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BOB PEASE ON ANALOG
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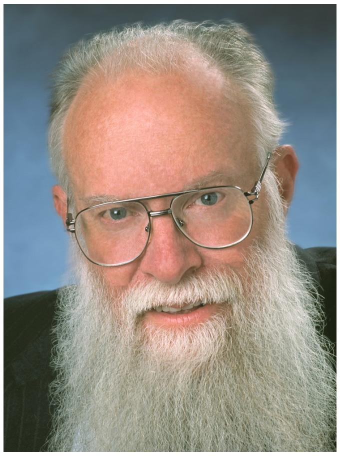

A compendium of technical articles
from legendary *Electronic Design*
engineer **Bob Pease**

A focused
collection of
columns from
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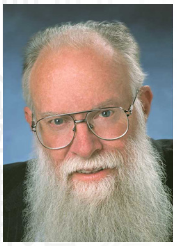
author **Bob Pease**



CHAPTER 1:

What's All This Analog Stuff, Anyhow?

BOB PEASE, Senior Analog Engineer

First published September 13, 1990, this is the first of a series of columns about analog and "linear" circuits written by Bob Pease, Staff Scientist at National Semiconductor Corp., Santa Clara Calif. We think our readers will get a lot out of Bob's seemingly off-the-wall, yet insightful views of the engineering world.

Why? Why am I going to all the trouble of writing about "linear" and analog circuits? Everybody knows that linear circuits are dead. Nobody's buying or designing in linear circuits; they are all being replaced by digital signal processors. Analog computers have been dead for years. Why bother?

Well, these days, even though there are tends to perform a lot of functions with digital computations, people are finding that there are still a huge number of things that cannot be done properly without analog circuits. It's true that some of the trendy new radios claim to use a lot of digital techniques, but even there, the receivers and amplifiers are analog circuits—even if the receiver's frequency appears to be digitally controlled.

When people are designing digital computers, they need analog techniques to make good layouts for fast buses. They need power supplies—either linear ICs or switch-mode circuits (which use analog circuits internally). And, as for us analog designers, the old-timers and the rookie engineers—well—this column is intended as a soapbox for me to talk about linear circuits, and then for me to listen to your opinions and comments and questions.

I have a lot of opinions, but I'm also very interested in what makes you tick. I may not be the smartest engineer in the whole analog jungle, but I have sort of volunteered to start writing this, and we'll see what happens—what interesting debates we get into. I have a bunch of opinions about Ics, data sheets, testing, computer simulation, education, troubleshooting, along with a whole slew of little topics.

In every darned issue of *Electronic Design*, I'll try to have some provocative or insightful topic. Some will be pretty technical, others will be more philosophical in nature. But one thing's for sure, I'll try not to bore you. For example: What's all this heuristic stuff, anyhow?

Heuristics?

The other day I was talking with a young college graduate from a prestigious Eastern engineering school. He explained that his specialty was analog synthesis. I perked up my ears—I hadn't heard much about that. Where could I read more about this? "Oh," he said, "in some of the IEEE journals." Hmm. He started to explain the approach. It's a heuristic approach, he said. Hmm. What's a heuristic?

He said, "You don't know what a heuristic is? Really?" I explained no, that we didn't have any heuristics when I was in school.

(Note: Mr. Webster says that heuristic means "serving to guide, discover, or reveal; specif.: valuable for stimulating or conducting empirical research but unproved or incapable of proof—often used of arguments, methods, or constructs that assume or postulate what remains to be proven or that leads a person to find out for himself.—from the Greek, heuriskein, to discover, find.")—Gee, that sounds a lot like analysis or optimization to me—not synthesis.

The young man explained that when you make a log of optimization experiments, heuristic refers to the starting place, the initial guess. H'mm. He said, "You feed in some requirements and some specifications, and it optimizes the performance." Hmm. Now, what circuit does it use? "Oh, it uses the circuit that you give it." Hmm.

The Key Question

If you give it a circuit that doesn't work well enough, how does it generate a circuit that works better? "Oh, it doesn't." I explained to this young fellow, that in our whole product line, about 99% of the circuits are not optimized at all—at least not "optimized" in the sense that he understands. If you really OPTOMIZED them, they would all be a little different than they are now. But each one has a different circuit that is a revolutionary—not just an evolutionary—change from any pervious circuit. So there may be places in our company where optimization is useful and a good idea.

But I wish he couldn't call it "analog synthesis," that seems to be a misnomer. The circuits around our area—the ones in the NSC Linear data books (and, I bet, in the PMI and Analog Devices data books, too), where not "synthesized" except by bright engineers who knew that the old circuits wouldn't cut it, and a new circuit was needed. Good luck, young fellow!

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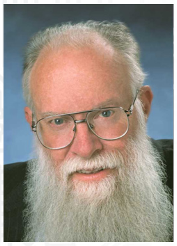

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CHAPTER 2:

What's All This Noise-Rejection Stuff, Anyhow?

BOB PEASE, Senior Analog Engineer

Bob Pease describes circuit solutions that solve common-mode noise problems.

