this we need a set of numbers that are not found in the Electrical Characteristics section of the datasheet. We need to know the maximum permitted IC junction temperature (150°C) and the maximum anticipated operating ambient temperature, which is 75°C for most consumer equipment. This gives a maximum junction-to-ambient temperature differential of 75°C. The other thing we need to know is the circuit thermal resistance. The circuit thermal resistance is the sum of three numbers: the junction-to-case thermal resistance of 1°C/Watt, the heatsink interface thermal resistance (silicon grease and an insulating mica or silicon washer) that is also around 1°C/Watt, and finally the heatsink thermal resistance. The heatsink is physically too large to be practical below 1°C/Watt, so this is the lowest number we will use for the heatsink thermal resistance. This gives us a total circuit thermal resistance of 3°C/Watt. What this latter number means is that each watt of power dissipation raises the junction temperature by 3°C. Dividing the thermal resistance into the maximum permitted temperature differential gives us the maximum power dissipation for the package . . . 25W.

If the package limits us to 25W power dissipation, then the maximum rated continuous output power is 40W, regardless of load or power supply voltage. What if we want more than 40W? For bipolar IC amplifiers this is possible by reducing the ambient temperature requirement or reducing the circuit thermal resistance, or both.

However, increasing the internal power dissipation is not the best way to go. Because the LM3875 has a very sophisticated set of protection features built in, there are few external components required. This means we can easily implement the discrete transistor solution to reduce the package power dissipation . . . use more than one package. As Figure 16 shows, two LM3875s connected in a bridge configuration is far less complex than an equivalent discrete single amplifier. Now the load is being driven from each end and, for an 8Ω speaker, each LM3875 "sees" a 4Ω load. The supply voltages are reduced to ±25V to prevent thermal shutdown at 75°C ambients, but each package can still dissipate 25W. Each IC amplifier of the bridge circuit will require a 1°C/Watt heatsink but the total maximum output power is doubled to 80W.

With the introduction of the Overture™ series, IC amplifiers have overcome the voltage and current constraints of high power audio amplifiers. The high level of integrated protection means that multiple package amplifiers are able to distribute the resulting power dissipation requirements without introducing a higher degree of circuit complexity.

Op Amp Test Circuits - Part II: Common-Mode Rejection

John Christensen

In the previous Linear Edge we went through the four tests that are usually at the top of the data sheet: offset voltage (V_{OS}), open loop gain (A_{VOL}), bias current (I_b) and offset current (I_{OS}). In this issue, we will continue down the data sheet through another of the DC parameters, Common-Mode Rejection.

A common-mode signal is a signal that is common to or applied to both inputs. CMR, Common-Mode Rejection or CMRR, Common-Mode Rejection Ratio is commonly defined as the ratio of the open-loop differential gain of the amplifier to the common-mode gain of the amplifier. The problem with this definition is that the open-loop gain of a high gain op amp is not something that can be easily measured in the real world. A number that is much easier to measure and is independent of open-loop gain is the input common-mode voltage change divided by the change in input offset voltage.

If the op amp input stage were perfectly matched, the applied common-mode signal would not cause a change in V_{OS}. Because of slight differences, a small V_{OS} shift is usually produced. This error occurs when the op amp is used as a non-inverting amplifier (Figure 17). Since the output causes both inputs to be equal, the input signal is a common-mode signal. This shows up as a distortion or varying error between the input and the output.

How should we go about measuring this change in V_{OS}? Why not use the same circuit we used before to measure V_{OS}? All we need to do is inject a common-mode signal, measure

