b) the transfer characteristic is quadratic,

(c) the Q of the input circuit is high (proper selectivity),

(d) transconductance is also very high [$K = (1/2)(S_3/S_1)\,V_2^2$, where S_1 represents transconductance],

(e) the required driving impedance for best noise properties is close to that required for good gain.

Conclusions

Calculations based on a typical input match (for a given antenna impedance) for an rf stage using the 3N140 indicate that the cross-modulation performance should be at least two to four times better than that of most bipolar transistors. The 3N128 provides comparable performance except for agc action. The 3N140 maintains its superiority over the full attenuation range, while the performance of the 3N128 is about equal to that of bipolar transistors at high attenuations.

Appendix

An analytic approach to the phenomenon of cross-modulation involves the power-series expression of the transfer characteristic [$I_D = f(V_G)$], as follows:

$$i_d = I_{do} + \alpha v_g + \beta v_g^2 + \delta v_g^3 \ldots \tag{1}$$

where i_d is the instantaneous value of the total drain current, I_{do} is the dc drain current, and v_g is the rf signal or change in gate voltage from the V_{GS} bias point. The coefficients α, β, and δ refer to the bias point V_{GS}. By use of Maclaurin's theorem, the following alternative series can be written:

$$i_d = I_{do} + S_1 v_g + 1/2\, S_2 v_g^2 + 1/6\, S_3 v_g^3 \ldots \tag{2}$$

where S_1 is equal to α, or the slope of i_d/v_g (transfer) characteristic; S_2 is equal to 2β, or the slope of the S_1/v_g characteristic; and S_3 is equal to 6δ, or the slope of the S_2/v_g characteristic.

Cross-modulation occurs when there are two signals (desired and undesired) present at the gate of the MOS device. The desired signal can be expressed as $V_1 \cos \omega_1 t$; the undesired signal can be expressed as $V_2 \cos \omega_2 t$. The gate voltage v_g is then given by

$$v_g = V_1 \cos \omega_1 t + V_2 \cos \omega_2 t \tag{3}$$

If Eq.(3) is substituted in Eq.(1) and the resulting expression is limited to terms involving only ω_1 (because a tuned circuit at the MOS drain filters only ω_1 terms) and all fourth-order and higher terms are omitted as insignificant, the following expression is obtained:

$$i_{d1} = \alpha V_1 \cos \omega_1 t \,(1 + \frac{3}{4}\frac{\delta}{\alpha} V_1^2 + \frac{3}{2}\frac{\delta}{\alpha} V_2^2) \ldots \tag{4}$$

If V_1 is considered small compared to V_2, Eq.(4) may be simplified, as follows:

$$i_{d1} = \alpha V_1 \cos \omega_1 t \,(1 + \frac{3}{2}\frac{\delta}{\alpha} V_2^2) \ldots \tag{5}$$

Eq.(5) confirms the statement that the amplification of the desired signal depends on the amplitude of the interference unless the δ term in the progression (the term that represents the third-order effects) is negligible. If the interfering signal is modulated, V_2 in Eq.(5) must be replaced by $V_2(1 + m_2 \cos p t)$, as follows:

$$\begin{aligned} i_{d_1} &= \alpha V_1 \cos \omega_1 t \{1 + \frac{3}{2}\frac{\delta}{\alpha} [V_2^2 + 2 V_2^2 m_2 \cos p t \\ &\qquad + V_2^2 m_2^2 (\cos p t)^2]\} \\ &= \alpha V_1 \cos \omega_1 t \,[1 + \frac{3}{2}\frac{\delta}{\alpha} (V_2)^2 (1 + 1/2\, m_2^2) \\ &\qquad + 3 \frac{\delta}{\alpha} V_2^2 m_2 \cos p t] \end{aligned} \tag{6}$$

The amount of modulation imposed on the carrier of the desired signal by the interfering transmission is given by

$$m_k = \frac{3\,(\delta/\alpha)\, V_2^2}{1 + (3/2)\,(\delta/\alpha)\, V_2^2\,(1 + 1/2\, m_2^2)} \tag{7}$$

If V_2 is not too large, Eq.(7) can be reduced as follows:

$$m_k = 3\,(\delta/\alpha)\, V_2^2\, m_2 = K\, m_2 \tag{8}$$

Eq.(8) expresses the interference transferred to the desired signal as a fraction K of the undesired-signal modulation m_2. The cross-modulation factor K is dependent on the amplitude V_2 of the undesired signal. K may also be defined as the ratio of the transferred modulation m_k to the modulation m_1 of the desired carrier when both carriers are modulated to the same degree ($m_1 = m_2$), as follows:

$$\frac{m_k}{m_1} = \frac{m_k}{m_2} = K \text{ when } m_1 = m_2 \tag{9}$$

The expression for K is determined from Eq.(8), as follows:

$$K = 3 \frac{\delta}{\alpha} V_2^2 = 1/2 \frac{S_3}{S_1} V_2^2 \tag{10}$$

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