IT'S WELL KNOWN THAT audio power amplifiers like to get a good set of grounds, or noise around the inputs may not be rejected properly, causing hum and buzz. So when a guy called me asking how to clean up his interface from his clean audio signals to his LM3886 power amplifier, whose ground system was pretty noisy and lumpy, I thought for a second and replied that the solution was easy (**Fig. 1**)!

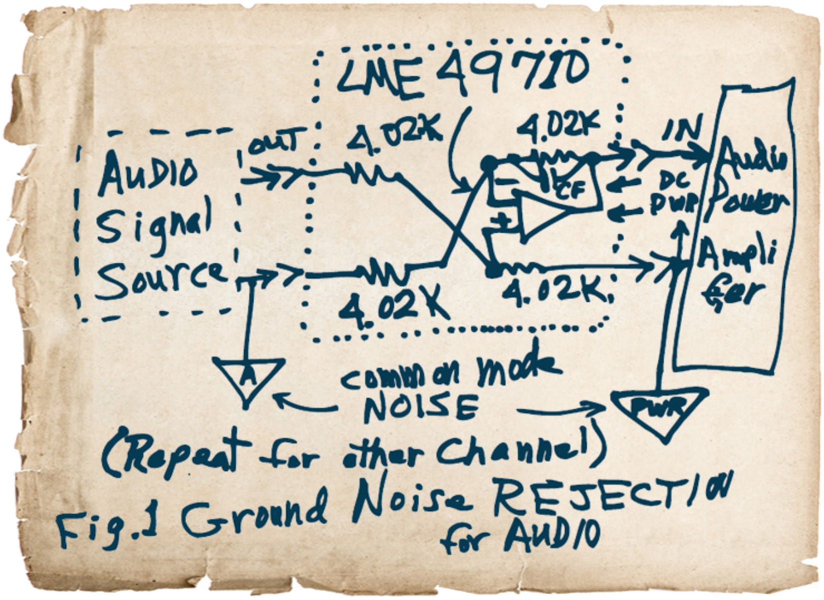

Fig.1 Ground Noise REJECTION for AUDIO

The set of four 4.02-k resistors (1% R's but matched to 0.1%) make up an adder-subtractor circuit—straight from Philbrick's 1955 Applications Manual. Let's stand at the power ground for our scope ground. If there is a quiet signal "out," referred to its analog ground "A," fine. But we must beware of any common-mode (CM) noise, which would appear between the A ground and the PWR ground. This noise might be related to any kind of ambient noise, or noise on the power amplifier's supplies, or (in a vehicle) automotive transients and radio-frequency interference (RFI). I told this guy, "This circuit will reject any and all such noises." The old equations give you:

$$\text{Amplifier } V_{IN} - (V_{PR}) = V \text{ (signal out)} - VA \text{ ground}$$

I recommended an LME49710, which has 50 MHz of bandwidth and good clean gain and linearity and noise rejection, per www.national.com/an/AN/AN-1671.pdf. I recommended a "gimmick" of teflon twisted pair as CF, starting at 6 in. and unwinding it as needed. I mean, we don't know exactly what kind of noises there will be, or how much of the wiring strays. Still, the op amp should be supplied with reasonably quiet power that is tied to the PWR ground. The 4k resistors do not have to be closely matched to ensure gain accuracy—but to help give you good common-mode rejection ratio (CMRR), much better than 1% R's will do. If you wanted 80 dB, you'd have to trim them.

Opportunity

After I hung up, I could not stop thinking of this "problem = ~ opportunity." Why does this look familiar? I thought about some of the problems my colleague Nick Gray had been trying to solve over the years. Hey! this looks just like the problem with analog signals that need to be sent to an analogto- digital converter (ADC)! You have an analog ground plane and a digital ground plane. But if you try to just strap the grounds together, you'll get absurd noises.

And if you have one ADC, the CM noise rejection can be bad. But if you have one analog ground plane and one big digital ground plane that are serving two or four or more ADCs, the CM noise coupling can be horrible! What's a mother to do?

The adder-subtractor shown in **Figure 2** will reject the CM noises, neatly, using one adder-subtractor per ADC channel. The clean, quiet voltage that is sent in between the signal and VA will re-appear at the adder-subtractor's output, referred to power ground, for each channel. The old equations give you:

$$V_{OUT} - (V_{DIGITAL\ GND}) = V\ (\text{signal out}) - \text{VA ground}$$

Now, the LME49710 mentioned above may have a signal bandwidth of just 20 MHz, plenty for audio or for some ADCs. It may be trimmable to a CMRR of 70 dB out to 2 MHz. That's what I saw when I actually built it. But what if you need a fast ADC and a lot of CMRR versus frequency?

A Key Modification

Let's swap in the LME49713, a fast current feedback amplifier (CFA), which will pass signals up above 90 MHz, and we may be able to get decent CMRR and noise rejection out past 20 MHz. (The resistors have been cut to 2k to make sure you can get full bandwidth. It might run even a bit faster if you chose 1.2k.)

