A Practical Approach to Designing MOSFET Amplifiers for a Specific Gain

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Abstract

Metal Oxide Semiconductor Field Effect Transistors (MOSFETs) are primarily recognized for their use as a solid state switch. They are often overlooked as amplifiers because, despite several advantages over Bipolar Junction Transistors (BJTs), they are seen as devices that can only deliver single digit voltage gains. Even though *calculating* the gain for a MOSFET amplifier design is a well understood exercise, *designing* a MOSFET amplifier for a specified, moderately high gain *at the outset* is not. This is because the gain parameter of a MOSFET, its transconductance, is both a function of, and interacts with, its bias point. This paper reinforces the Problem Definition, Paper Design, Validation, and Verification engineering process with a step-by-step method to design a modern MOSFET amplifier to a specific, moderately high voltage gain. Instead of the "design the amplifier, analyze it to determine its gain, and modify" iterative process, this approach greatly reduces the empirical nature of the typical FET Amplifier design process taught in text books by introducing the engineering student to a comprehensible method to approximate the desired voltage gain on the first iteration.

Introduction

MOSFET transistors offer versatility in a number of applications that Bipolar Junction Transistors do not. Due to its very low (sub-ohm) R_{DS} resistances, the MOSFET transistor has low V_{DS} saturation voltages as compared to the V_{CE} saturation voltages of a BJT. Even low power MOSFETs have V_{DSon} voltages of 0.1 volts or less when the BJT V_{CEsat} voltages can be in the 0.5 to 1 volt range and even higher. This allows the MOSFET to have significantly improved switching characteristics over the BJT in comparable applications. In addition, most MOSFETs have an output protection diode which allows utility comparable to a BJT in switching inductive loads such as relays and motors.

Along the same lines, the MOSFET retains its amplification characteristics much closer to its turn-off and saturation points than does a BJT. This allows MOSFET transistor amplifiers to

have improved output voltage range over that of a BJT particularly in applications where the power supply may be 5V or even less. Another advantage of the MOSFET, particularly those constructed in VMOS is its improved temperature stability over the BJT greatly reducing the possibility of thermal runaway[1,2].

Perhaps the most attractive characteristic of the MOSFET transistor is its high input impedance. Regardless of whether it is used in a switching or amplification application, the MOSFETs high input impedance allows it to be driven by a low or a higher impedance source with little or no degradation of the signal.

Many designers ignore the MOSFET in linear applications because they believe that the gains that can be realized by MOSFET amplifiers are too low to bother with. [3,4] However, if they are carefully biased and utilized at frequencies below 100KHz, gains from a MOSFET amplifier can *practically* approach 50 while offering much higher input impedances, as mentioned above. This is possible because the gain parameter of a MOSFET, its transconductance (y_{fs}), is a function of its bias point (Q point). In contrast, the current gain function of a BJT (h_{FE}) is approximately constant over most its range of bias points, relative to a MOSFET.

Practical MOSFET Amplifier Design

Problem Definition and Design Constraints

Given a specific input voltage (Vi = 50mVp which is the approximate output amplitude of a piezoelectric microphone) and frequency range (audio), an amplifier is primarily intended to provide a certain minimum gain. To demonstrate our hypotheses, we will select an inverting gain of 50. Also, the power supplies available in the system, V_{DD} (+12V) have usually been decided upon by this point in the design process. Within these parameters, a MOSFET based amplifier design can be accomplished. Finally, to demonstrate the design process, a readily available, general purpose MOSFET was selected – the 2N7000. For the purposes of this study, the following characteristics of the 2N7000 manufactured by Vishay[5] were utilized:

$V_{GSoff} = 2.0V$

$V_{GSon} = 10V$

$I_{Don} = 1A$

These characteristics are fairly typical, though not identical, to those found for 2N7000s from other suppliers.