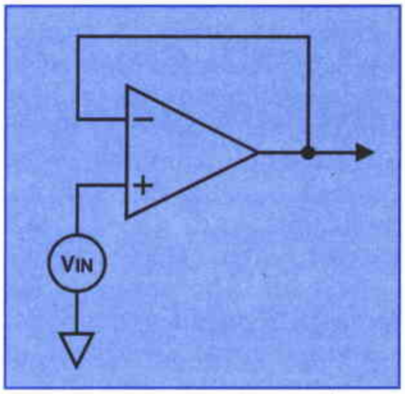

Figure 17: Buffer

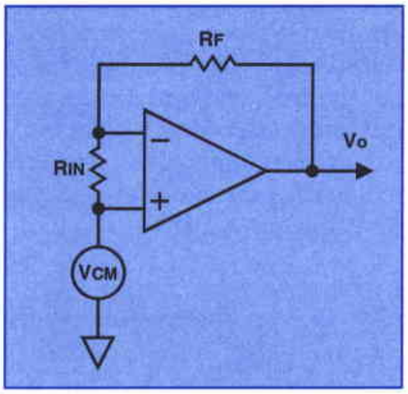

Figure 18: Poor Common-mode Test Circuit

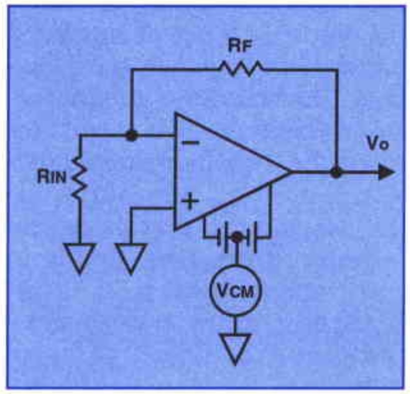

Figure 19: A Common Method Of Measuring CMR

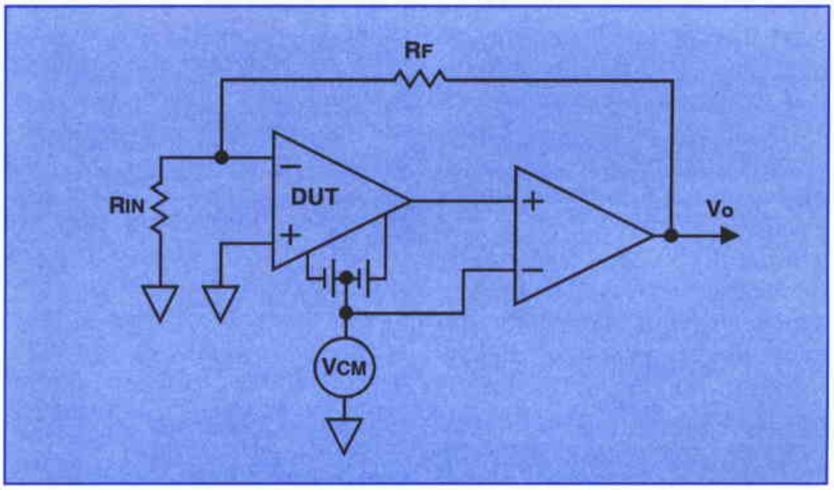

Figure 20: CMR Measurement with Buffer Added

the change in V_{os} and we have our answer (Figure 18).

Haven't we just moved the inputs up and down and left the rest at ground? Almost but not quite. We need to keep both inputs at the same common-mode voltage. To do this, the output stage must follow the input voltage. If the output doesn't move, the voltage at the inverting input will be a little lower than the non-inverting input for a positive V_{cm}. To make the output move requires a differential voltage at the input that is equal to the common-mode voltage divided by the open-loop gain. The result is that the output moves by:

Equation 2-0:

$V_{cm} - ((V_{cm} * (R_f/R_{in}))/A_{vol})$.

What we see at the output is the common-mode voltage plus the error caused by the finite open-loop gain.

Equation 2-1

$V_o = V_{cm}$
$- ((V_{cm} * (R_f/R_{in}))/A_{vol})$

If this looks familiar, it is because it looks a lot like our previous open-loop gain test. In fact, electrically, that is what it is except that we have added the common-mode voltage to the output.

When you looked at the formula (2-1), you probably noticed that there was no common-mode error included. If we add the common-mode error voltage, V_{cme}, into the formula, we end up with this mess.

Equation 2-2

$V_o = V_{cm}$
$+ ((\pm V_{cme} * (R_f/R_{in}))$
$- ((V_{cm} * (R_f/R_{in}))/A_{vol}))$

To get a useful answer requires that we make a differential measurement between V_{in} and V_o. Because the gain error and the common-mode error are added together, the result will only resemble the common-mode error if the common-mode error is much larger than the gain error.

In most cases, just the opposite is true. The CMR is often greater than the open-loop gain of the amplifier. When this happens, the output will just be the open-loop gain signal. As my friend Bob Pease says in his excellent book "Troubleshooting Analog Circuits",[1] "There are still a few op amp data sheets where the CMRR curve is stated to be the same as the Bode plot."

What Should We Do?

A common (do you have the feeling that this word is becoming very common in this article?) method of measuring CMR is to connect the amplifier as shown in Figure 19.

