

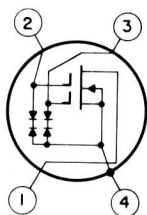
RF Applications of the Dual-Gate MOS FET up to 500 MHz

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The RCA dual-gate protected, metal-oxide silicon, field-effect transistor (MOS FET) is especially useful for high-frequency applications in RF amplifier circuits. The dual-gate feature permits the design of simple AGC circuitry requiring very low power. The integrated diodes protect the gates against damage due to static discharge that may develop during handling and usage. This Note describes the use of the RCA-3N200 dual-gate MOS FET in RF applications. The 3N200 has good power gain and a low noise factor at frequencies up to 500 MHz, offers especially good cross-modulation performance, and has a wide dynamic range; its low-feedback capacitance provides stable performance without neutralization.

Gate-Protection Diodes

Fig. 1 shows the terminal diagram for the 3N200. Gate No. 1 is the input signal electrode and Gate No. 2 is normally used to obtain gain control. The back-to-back diodes are connected from each of the gates to the source terminal, lead No. 4. If short duration pulses greater than ± 10 volts, generated for example by static discharge, are inadvertently applied to either gate, the protective diodes limit these voltages and shunt the current to the source terminal. Thus the gates, under normal operating conditions, are protected against the effects of overload voltages.¹



LEAD 1 - DRAIN
LEAD 2 - GATE NO. 2
LEAD 3 - GATE NO. 1
LEAD 4 - SOURCE, SUBSTRATE,
AND CASE

Operating Conditions

Typical two-port characteristics at 400 MHz including both "y" and "s" parameters, are given for the 3N200 in the RCA technical bulletin, File No. 437. This note makes use of the "y" parameters; however, designers who prefer the alternate method can, by parallel analysis, make use of the "s" parameters.

A recommended operating drain current (I_D) for the 3N200 is approximately 10 milliamperes with Gate No. 2 sufficiently forward biased such that a change in the bias voltage does not greatly affect the drain current. An adequate Gate No. 2-to-source voltage (V_{G2S}) is approximately +4 volts. The forward transadmittance (y_{fs}) increases with drain current, but saturates at higher current levels. The increase in RF performance at drain currents above 10 milliamperes is achieved only with less efficient use of input power.

To establish the optimum operating conditions for a type, consideration must be given to the range of variations in characteristics values encountered in production quantities of the type.² One important measure of type variation is the range of zero bias drain current (I_{DS}). The current range given in the 3N200 technical bulletin for I_{DS} is from 0.5 mA to 12 mA. A fixed bias condition intended to center the range of drain current at the desired level, still will produce an operating drain current range of 11.5 milliamperes with a resultant wide range of forward transconductance (g_{fs}). The drain current can be regulated by applying dc feedback with a bypassed source resistor (R_S). A good approximation of R_S (where $I_{DQ} \geq I_{DS}/2$) can be calculated by the use of the following formula*, assuming that V_{G1S} vs. I_{DS} is linear over the current range under consideration:

$$R_S \approx \left(\frac{1}{g_{fs}(\text{min.})} \right) \left(\frac{\Delta I_{DS}}{\Delta I_{DQ}} - 1 \right) \quad \text{Eq. 1}$$

*See Appendix

Fig. 1 - Terminal diagram for the 3N200.

where:

ΔI_{DS} is the current range given in the 3N200 technical bulletin

ΔI_{DQ} is the desired range of operating current

$g_{fs}(\text{min.})$ is the minimum forward transconductance at 1000 Hz

With the value of R_S established, then the Gate-No. 1 Voltage (V_{G1}) can be calculated from the equation

$$V_{G1} = V_{G1S} + I_{DQ} R_S \quad \text{Eq. 2}$$

where V_{G1S} is estimated by:

$$V_{G1S} \approx \frac{I_{DQ} - I_{DS}}{g_{fs}(\text{avg.})} \quad \text{Eq. 3}$$

where:

$g_{fs}(\text{avg.})$ is the average forward transconductance

To establish the Gate-No. 2 Voltage (V_{G2}), follow the same procedure described for calculating the Gate-No. 1 Voltage, except that a fixed V_{G2S} of approximately 4 volts is adequate.

If gain control is desired, apply a negative-going voltage to Gate No. 2. Because Gate No. 2 has little control in the

voltage range of +2 to +5 volts, this characteristic may be used to effect AGC delay of the device in order to maintain the low noise figure until the RF signal is out of the noise-level range.