The LMH6714 can go out to even faster, 400 MHz. But the '49713 may have better linearity. Hard to tell. When you need to do precision work plus fast bandwidth, everything has to be engineered and tested. The circuit

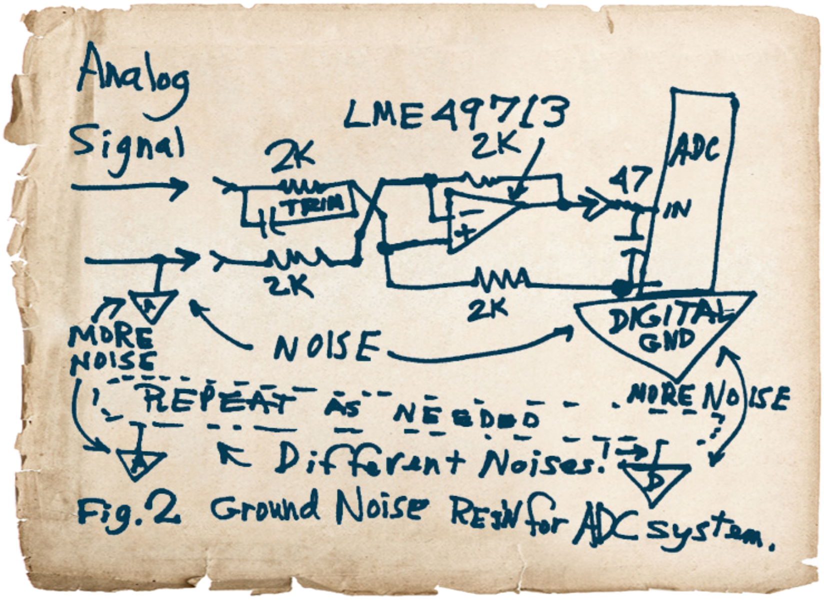

Fig.2 Ground Noise Rej'n for ADC system.

might change slightly if you need fast, clean step response rather than just wide bandwidth for sines.

Will that be fast enough for a pretty broad-band ADC? Well, that will depend on the actual circumstances of your system needs. I mean, you could always have more CM noise than this adder/subtractor could quash. I did my testing with 10 V p-p. Could this be used in addition to a Balun? Insert that ahead of the first two 2k resistors. Probably. I mean, when things get fast, then you always have to be prepared to do some real engineering. Still, this is one good tool to add to a good toolbox, when you have to accommodate (and reject) many kinds of nasty noises, conducted and induced from radiated noise.

For this case, adding a few pF of Cf seemed to be doing more harm than good, so I installed several inches of "gimmick" = twisted pair as a capacitive CMRR trim, for the signal path going to the positive input.

What's the big deal with the "adder-subtractor"? When George Philbrick developed the K2-W, which was one of the first operational amplifiers with differential inputs, it facilitated simple adder-subtractors that did not need a dozen resistors and two or more chopper-stabilized amplifiers and hundreds of watts. George never had access to any 400-MHz op amps as we can easily buy these days. He'd be impressed with modern op amps—and applications circuits.

Can these amplifiers provide a voltage gain other than 1.0? Sure—and you have to engineer it. And to get good results, you always have to plan a good layout. This is just a start, to indicate all the things you can do.

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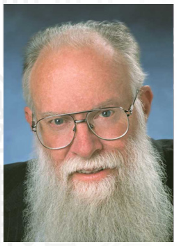

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CHAPTER 3:

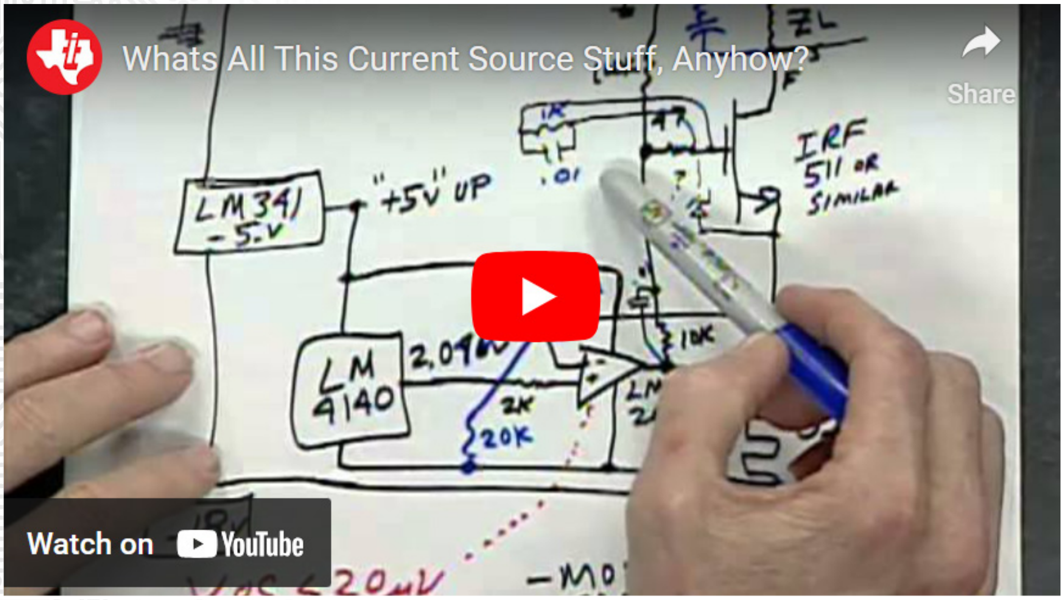
What's All This Current-Source Stuff, Anyhow?