The transfer function of a MOSFET relates its output current (I_D) to its input voltage. Over the range of its bias points, this parabolic function is defined by the equation:

$I_{Dq} = K * (V_{GSq} - V_{GSoff})^2$

Once the MOSFET is turned on, the constant K is the ratio of drain current per the square of the voltage from its gate to source (V_{GS}) at a point midway along its transfer curve. K is specific to each type MOSFET but once calculated may be utilized for most practical bias conditions.

$$K = I_{Don} / (V_{GSon} - V_{GSoff})^2$$

Utilizing our 2N7000 characteristics outlined above:

$$K = 1A / (10V - 2V)^2 = 15.6 \text{ mA/V}^2$$

So the transfer function equation for this 2N7000 MOSFET, Equation 1):

Transfer Function – Equation 1

1) $I_{Dq} = .0156 \text{ A/V}^2 * (V_{GSq} - 2)^2$

Paper Design

Calculating the Q Point

The gain (inverting) of a MOSFET Common Source amplifier is a function of its output impedance (Zo) and its transconductance at the bias point (y_{fq}) and is defined by the equation:

$$Av = - y_{fq} \text{ Zo}$$

By analyzing the AC equivalent model of a properly decoupled and bypassed Common Source MOSFET amplifier (see Fig 1 below),
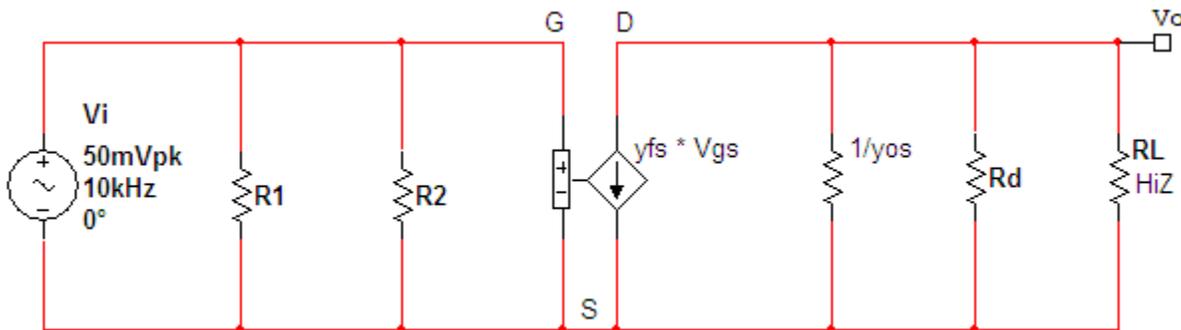

Figure 1. Common Source MOSFET – AC Equivalent Model

it is apparent that Zo equals the load resistance(R_L) in parallel with the drain resistor (R_D) in parallel with the output impedance of the transistor (which is 1/ the output conductance (yos)).

$$Zo = R_L \backslash\backslash R_D \backslash\backslash 1/yos$$

To limit the scope of the design, we will assume the amplifier is feeding a known, high impedance load. Therefore the amplifier's output impedance may be approximated by R_D // 1/yos. The output conductance of many FETs is frequently an underspecified or, more often, unspecified parameter[2,5] because it is frequently assumed to be sufficiently high as to not affect the overall output impedance of most FET applications [6]. For the same reasons, many simulation models do not reflect accurate values of yos as a function of bias conditions. In *practice*, the output conductance of an FET does affect the performance of amplifiers biased at lower drain currents (I_D). But without better data to determine what the output conductance is at lower bias levels, we have no choice but to move forward and to compensate for yos whenever possible. Therefore we will assume for the time being that:

$$Z_o \approx R_D$$

so

    a) $A_v \approx - y_{fq} R_D$

The practical output conductance will be addressed later in the paper.

In order to avoid the non-linear affects of a transistor's saturation and turn off areas of operation, and assuming our output is symmetrical (a sign wave), a good place to bias the drain of the amplifier (V_D) is at approximately $0.5 * V_{DD}$[7]. Absent other design criteria, this assumption will normally allow for close to optimal output amplitude. Therefore:

$$V_{RD} = V_{DD} - 0.5 * V_{DD} = 0.5 * V_{DD}$$

so

    b) $R_D = 0.5 * V_{DD} / I_{Dq}$

The transconductance of a MOSFET can be found by taking the derivative of the transfer function for I_D at the V_{GS} bias point of the amplifier:

$$y_{fq} = \delta I_D / \delta V_{GS} = \delta [K * (V_{GSq} - V_{GSoff})^2] / \delta V_{GS} = 2*K*(V_{GSq} - V_{GSoff})$$

so

$$y_{fq} = 2 * .0156 \text{ A} / V^2 * (V_{GSq} - 2)$$

    c) $y_{fq} = .0312 \text{ A} / V^2 * (V_{GSq} - 2)$

By substituting equations b, and c, into a, the inverting gain of our MOSFET amplifier can be approximated by:

$Av \approx - [.0312 \text{ A} / V^2 * (V_{GSq} - 2)] * [0.5 * 12V/ I_{Dq}]$

Finally, through one more step of algebraic manipulation, I_D as a function of gain (-50) can be realized:

I_D as a Function of Gain – Equation 2

2) $I_{Dq} \approx [.0312A / V^2 * (V_{GSq} - 2)] * 6V / 50$

Now, by substituting in our desired Av, we can simultaneously solve the transfer function and the gain function (Equations 1 and 2) to find the approximate bias condition. The utilization of a graphing calculator such as a TI 86 or 89 will solve equations 1 and 2 both simultaneously and graphically. Displaying the equations graphically results in a transfer curve that is very useful in determining other bias conditions as well. The equations as entered into the TI-89 are shown in Figure 2 below:
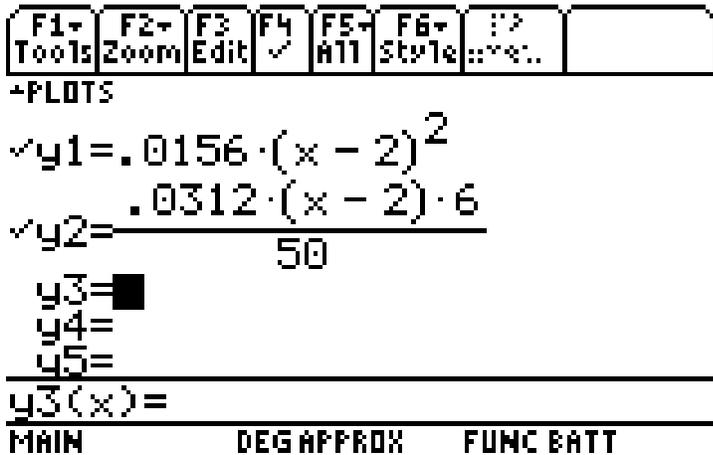

Figure 2. Transfer Curve and Gain Equations – TI-89

Graphing and solving simultaneously on the TI – 89 by utilizing its "Math" and "Intersection" functions results in an operating bias (Q) point of:

$V_{GSq} = 2.24V$ **$I_{Dq} = 897uA$**

as seen in Figure 3:

```
F1▾    F2▾  F3    F4      F5▾   F6▾  F7▾ ░░
Tools Zoom Trace ReGraph Math Draw Pen ⋮

Intersection
xc:2.23974e0    yc:896.64e⁻6
MAIN          DEGAPPROX    FUNC BATT
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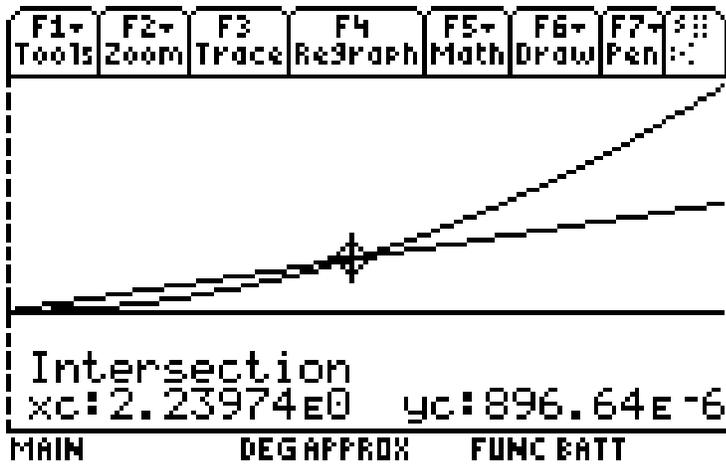
Figure 3. Graphing and Solving Simultaneously - TI – 89

Note that, if a gain outside of the capability of the MOSFET is selected, the graphs will not intersect and there will be no simultaneous solution.

Again utilizing the TI-89's "Math" functions, the transconductance at the bias (Q) point can be readily determined as well as shown in Figure 4:

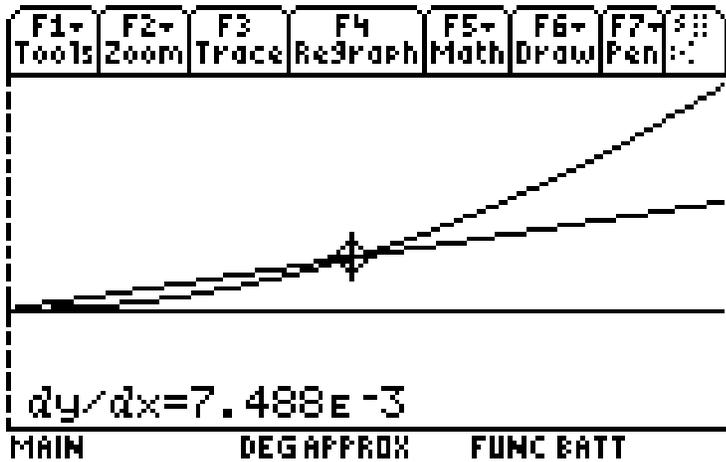$y_{fq} = \delta\, I_D\, /\, \delta\, V_{GS}\ \ = \textbf{7.49mS}$

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F1▾    F2▾  F3    F4      F5▾   F6▾  F7▾ ░░
Tools Zoom Trace ReGraph Math Draw Pen ⋮

dy/dx=7.488e⁻3
MAIN          DEGAPPROX    FUNC BATT
```