Stability Considerations

Typical "y" parameter data as a function of frequency are given in Table 1. Maximum available gain (MAG) calculated from these data are also included to indicate ideal gain performance (i.e., $y_{rs} = 0$). The ability of the MOS FET to approach these gain levels depends on the device maintaining stable performance at the required operating frequency.

There are several methods which may be used to test for gain vs. stability. One of these methods, the Linvill Criteria (C), is defined by the equation:

$$C = \frac{y_{rs} y_{fs}}{2g_{is} g_{os} - R_e (y_{rs} y_{fs})} \quad \text{Eq. 4}$$

A value for C which is less than 1 indicates unconditional stability. Applying the 400-MHz values taken from Table 1 to the Linvill Criteria yields a value of $C = 0.615$; substantially less than the value indicating unconditional stability.

CHARACTERISTICS	SYMBOL	FREQUENCY (MHz)					UNITS
		100	200	300	400	500	
<u>y Parameters</u>							
Input Conductance	g_{is}	0.25	0.8	2.0	3.6	6.2	mmho
Input Susceptance	b_{is}	3.4	5.8	8.5	11.2	15.5	mmho
Magnitude of Forward Transadmittance	$ y_{fs} $	15.3	15.3	15.4	15.5	16.3	mmho
Angle of Forward Transadmittance	$\angle y_{fs}$	-15.0	-25.0	-35.0	-47.0	-60.0	degrees
Output Conductance	g_{os}	0.15	0.3	0.5	0.8	1.1	mmho
Output Susceptance	b_{os}	1.5	2.7	3.6	4.25	5.4	mmho
Magnitude of Reverse Transadmittance	$ y_{rs} $	0.012	0.025	0.06	0.14	0.26	mmho
Angle of Reverse Transadmittance	$\angle y_{rs}$	-60.0	-25.0	0	14.0	20.0	degrees
Maximum Available Gain	MAG	32.0	24.0	17.5	13	10.0	dB

Table 1 - "y" Parameters from 100 to 500 MHz

The following equation for Maximum Usable Gain (MUG)³ is:

$$\text{MUG} = \frac{2K |y_{fs}|}{|y_{rs}| (1 + \cos \theta)} \quad \text{Eq. 5}$$

where:

$$\theta = \angle y_{fs} + \angle y_{rs}$$

K = skew factor

$\angle y_{rs}$ = angle of reverse transadmittance

$\angle y_{fs}$ = angle of forward transadmittance

The skew factor, introduced in this equation, is a safety measure that establishes an arbitrary degree of skewing in the frequency response which may be introduced by regeneration. A value of 0.2 for K has been established on the basis of past experience. The value of MUG calculated at 400 MHz is 13.8 dB. This value of MUG is greater than the value of MAG, again indicating unconditional stability, since MAG, ignoring inherent feedback, is the conjugately matched gain. Therefore, neutralization or circuit loading is not required to insure stable performance, and the gain can approach MAG, limited only by circuit losses.

Reverse transadmittance (y_{rs}) is composed of several components, but the major ones are feedback capacitance (C_{rs}) and source-lead inductance (L_s). Therefore, care must be exercised in the application of the y_{rs} values, shown in Table 1, at the upper end of the usable frequency range. The 3N200 utilizes a JEDEC TO-72 package that has 4 leads. The data in Table 1 was compiled with the use of a socket which contacts the leads of the 3N200 as close as possible to the bottom of the package as specified by the JEDEC Standard Proposal SP-1028 "Measurement of VHF-UHF 'y' Parameters". The leads are shielded from each other to eliminate stray capacitance between the leads, but some lead inductance is inevitable. If the device is soldered directly to the circuit components using commercial production techniques rather than by precise laboratory methods, then additional source lead inductance can be expected. Also, some additional capacitive coupling may result if the input and output circuits are not completely isolated from each other.

Because the published y_{rs} value for the 3N200 is very small, the circuit y_{rs} values may differ significantly from the y_{rs} values shown in Table 1 and hence, may result in an unstable operating condition. It is impossible to provide data for all possible mounting combinations, therefore, a recommended mounting arrangement is shown in Fig. 2. The source and substrate in the TO-72 package of the 3N200 are internally connected to lead No. 4 and the case. The source-lead inductance can be reduced, if the case is used as the source connection. Fig. 2 illustrates a partial component layout in which the case is held by a clamp or other fingered

device. The clamp is soldered to a feedthrough capacitor to provide an effective, very-low inductance bypass to RF signals. This mounting arrangement still permits the use of a source resistor for DC stability, and enables the case to provide isolation between the input and output circuit in addition to the isolation afforded by the shield.