BOB PEASE, Senior Analog Engineer

If you need a current source, and you don't know where to look, it is not easy to find advice on how to make them.

Recently, a guy asked me how to draw a constant 1.00 mA from a node of a circuit. Of course, he did not tell me what volts, ohms, or frequency. But, he admitted, he basically did not know how to design a current source.

So I'm sorry to waste the time of all you guys who do know how to design a current source. But maybe this lecture can help and save you some time so you don't have to teach all the young kids. If you need a current source, and you don't know where to look, it is not easy to find advice on how to make them. I looked and could not find valid advice on how to do this! So, here you go.



Editor's Note: The video is one of The Bob Pease Show episodes that was recorded in the same general timeframe.

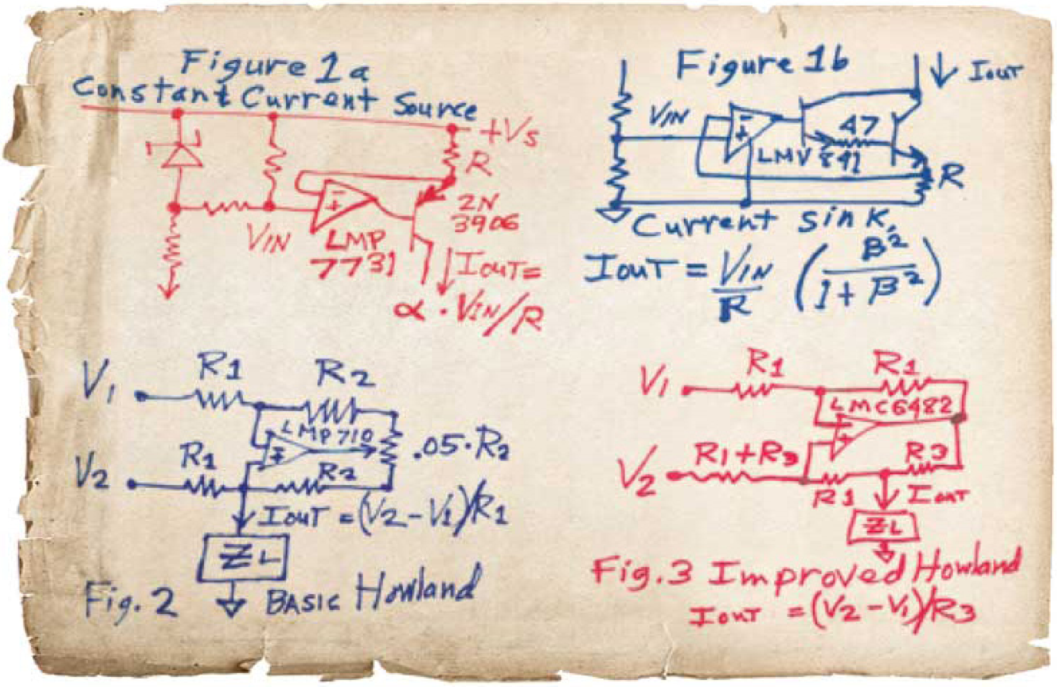
Step-by-Step

Figure 1 is a basic (unidirectional) current source that can spit out any positive current you want. You want 10 or 100 µA? 10 or 100 mA? 10 or 100 nA? 10 or 100 A? Be my guest. It does a good job, putting out current in one direction—but not both.

The current sourced is $I = V_{IN}/R$ (**Fig. 1a**). Of course, you need an op amp whose common-mode (CM) range extends to the right voltage and whose IB is small enough. For current sourcing, you need an amplifier whose CM range and output go approximately to the positive rail, and you need PNP transistors. Sometimes you can arrange it so the output does not have to go too close to the rail, using a resistive divider. The transistor is shown in a nominal way.

For 1-mA full-scale current, an ordinary 2N3906 can give you a Z_{OUT} of 50 MΩ. If you want a really high Z_{OUT}, like 1000 M Ω, you might put in a Darlington (**Fig. 1b**). Or for large currents, a Trarlington could be justified. If you need to put current into a fast-moving signal, you might need to add some extra cascoding to I_{OUT}.

For sinking current, you need an amplifier whose CM range (and output swing)



extends to (or near) ground, or $-V_S$ (Fig. 1b). And, you need an NPN transistor. This is often called a constant current source.

Well, it does not have to be absolutely "constant." It can be "modulated" or adjusted by changing the V_{IN}. You could put in some ac. But don't allow the current to get "modulated" to zero, or you might get some strange response from the unhappy amplifier.

A Howland current pump can put out positive or negative current—or zero (**Fig. 2**). Then there is the "improved" Howland current pump (**Fig. 3**). Both are wonderful when you have defined what ranges of V and I you want. Neither one has a great weakness, depending on what you want. If you want the output to go close to the rails, the "improved" Howland can usually be arranged to swing closer.

Some Extra Info

I recently did a complete analysis of the Howland circuits, and I wrote it up as an Application Note, AN-1515, at www.national.com/an/AN/AN-1515.pdf. I included some notes on trimming the resistors, because many times the Howland is just presented as a nominal circuit, with no trimming indicated. But to get high $Z_{OUT,}$ you usually do have to trim.