Figure 4. Transconductance at the Q Point

Calculating Circuit Component Values

Calculating R_D:

$R_D = 6V\, /\, 897uA = 6.69K\Omega$

6.69 KΩ is not a standard value. In order to compensate somewhat for an unspecified output conductance (yos), we will *round up R_D* to the next higher standard value resistor.

$R_D = 6.8K\Omega$

To provide additional thermal stability in a similar fashion to a BJT amplifier, we will utilize an R_S resistor.

Assuming that the bias point of the gate voltage (V_G) is not dependent on the characteristics of the input voltage (Vi), an acceptable place to bias V_S is at $0.1 * V_{DD}$. This rule of thumb also minimizes the voltage requirements and therefore the physical size and cost of C_S, the bypass capacitor. Therefore:

$R_S = 0.1 * 12V / 897uA = 1.34K\Omega$. Standard value $= 1.5K\Omega$

So:

$V_S = 897uA * 1.5K\Omega = 1.35V$

Note that:

$V_G = V_S + V_{GS} = 1.35V + 2.24V = 3.59V$

Recalling that the input impedance of a MOSFET transistor is close to infinity, the R_1 and R_2 resistors may be selected as if a simple voltage divider. In order to maintain the feature of high input impedance for our amplifier, we will select $R_2 = 2M\Omega$. Therefore:

$3.59V = 12V * 2M\Omega / (2M\Omega + R_1)$

Solving, $R_1 = 4.68M\Omega$ or $4.7M\Omega$ standard value.

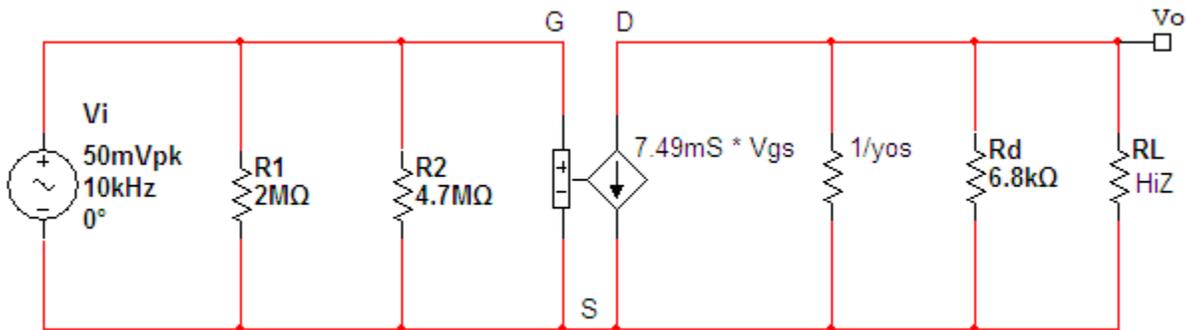The ac equivalent model specific to our design may now be shown in Figure 5 below:

Figure 5. AC Equivalent Model - Specified

In selecting the capacitors for a traditional Common Source Amplifier design, we select their impedances to be a relative ac short compared to the Thevenin equivalent resistance seen by each capacitor. We will assume a relative ac short exists when the reactance of the capacitor, Xc ≤