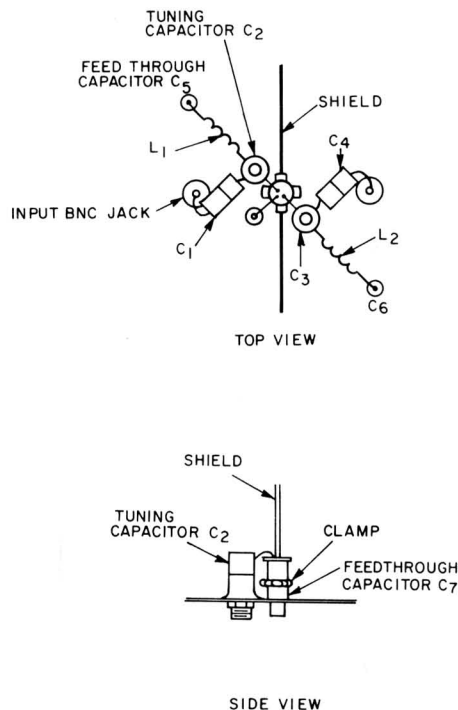


Fig. 2 – Partial component layout of 400-MHz amplifier circuit

The reduction of source-lead inductance provides in addition to greater stability, a lower input and output conductance. Table 2 shows the differences in "y" parameter values at 400 MHz when measured with the source connection made to lead No. 4 (in accordance with the published data for the 3N200) and when measured with the case connected directly to the ground plane of the test jig. The magnitude of reverse transadmittance is halved with a significant change in its phase angle. The input conductance is reduced by 30%, and the output conductance is reduced by 13%. A recalculation of the expressions for MAG, MUG, and Linvill Criteria (C) shows a significant improvement in gain and circuit stability.

While it is difficult to provide accurate information on the effects of shielding between the input and output circuits, its effect can be demonstrated when all other feedback components have been reduced to negligible values. The circuit, shown in Fig. 3 (for component layout see Fig. 2), was measured both with and without a shield. The maximum gain, without the shield, averaged 0.8 dB lower than with the use of the shield.

When receiver sensitivity is an important consideration in the design of an RF amplifier, a compromise must be made in the circuit power gain to achieve a lower noise

CHARACTERISTICS	SYMBOL	FREQUENCY (f) = 400 MHz		UNITS
		Normal Connection	Case Grounded	
Maximum Available Power Gain	MAG	13.0	15.7	dB
Maximum Usable Power Gain (unneutralized)	MUG	13.8	19.4	dB
Linville Stability Factor, C	C	0.615	0.335	mmho
<u>"y" Parameters</u>				
Input Conductance	g_{is}	3.6	2.5	mmho
Input Susceptance	b_{is}	11.2	11.7	mmho
Magnitude of Forward Transadmittance	$ Y_{fs} $	15.5	15.5	mmho
Angle of Forward Transadmittance	$\angle Y_{fs}$	-47.0	-40.0	degrees
Output Conductance	g_{os}	0.8	0.65	mmho
Output Susceptance	b_{os}	4.25	4.25	mmho
Magnitude of Reverse Transadmittance	$ Y_{rs} $	0.14	0.07	mmho
Angle of Reverse Transadmittance	$\angle Y_{rs}$	14.0	49.0	degrees

Table 2 – "y" Parameters at 400 MHz with source connection to lead No. 4 and with case connected to ground plane of test jig

factor. A contour plot of noise figure as a function of generator source admittance is shown in Fig. 4. Each contour is a plot of noise figure as a function of the generator source conductance and susceptance. Data for the noise figure were obtained from a test amplifier designed with very low feedback. Even though the area of very low-noise figure in the curves in Fig. 4 cover a broad range of source admittance, impedance-matching for maximum power gain could result in

a relatively poor noise figure. As shown in Table 2, the input conductance (g_{is}) with the case grounded is 2.5 mmho. With the reactive portion tuned out, the noise factor at power matched conditions is almost 1 dB higher than the optimum noise figure. However, matching to 5.0 mmho results in a near optimum noise factor with a loss of only 0.5 dB in gain. In addition, impedance matching to high conductance