For a 1-mA output (all R's = 10k ±1%), the Z_{OUT} might be as poor as 0.25 MΩ, +0.25 MΩ or even –0.25 MΩ. So, you have to trim. (See the little trim on Figure 2. Even 0.1% resistors would only provide a moderate improvement, to ±2.5 MΩ.) That App Note also shows how to get high Z_{OUT} without any pots.

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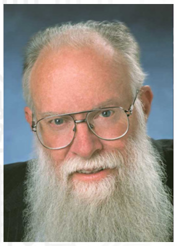

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CHAPTER 4:

What's All This Transconductance Stuff, Anyhow?

BOB PEASE, Senior Analog Engineer

Bob debates a reader's assertions regarding the beta and transconductance of transistors.

Recently, I received this letter: *Sir: I haven't written you in a long time since you give me the "Taguchi Treatment." This is the typical "professional correctness" approach in our industry.*

I was being laughed at for calling for reduced voltages applied to the output circuits of active devices OVER thirty years ago; today, two to three volts or less is becoming standard in computing. I was able to get a two-to-one change in output current in FETs, with 18-mV change in gate voltage, and with drain supplies of a few tens of millivolts applied (20 mV, for example). I knew the physics of how to do this even then.

The LM4250 op amp made by NSC is an excellent example of why this works (+/–1.5 V supply voltage). Unfortunately, I need some more of them. In measuring currents with bipolar devices, it is desirable that the current be measured with a voltage burden of not over about 5 mV. Which means that a semi-precision op amp capable of boosting a voltage of 0.5-5 mV to 20-200 mV, with a precision 5-10%, and an offset error at the input of probably not more than 0.5 mV is needed. Where can one get such beasts (instrumentation amplifiers)? Note that a burden of 200 mV measuring the current in the collector of a bipolar transistor increases device voltage gain by about 7.5, which can be enough to generate instability occasionally. Clearly, we need to be more open-minded in our profession.

Very truly yours, Keats A. Pullen

I was compelled to reply: Dear Mr. Pullen, I don't criticize anybody because of political correctness, or the lack of such correctness, but because of a lack of technical correctness.

I have noticed that for many years, you have strongly criticized engineers who think that the beta of transistors is important. In your mind, that may be a matter of political correctness, or technical correctness. BUT in many technical areas it is important to have bipolar transistors with good, well-controlled beta. In other cases it's important to have well-matched beta. If you want to argue that's not true, let's talk. But, if you persist in your insulting of any engineer who talks about "beta," and who uses beta to analyze circuits, I shall rebut you with great vigor.

I will even be happy to send you some LM4250s, as free samples, but you should be cautious because the input transistors have high beta and are matched.

The internal transistors have high, matched betas, too.

I was reading recently about the "good old days" of 20 years ago, when many people were building digital circuits, computers, calculators, etc., using FETs whose thresholds might shift and move several volts over their lifetime. So, 9-V power supplies were needed to make systems that would not stop working right away. You could demonstrate that a 2-V supply was enough on any given day, but the product would not keep working. Stability of operating bias isn't trivial today and it was a serious deal 20 years ago. A scientist might be called a person who can show something working once. An engineer must figure out how to make products that will last considerably longer than the warranty period.

I will certainly agree that the trans~conductance of transistors is very important. We usually take it for granted because it's so consistent. But you should not say, as you usually do, that we engineers don't appreciate the g_m of devices just because we don't talk about it a lot. We take it for granted.

However, I have heard your claims, over the past several years, that FETs have a transconductance so good that you can get a two-to-one change in output current for an 18-mV change in the gate voltage at room temperature. Of course, that's the same as 60 mV per decade, and is equivalent to saying that the g_m/mA is 38.6, which is a well-known theoretical maximum amount of g_m for bipolar transistors at room temperature.

Okay, Mr. Pullen, I'll call your bluff: on EXACTLY what FET device that you have bought, or borrowed, or fabricated, have you ever measured a transconductance as good as 18 mV per octave? Or, have you seen anybody else measure such a FET? I agree that 18 mV is a theoretical limit, never to be exceeded. No FET device ever made is that good, whether MOSFET or JFET. Not just by a few percent, but by a factor of 2 or maybe, at best, 1.5. So you won't find an octave change of drain current per 18 mV; you'll find, perhaps, at best, a factor of 27 or 36 mV per octave----not NEARLY as good as 18.

I asked a large number of my friends: What is the best published data, in all of the technical literature, that indicates the transconductance of any FET----MOSFET or JFET----is as good as XX mV per octave of output current? Some of them said that the g_m per mA can get extremely high, considerably *higher than* g_m= 38.6 X I. That's if you trust your computer simulation, and if you operate the transistors at extremely low density down to the sub-threshold region, such as 0.1 μA through a transistor 1 micron long by 5000 microns wide.