- Use minigator to select R100
- Let R3 = 40 X R12
  - R4 = R12/20
  - R1 = R11
  - R2 = R12
- All Rs should be ±1%
- VCM = VIN X R2/(R1 + R2)
- Verror = VCM/CMRR
- Vout is related to Verror

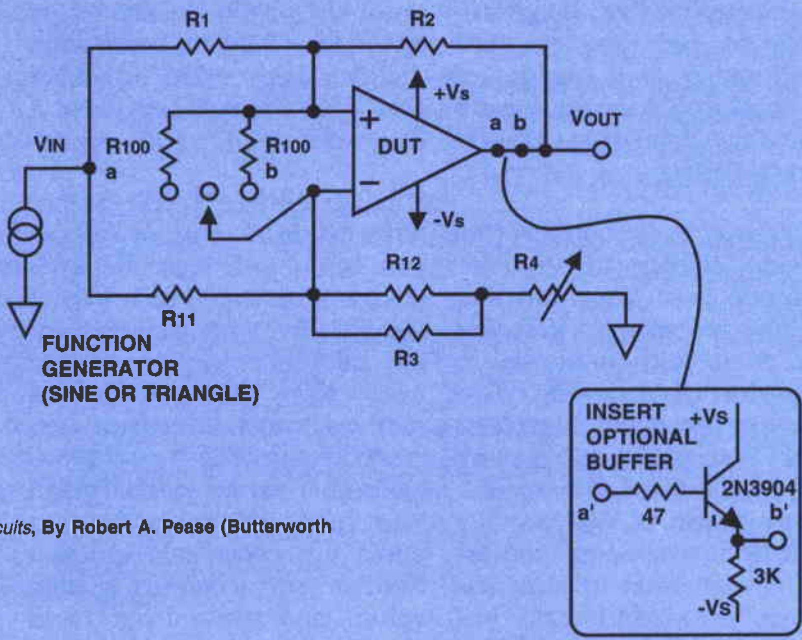

Figure 8.7 from ***Troubleshooting Analog Circuits,*** **By Robert A. Pease (Butterworth Heineman, Boston, MA, 1991)**

At first this looks like a workable circuit. The power supplies, the output and everything else moves up and down together while the inputs stay at ground.

But do they? Isn't this just the same as if the inputs were moved up and down while everything else remained constant? Not quite! If you look closely, you will see that this circuit is very much the same as the previous one. It has one less problem than the last one. In this case, the output does not include the V_{cm} but to do that, it has to move in the opposite direction as V_{cm}. How does it move? By producing a signal at the input that is equal to V_{cm}/A_{vol}. All we need to do is add the R_f/R_{in} term and, lo and behold, we have measured open-loop gain again.

How do we get out of this problem of adding gain error with common-mode error? If we raise the gain to infinity, the gain error becomes zero. Can we do this? Not completely, but we don't have to go all the way to infinity. In Figure 20, we have added another amplifier into the loop. This raises the open-loop gain to a very high value. Probably higher than the CMR. If the gain error is much smaller than the common-mode error, we can ignore it.

Is This the Answer? Is There a Free Lunch?

This circuit (Figure 20) adds the phase shift of the second amplifier to the loop. Some frequency compensation will probably be required and it can be messy to figure out the best component values. This circuit, or minor variations on it, is the one that is most commonly used in production testing.

Is this all there is? Not quite. There is yet another circuit. I have not seen it used in production testing, but as a bench test it looks like a good alternative to the previous circuit. I quote it directly from the Bob Pease book I mentioned above [1].

"Figure 8.7 is a darned fine circuit, even if I did invent it myself about 22 years ago. It has limitations, but it's the best circuit I've seen. Let's choose $R_1 = R_{11} = 1k$, $R_2 = R_{12} = 10k$, and R3 = 200k and R4 = a 500Ω pot, single-turn carbon or similar. In this case, the noise gain is defined as $1 + [R_f/R_{in}]$, or about 11. See pages 100-101 for discussion of noise gain. Let's put a ±11-V sine wave into the signal input so the CM voltage is about ±10 V. The output error signal will be about 11 times the error voltage plus some function of the mismatch of all those resistors. Okay, first connect the output to a scope in cross-plot (X-Y) mode and trim that pot until the output error is very small - until the slope is nominally flat. We don't know if the CMRR error is balanced out by the resistor error, or what; but, as it turns out, we don't care. Just observe that the output error, as viewed on a cross-plot scope, is quite small. Now connect in R100a, a nice low value such as 200Ω. If you sit down and compute it, the noise gain rises from 11 to 111. Namely, the noise gain was $(1 + R_2/R_1)$, and it then increases to $(1 + R_2/R_1)$ plus $(R_2 + R_{12})/R_{100}$. In this example, that is an increase of 100. So, you are now looking at a change of V_{out} equal to 100 times the input error voltage, (and that is V_{CM} divided by CMRR).