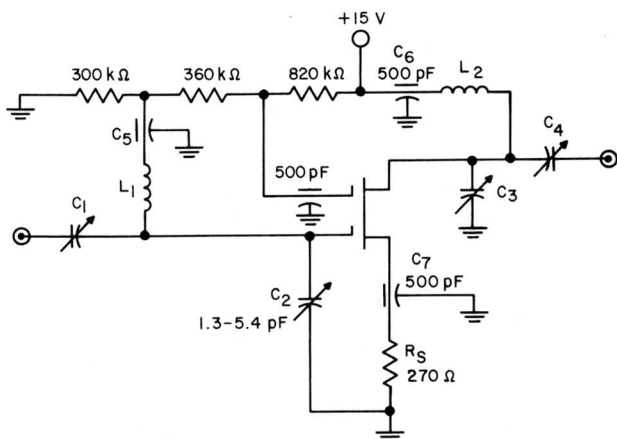


Fig. 3 – 400-MHz amplifier circuit

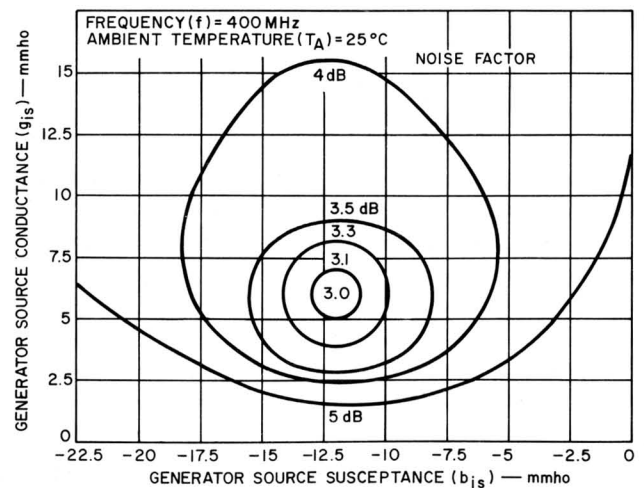


Fig. 4 – Noise factor vs. generator source (input) admittance (Y_{is})

also benefits crossmodulation performance, as will be discussed in a later section.

Gate Protection Diodes

The diodes incorporated into RCA dual-gate MOS FETs, for gate protection, have been designed to minimize RF loading on the input circuits. The small amount of RF loading results in only a fraction of a dB loss in power gain and a negligible increase in the noise figure. The advantages of diode protection, greatly outweigh the slight loss in power gain, especially in an RF amplifier intended for the input stage of a receiver.

In addition to the protection afforded in normal handling, the diodes also provide in-circuit protection against events such as: static discharge due to contact with the antenna, delay in transmit-receive switching, or connection of an antenna with an accumulated charge to the receiver.

Crossmodulation

Crossmodulation is an important consideration because it is an inherent device characteristic where circuit considerations are secondary. Crossmodulation is the transfer of modulation from an undesired signal on a desired signal caused by the non-linear characteristics of a device.

Crossmodulation is proportional to the third-order term of the expansion of the $I_D - V_{GS}$ curve. It is normally specified as the undesired signal voltage required to produce a crossmodulation factor of 0.01. The crossmodulation factor is defined as the percent modulation on a desired carrier by the modulated undesired signal divided by the percent modulation of the undesired signal.⁴

Inspection of the $I_D - V_{G1S}$ curve of Fig. 5 offers an insight to the possible crossmodulation as a function of gain-reduction performance. When both channels of the 3N200 are fully conducting current, as shown by the $V_{G2S} = 4$ -volt curve, the device approximately follows a square-law characteristic. If the $I_D - V_{G1S}$ curve was ideal, the third-order term would be zero; but in practical cases, the

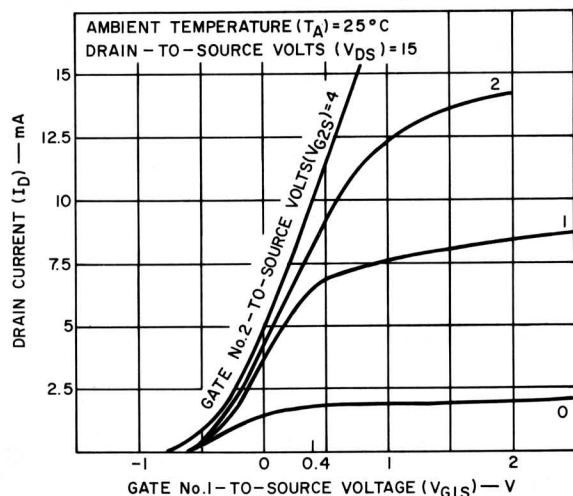


Fig. 5 — Drain current (I_D) vs. gate No. 1-to-source voltage (V_{G1S})