I explained to these people that the story their computer tells them is untrue, because their computer is using an oversimplified model for FETs, with bad accuracy at starved levels. At high levels, every time you decrease the drain current by a factor of 4, the g_m/ I does improve by a factor of 2----but that runs out of gas before you get to g_m= 38 X I. The g_m doesn't ever exceed 38 X I; in fact, it never attains 38 X I, yet does approach it. But not very closely. Other people knew that g_m/mA approaches 38.6, and doesn't exceed it----but they agreed that they didn't know how closely, nor why. Nobody recalled reading any "best published value." Several guys pointed out that they have seen 100 mV per decade on big MOSFETs operating below 0.01 μA. That's about 30 mV per octave. Nobody had ever seen this number in the 20s.

I was helping to interview a young engineer recently, and I bounced this question off him, not expecting an answer. He simply explained that there's a virtual gate, under the gate oxide of a MOSFET. If the capacitance of this gate to ground (or, to V_{source}) is relatively large, the capacitance of the C_{0X} would cause a capacitive voltage-divider effect, so the virtual gate never sees all of the gate voltage's change. In this case, the g_m/ I will typically be reduced below 38.6 by a factor of C_{0X}/ (C_{0X}+ Cgs). This factor rarely gets better than 0.6. Even in cases where the gate oxide is very thin, as thin as 50Å, the capacitance from the virtual gate will still be almost as big as the C_{0X}. So the g_m/ mA will be poor for devices with thick gate oxide----and it can get better when the gate oxide gets thinner. However, the g_m/mA never gets very close to the maximum theoretical value. Has anybody seen better than 30 mV per octave on a MOSFET?

And JFETs (junction FETs) aren't very good, either. I measured some 2N5486s----a modern, high-performance N-channel JFET----and found the $g_m$as good as 25 mV per octave, but no better, not even at starved levels such as 0.1 nA. JFETs don't give high g_m/ mA, because the back gate has inferior sensitivity to the front gate. If you use a tetrode FET, with the back gate brought out separately from the front gate, the gm is even worse.

I once saw some JFETs that had g_m about 38 X I, because they were implanted with such a light channel that they wouldn't conduct any current until they were forward-biased. Then they ran like a bipolar transistor, with injection and with finite base current, too. They were Enhancement-Mode JFETs.

Well, after this interview, the first thing I told my boss was, "We want to hire this guy who understands the g_m limitations of MOSFETs." And then I wrote down this letter.

So, Mr. Pullen, if you know where to measure any FET with a millivolts-per-octave ratio better than 25, please let us know exactly where and how we can repeat this experiment.

Yours truly, /Robert A. Pease

p.s. Operational amplifiers with offset voltage better than 0.5 mV, such as the LM308A and many others, have been around for over 20 years. If you just connect a trim pot and optimize the offset, it's not hard to trim an op amp's offset within a few microvolts of zero. These days, offsets less than 0.025 mV or 0.010 mV, on an OP07-type amplifier, are not news----these have been around for at least 10 years. So, I don't see what your problem is. /RAP

Then I mailed the letter.

When I was a youngster in Connecticut, and knew almost nothing about electronics, I recall, VERY PLAINLY, reading in the Hartford Courant (about 1958), that somebody had invented a "spacistor" that would put out its current better than any existing transistor. Well, what the heck does that mean?

After about 15 years, I realized that these guys were probably talking about the first JFETs. All very good. And to this day, JFETs do many tasks very well. And MOSFETs, by the hundreds of billions, also do their jobs. They have small input currents. But, while their g_m/mA is *adequate*, it's not astonishing. Does anybody remember the "spacistor?" Who made it?

When I began measuring real transistors back in 1961, I was really impressed with the bipolar transistor's wide range, in which the transconductance is nicely proportional to the IC , from 1 mA to 1 nA, or even lower. Hey, that's pretty impressive. And I have to concede some more respect for the late Bob Widlar, for showing that a transistor's g_m can work even when the emitter voltage is higher than the base voltage, on an npn (bipolar) transistor. You can measure this on any germanium transistor at room temperature, or on any good npn at about 250ºC. So the transconductance of a bipolar transistor is a very useful, well-known, and predictable quantity. In fact, it's only when the g_m gets *worse* than 38.6 X I that we take notice.

Ah, yes----when we work with transconductance, why do we talk about "g_m?" Well, in basic circuit theory, just as the resistance of a circuit is labeled r, the conductance of a circuit is labeled g. And when vacuum tubes came along, their gain, $\Delta I_{out}/\Delta V_{in}$, was in the form of a *conductance*. Furthermore, this was considered a mutual conductance----the voltage was applied between the grid and the cathode, and the output current flowed between the plate and the cathode. Vacuum tubes thus were characterized in terms of their mutual conductance, or g_m , at a given bias. When transistors came along, we kept the same term. So that's why we have the term g_m, and we still use it because it's much more compact than the phrase "transconductance."

Keats Pullen showed me an old clipping from a 1966 *IEEE Proceedings*, indicating that he had measured the g_m of some of the first experimental JFETs he ever saw, back in 1964, and he claimed they did have 38.6 milli-mhos per milliampere. But there wasn't any technical information on the device, no type number or model or part number----not even the manufacturer. In other words, this was not exactly a reproducible experiment.

More recently, Mr. Pullen argued that if we're just able to build FETs out of *perfectly pure silicon,* it stands to reason that the g_m will behave perfectly. Sorry, but that kind of appeal to passion falls apart when you remember that you won't have any transistor at all, unless the "perfectly pure silicon" gets doped.