Of course, it is unlikely for this error voltage to be a linear function of V_{CM}, and that is why I recommend that you look at it with a scope in cross-plot (X-Y) mode. Too many people make a pretend game, that CMRR is constant at all levels, that CM error is a linear function of V_{CM}, so they just look at two points and assume every other voltage has a linear error; and that's just too silly. Even if you want to use some ATE (Automatic Test Equipment) you will want to look at this error at least three places - maybe at four or five voltages. Another good reason to use a scope in the X-Y mode is so you can use your eyeball to subtract out the noise. You certainly can't use an AC voltmeter to detect the CMRR error. For example, in Figure 8.6, the CM error is fairly stated as 0.2 mV p-p, not 0.3 mV p-p (as it might be if you used a meter that counted the noise).

Anyhow, if you have a good amplifier with a CMRR of about 100 dB, the CM error will be about 200µV p-p, and as this is magnified by 100, you can easily see an output error of 20mV p-p. If you have a really good unit with CMRR of 120 or 140 dB, you'll want to clip in the R100b, such as 20Ω, and then the A (noise gain) will be 1000. The noise will be magnified by 1000, but so will the error and you can see what you need to see. Now, I shall not get embroiled in the question, are you trying to see exactly how good the CMRR really is, or just if the CMRR is better than the datasheet value; in either case, this is the best way I have seen.

For use with ATE, you do not have to look with a scope; you can use a step or trapezoidal wave and look just at the DC levels at the ends or the middle or wherever you need. Note that you do not have to trim that resistor network all the time, nor do you have to trim it perfectly. All you have to know is that when the noise gain changes from a low value to a high value, and the output error changes, it is the change of the output error that is of interest, not really the p-p value before or after, but the delta. You do not have to trim the resistor to get the slope perfect; but that is the easy way for the guy working at his bench to see the changes.

This is a great circuit to fool around with. When you get it running, you will want to test every op amp in your area, because it gives you such a neat high-resolution view. It gives you a good feel for what is happening, rather than just hard, cold, dumb numbers. For example, if you see a 22-mV p-p output signal that is caused by a 22-µV error signal, you know that the CMRR really is way up near a million, which is a lot more educational than a cold "119.2 dB" statement. Besides, you learn rather quickly that the slope and the curvature of the display are important. Not all amplifiers with the same "119.2 dB" of CMRR are actually the same, not at all. Some have a positive slope, some may have a negative slope, and some curve madly, so that if you took a two-point measurement, the slope would change wildly, depending on which two points you choose. (If you increase the amplitude of the input signal, you can also see plainly where severe distortion sets in - that's the extent of the common-mode range.)

Limitations: If you set the noise gain as high as 100, then this circuit will be 3 dB down at F_{GBW} divided by 100, so you would only use this up to about 1 kHz on an ordinary 1 MHz op amp, and only up to 100Hz at a gain of 1000. That's not too bad, really.

To look at CMRR above 1kHz, you might use R100c = 2k, to give good results up to 10 kHz. In other words, you have to engineer this circuit a little, to know where it gives valid data. Thinking is required. Sorry about that.

For really fast work, I go to a high-speed low-gain version where R1 = R11 = 5k, R2 = R12 = 5k, and R100 = 2k or 1k or 0.5k. This works pretty well up to 50 kHz or more, depending on what gain-bandwidth product your amplifier has.

For best results at AC, it's important to avoid stray capacitance of wires or of a real switch at the points where you connect to R100a or R100b. Usually I get excellent results from just grabbing on to the resistor with a minigator clip. You can avoid the stray pf that way; if you use a good selector switch, with all the wires dressed neatly in the air (which is an excellent insulator) you may be able to get decent bandwidth, but you should be aware that you are probably measuring the AC CMRR of your setup, not of the op amp.

I was discussing this circuit with a colleague, and I realized the right way to make this R100 is to solder, for example, 100Ω to the + input and 100Ω to the - input, and then just clip their tips together to make 200Ω-balanced strays, and all that.

If you have an op amp with low gain or low g_m, you may want to add in a buffer follower at a-b, so the amplifier does not generate a big error due to its low gain. The National LM6361 would need a buffer as it only has a gain of 3000 with a load of 10kΩ, and its CMRR is a lot higher than 3000.

Altogether, I find this circuit has better resolution and gives less trouble than any other circuit for measuring CM error. And the price is right: a few resistors and a minigator clip."

So this is the way Bob Pease suggests we measure CMR. I don't know why it is not more commonly used.

What, only one test in this whole article? Well CMR is a messy measurement and we have only discussed the DC measurement. Later we can play with the AC measurement.

This may help explain why so many digital engineers get into trouble when they do analog designs. It's not just ones and zeros.

[1] Robert A. Pease, *Troubleshooting Analog Circuits*, Butterworth Heineman, Boston MA, 1991.

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