$0.1 * R_{TH}$. Note that the worst case reactance of a capacitor utilized in the audio frequency range is:

$$Xc = 1/ (2\pi*20Hz*C)$$

Therefore through some algebraic manipulation, we will select our decoupling capacitors Cg, Cd, and the source resistor bypass capacitor Cs according to the following equation:

$$C \geq 1/ (2\pi*20Hz*0.1 * R_{TH})$$

Cg, Cd and Cs are calculated below:

R_{TH} at the gate of the amplifier equals 2MΩ // 4.7MΩ or 1.4 MΩ therefore;

$$Cg \geq 1/(2\pi*20Hz*0.1 * 1.4M\Omega)$$

$$Cg \geq 57nF$$

R_{TH} at the source is approximately equal to Rs or 1.5KΩ therefore;

$$Cs \geq 1/(2\pi*20Hz*0.1 * 1.5K\Omega)$$

$$Cs \geq 53uF$$

R_{TH} at the drain is approximately equal to Rd + RL, a very high impedance, therefore we will select an inexpensive generic Cd;

$$Cd = 100nF$$

Finally the completed paper design of a Common Source Amplifier with very high input impedance and a designed-in gain of -50 is shown in Figure 6 below. It utilizes a 2N7000 Enhancement Mode MOSFET with a 12V power supply. Because of slight differences between the V_{GSoff} of the transistor model in Multisim and that of the Vishay specification, R1 was modified to the next highest standard resistor, 5MΩ. This modification resulted in an I_{Dq} closest to the specified I_D of 897uA.
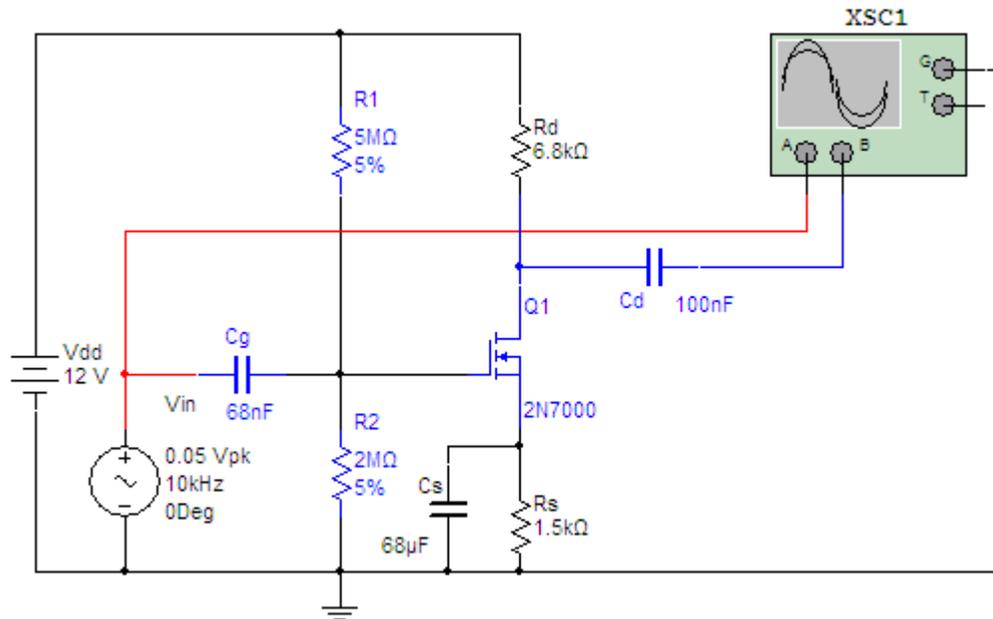

Figure 6. CS MOSFET Amplifier with a Gain of -50 – Paper Design

Design Validation

When simulated using National Instrument's MultiSim as shown in Figure 7 below, the *simulated* gain may be calculated:

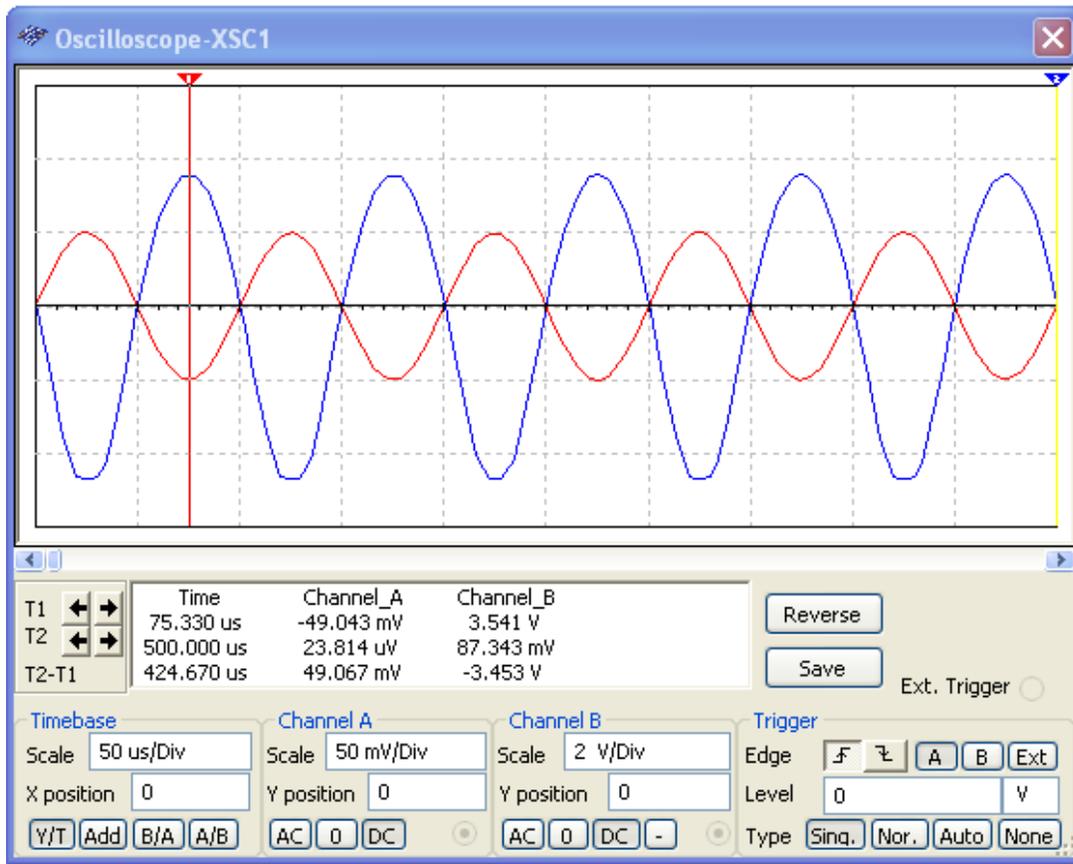**Figure 7. CS MOSFET Amplifier – Multisim Simulation**

$Av_{sim} = -Vo / Vi = -3.54Vp / 49mVp = -72$

A good bit higher than our design goal of -50!