Furthermore, Mr. Pullen argued that vacuum tubes also have a g_m of 38.6 X I. But this g_m is only applicable at 25º C, where the value of q / kT is 38.6. At higher temperatures, g_m falls. Why would a vacuum tube's g_m not correspond to the much hotter temperature of the orange-hot cathode, or the warm space-cloud of electrons swarming around it? In fact, a vacuum tube's g_m / I is rarely better than 4, which is a long way away from 38.

Then I got another letter from Mr. Pullen, pointing out that the millivolts per octave of these 1964 FETs was not 18 or 19----but 20 mV in one case, and 25 mV in the other case. All of these years, Keats has been telling us that the millivolts per octave that he had measured was 18, when actually it was 25 or 20.

I went back and measured some 2N4393s (a basic 30-Ω switch). They had 21.5 mV per octave----a considerable ways away from 18----but not much different from the ones Keats saw in 1964.

Keats Pullen has always insisted that FETs could show an octave change of $I_D$per 18 mV. Finally, he had to admit, when pressed, that the ones that he thought were best weren't actually as good as 18 mV per octave, but 20 mV. Well, if you mean 20, say 20, not 18.

Now, everybody I talked to said that JFETs were worse than MOSFETs, and the millivolts per octave was always worse than 25. And I was sure of that, too. And they were wrong. And I was wrong. Just goes to show... "what everybody knows" can be wrong. There's nothing like a real experiment, with real tests and real data, to puncture "what everybody knows." If any reader can tell me of any published or unpublished data, preferably from an experiment we can reproduce, showing gm better than 20 mV per octave, on any MOSFET or JFET, at 25ºC, I'd be interested to hear about it. Any theoretical analysis that explains the reasons why a JFET's millivolts per octave never gets too close to 18 would be greatly appreciated, too. Is that limitation due to doping, or geometric factors? Or something else?

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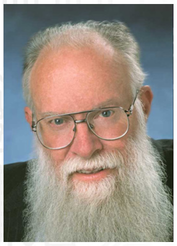

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CHAPTER 5:

What's All This Absurd Filter Stuff, Anyhow?

BOB PEASE, Senior Analog Engineer

This morning, some darned kid engineer (rather bright for a 15-year-old!) asked me if I could show him anywhere he could add a little narrow-band boost or cut to a circuit for an audio signal, perhaps with a Q of 4 or 5. I told him sure, I could find a cookbook circuit in any number of audio handbooks, such as the recent reprint of the old 1976 NSC Audio-Radio handbook, available at www.AudioXpress.com for about $16.

Then he asked me if he could have this circuit with an adjustable center frequency. I mean, okay, some dynamic equalizers do many things well. But a person might want the center-frequency to be adjustable for a special case.

Oh?

"Cut that out!" I told him. "Nobody has ever seen that or done that as an adjustable-frequency filter! And it's the most silly-assed question I've seen all week!! And it's absurd and impossible..."

Then I pondered this and told him, "But, on the other hand, I just remembered that I do know how to do it. There's probably nobody else in the world who knows how to do it, but I can do it. And changing the center frequency doesn't goof up the gain or the bandwidth."

Doing the Absurd and Impossible

Take that bandpass filter on page 237 of my orange-cover book, Analog Circuits (World Class Designs), the one where trimming the F_{Center} does not goof up the gain or BW, per the figure. Set up your desired BW. Get it running. The best thing about this bandpass circuit is that the bandwidth depends only on R3 and the C's: $BW = 1/(\pi \times R3C)$.

The gain depends only on $R3/(2 \times R1)$. If $R3 = R1 \times 2$, the gain is 1. No matter what it looks like, R2 is not a trim on gain. It has no effect on gain or BW. R2 is just a trim on the center frequency. The center frequency depends fairly accurately on $1/[\sqrt{(R2 \times R3)} \times (2\pi C)]$. So, the R2 resistor has sort of a square-root effect on the center frequency. It's not really linear, but it's usable. (You could use a logarithmic pot for this if you want to.)