third-order term and crossmodulation have some low values. When the gain is reduced, by the application of bias to Gate No. 2, the square-law characteristic changes to a curve with a knee. Sharp curvatures usually result in larger high-order terms and poorer crossmodulation performance can be expected at lower gain conditions. If in Fig. 6, Circuit A, we assume a fixed bias (V_{G1S}) of approximately +0.4 volt, then the expected variation in crossmodulation is determined at the points where the ordinate at $V_{G1S} = +0.4$ volt crosses the curves. Crossmodulation performance at values of $V_{G2S} = +4$ volts to cutoff is as follows: good (low crossmodulation) at +4 volts, poorer at +2 volts, poorest at +1 volt, and again improves from zero volts to cutoff.

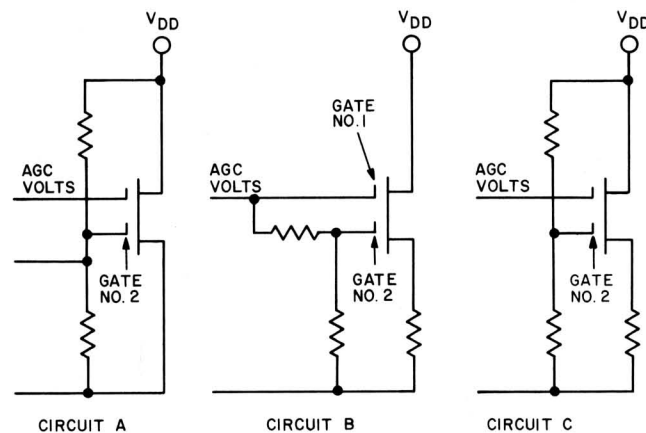


Fig. 6 — Biasing circuits using the 3N200

Curve A, Fig. 7 shows a curve of the undesired signal with a crossmodulation factor of 0.01 as a function of gain reduction. The curve indicates performance is poorest when gain reduction is in the 3- to 15-dB region; this region represents a Gate No. 2-voltage range of approximately 0.5 volt to 2 volts. The exception to the poor crossmodulation perform-

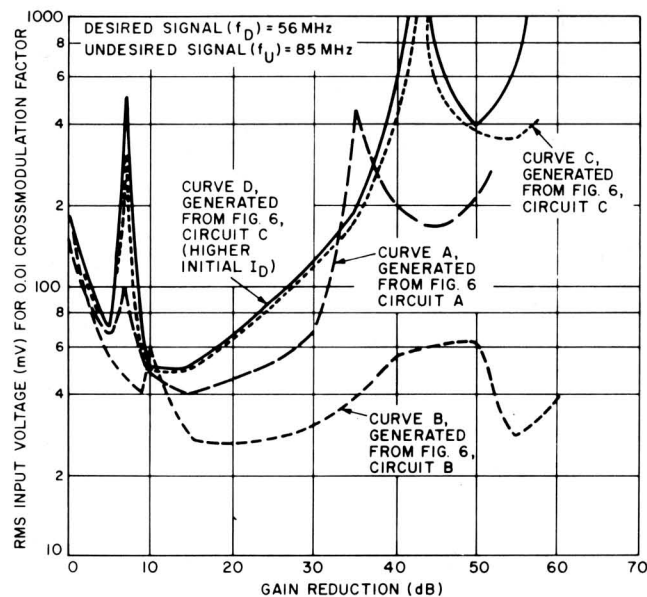


Fig. 7 — Crossmodulation vs. gain reduction using biasing circuits shown in Fig. 6

ance in this range is the sharp peak which occurs at the 5-dB level and is due to a curve inversion that takes place just prior to the knee. Beyond the 15-dB level, crossmodulation generally shows an improvement.

If Gate No. 1 is also reverse biased in conjunction with Gate No. 2 in the manner shown in Fig. 6, Circuit B, then the overall performance is poorer because the Gate No. 1 voltage will tend to follow the knee of each curve. This occurrence is evident in Fig. 7, Curve B. If Gate No. 1 is biased as shown in Fig. 6, Circuit C, the Gate No. 1-to-Source voltage intercepts the Gate No. 2 curves where the curvature is less severe, indicating as shown by Fig. 7, Curve C an improvement in crossmodulation performance. A further slight improvement is possible by the use of a higher initial operating drain current, which effectively moves the intercepts to the right on each curve. This improvement is indicated in Fig. 7, Curve D.