Design Verification

When the circuit was prototyped in the laboratory it resulted in an actual output seen in Figure 8 below.
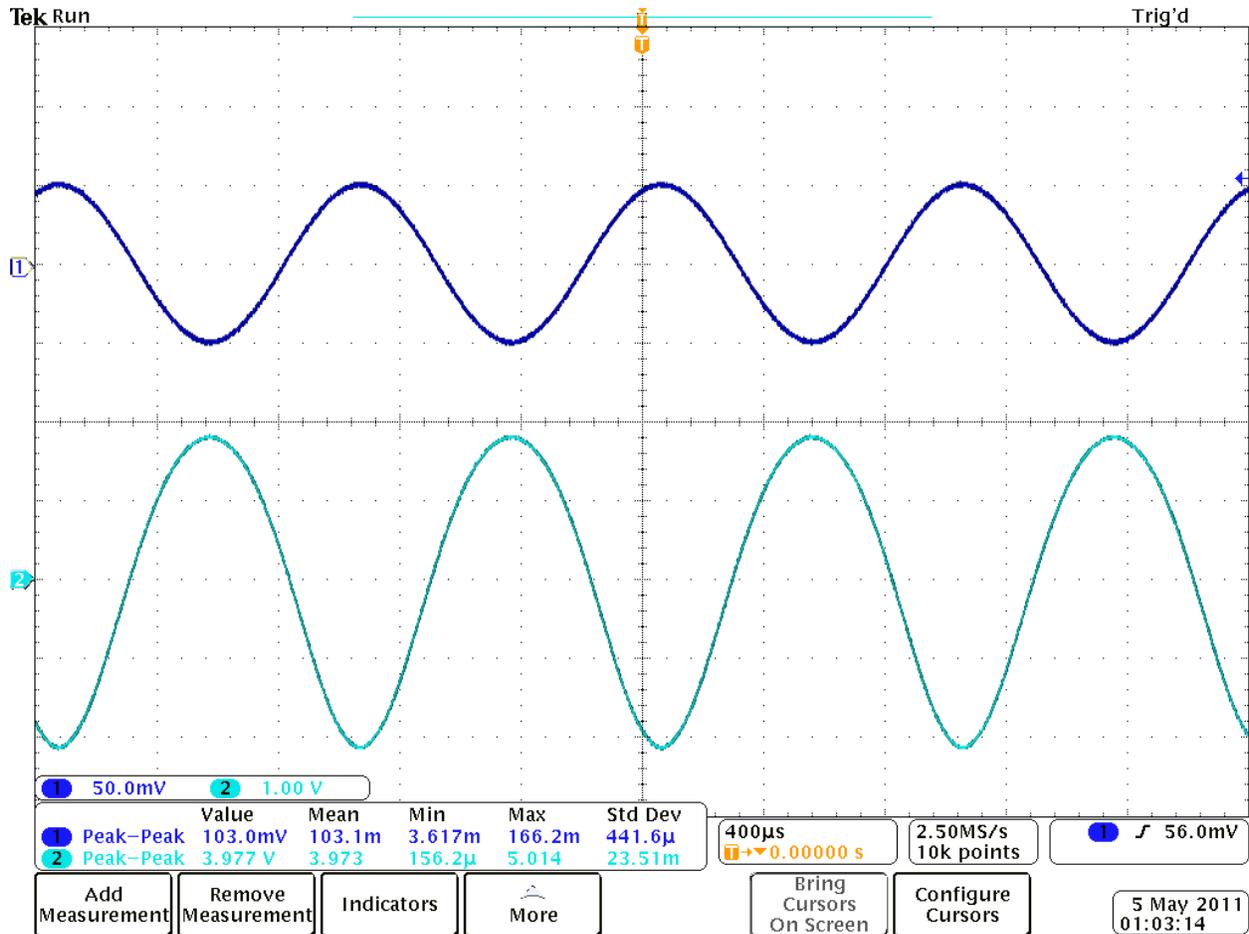

Figure 8. CS MOSFET Amplifier – Actual Output

Now the *actual* gain may be calculated:

Av_{act} = -Vo / Vi = -3.98Vp-p / 103mVp-p = -39

A bit lower than -50.

Analysis

It is apparent that the Multisim simulation and the actual prototype demonstrate significantly different results from our paper design.

	Paper Design	Multisim Simulation	Prototype (Actual)
Voltage Gain (Av)	-50	-72	-39

This is not at all unusual in the area of semiconductors and their design applications. The educational version of Multisim has one 2N7000 model that is selected to reasonably emulate devices fabricated by multiple manufacturers. In addition, few if any of the manufacturers specify the output conductance (y_{os}) as a function of bias conditions.

The prototype exhibits a gain closest to the paper design which, considering an amplifier had to be created utilizing a transistor with an unspecified yos, reinforces our design methodology. In fact, with actual data, we can utilize our prototype results to closely approximate the actual output conductance of the transistor. Revisiting the original gain equation,

$Av = - y_{fq} Zo$

Av and the transconductance can be substituted back in to determine the actual output conductance.

$Zo = - 39 / - 7.49mS = 5207\Omega$

Recall that

$Zo \approx R_D \backslash\backslash \ 1/yos$

Therefore, the actual output conductance of our 2N7000 can be closely approximated:

$yos = 1/ 5207\Omega - 1/ 6800\Omega = 45.0 \ uS$

Conclusions

Now that our actual output conductance and overall output impedance have been closely approximated, this information could be applied to slightly modifying the bias conditions to result in an actual gain closer to the design goal, if necessary. This is another application where utilizing the TI 89 to graph the transistor's transfer function comes in handy. Having the transfer function graphed on the calculator allows the "Trace" function to be utilized to move the bias conditions slightly up or down the transfer curve until the desired transconductance and corresponding voltage gain are reached.

This paper demonstrates that utilizing the appropriate transfer equation for the MOSFET in conjunction with developing a corresponding equation for I_D as a function of gain and solving them simultaneously can take much of the uncertainty out of designing an amplifier with a specific gain. In addition, utilizing the advantages of modern MOSFETs in amplifier design doesn't necessarily mean that the designer must compromise with low gains. Finally, because the output conductance of some transistors is incompletely specified, it may be necessary to make a second minor design iteration once the actual Q point and output conductance are closely approximated.

References

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