The curves in Fig. 7 establish that the biasing arrangement which provides optimum crossmodulation performance is the one in which Gate No. 1 forward bias increases as Gate No. 2 controls the gain. This biasing arrangement is easily accomplished by the use of a fixed Gate No. 1 voltage and a source resistor. As the Gate No. 2 bias voltage reduces the drain current, there is also a decrease in source voltage and an increase in the Gate No. 1-to-Source voltage. The gate-to-source voltage ratings must not be exceeded under any circumstances.

Summary

An RF amplifier, ideally, should provide high gain, a low-noise figure, and low crossmodulation. The 3N200 offers a good compromise in providing these three features. As indicated in the section on "Stability Considerations" a mismatch at the circuit input to a higher conductance level, provides an improved noise figure. The same mismatch condition also improves crossmodulation performance. The input signal at the gate of the device, when mismatched as indicated above, is lower than if it is power matched. The same ratio applies to any undesired signal and, thus, reduces the possibility of crossmodulation interference.

Appendix

The drain current of a device is established by the relationship

$$I_D = g_{fs} V_{G1S} + I_{DS}$$

where:

I_{DS} = drain current

at:

$$V_{G1S} = 0, \quad V_{G2S} = +4 \text{ volts.}$$

If a source resistor is used, as shown in Fig. A1, the gate No. 1-to-source voltage is

$$V_{G1S} = V_{G1} - I_D R_S$$

then

$$I_D = g_{fs} (V_{G1} - I_D R_S) + I_{DS} \quad \text{or}$$

$$I_D = \frac{g_{fs} V_{G1}}{1 + g_{fs} R_S} + \frac{I_{DS}}{1 + g_{fs} R_S}$$

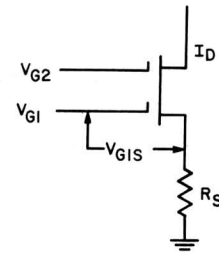


Fig. A1 – Bias circuit using the 3N200

The typical curves in Fig. A2 show drain current vs. Gate No. 1-to-Source Voltage as a function of I_{DS} level. These curves are almost linear when the typical operating drain current is in the 10-milliampere region. For the remainder of the analysis a linear relationship will be assumed for the required range of quiescent current. This assumption of linearity dictates that g_{fs} is a constant.

The required range of drain current is $I_{D2} - I_{D1}$

where:

$$I_{D2} = \frac{g_{fs} V_{G1}}{1 + g_{fs} R_S} + \frac{I_{DS} (\text{max.})}{1 + g_{fs} R_S}$$

$$I_{D1} = \frac{g_{fs} V_{G1}}{1 + g_{fs} R_S} + \frac{I_{DS} (\text{min.})}{1 + g_{fs} R_S}$$

$$\Delta I_D = I_{D2} - I_{D1} = \frac{I_{DS} (\text{max.}) - I_{DS} (\text{min.})}{1 + g_{fs} R_S} = \frac{\Delta I_{DS}}{1 + g_{fs} R_S}$$

Solving the above equation for R_S gives

$$R_S = \frac{(\Delta I_{DS} / \Delta I_D) - 1}{g_{fs}}$$

where:

g_{fs} is equal to the expected minimum value at the required I_D

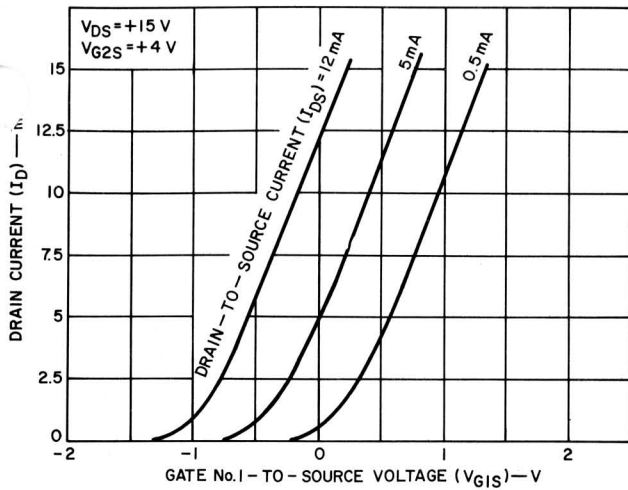


Fig. A2 – Drain current vs. gate No. 1-to-source voltage

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