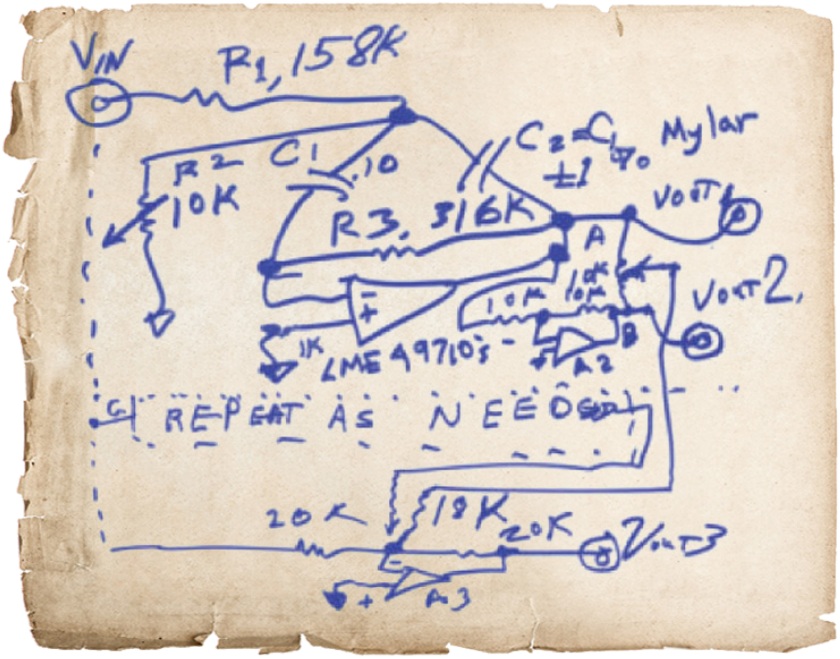
Note that I did say that the BW is not affected by R2. But if the center frequency changes, and the BW does not, then the Q can change somewhat. If you vary the center frequency a bare half octave, up or down, the change in Q may not be too bad. If that's okay, it's okay. If not, you might use a multipole double-pole double-throw (DPDT) switch to switch some capacitors out and others in. Be my guest. But it's worth a try, I think. It depends on what kind of band-pass filter you are trying to make. Does this circuit do the job okay? Do you need absolute precision? Put in matched polys rather than cheap mylar.

Now, take this filter's output (A) and shove it through a unity-gain inverter (A2) to make (B) = (–A). Run a pot such as 10k from A to B. Send the wiper to a 20k resistor to a summing point of a third amplifier A3. Send a suitable fraction of A through 20 or 18k to the summing point. So now the R2 that goes to ground can tweak the center frequency without goofing up the rest of the filter, maybe over a 2:1 or 3:1 frequency range. And, the new 10k pot can give you boost (+6 dB) or cut (–36 dB or more).

Not bad. I had to save this one to publish as a column. And I first wrote up that basic band-pass filter in 1971, 39 years ago! I still have a good copy of that Philbrick Applications Note, and it had no errors in it. It is still a useful bandpass filter, and it can now help you make a notch filter, which I did not realize at the time.

If you are trying to "add" a little cut (or boost) to some audio circuit frequency response, try it. It might just work well. Are you trying to notch something out? How does it sound? This is such a small, cheap, simple circuit, you could add in two tuned circuits to subtract harmonics for higher resolution.

The values shown are for 50 or 60 Hz. If you want to add a second filter for 120 or 180 Hz, as shown at Fig. C, change the R's or the C's and be my guest.



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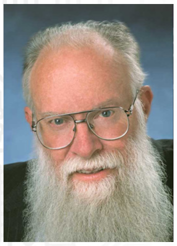

CHAPTER 6:

What's All This Time-Domain Stuff, Anyhow?

BOB PEASE, Senior Analog Engineer

Bob Pease column on thinking about amplifers in the time domain, with examples.

Last night, I was attacking a thorny problem and thought about the time domain. I think about circuits, as an engineer, in the time domain. When something happens, or changes, then something else can happen—or may start to happen. Is that something that I like? Or is it something I don't like?

I have used this analysis many times, as in "What's All This Fuzzy-Logic Stuff, Anyhow? (Part 4)" and "What's All This Ball-On-Beam Balancer Stuff, Anyhow?" *(Nov. 20, 1995,)*. I know some engineers who like to work in the time domain and some guys who like to work in the frequency domain. We have different kinds of heads. We may each be able to solve a problem, but from completely different angles. Different strategies. And sometimes we have to collaborate. That can be fun! I mean, I am not completely ignorant of the F domain, but I rarely find it helpful.

I know a lot of good engineers who work primarily in the time domain. Often we can solve some problems that the frequency-domain guys have trouble with, such as the ball-on-beam balancer (BOBB) and the fuzzy controller for steam boilers. I get insights that the fuzzy-logic guys and the F-domain guys don't.

Key Questions

Several years ago, a guy asked me, "When an LM308 has its dc gain increase, don't you get in trouble when its ac gain increases proportionately?" I asked him where he got that notion from. He said he read it in a book. I told him to drag out that book and X out that idea.

I explained that the gain-bandwidth of any modern op amp (designed in the last 40 years) is invariant of the dc gain. He said his simulations did not show that. I told him his simulations and models were just wrong. The book was wrong.

Then I asked him if he ran a simulation of an LM108 with high gain (–500,000), another one with low gain (–50,000), and another one with reversed gain (+500,000), what if the simulation told him some of them would not work well? What if he ran the amplifiers and they all worked well (as I am sure they would)? Which would he believe, the simulation or the silicon?

He did not know how to answer my questions. He went away. He never came back. I hope he believed the real amplifiers. A few weeks ago, I bumped into another guy who still believed that:

$$A_V = A_{VO} \times 1/(1 + s \times F_O)$$

where F_O is the purported "low-frequency rolloff" frequency. Even at Philbrick, we used to say that. Even when we were wrong. Even when we should have known better. For him, I cooked up a better expression. For mid-frequencies, it is fair to say:

$$-V_O = 2\pi f_H \int V_{IN}\, dt$$

which is the same as saying:

$$-V_{IN} = p \times V_{OUT}/2\pi f_H$$

where f_H is the gain-bandwidth product or the unity gain frequency. Or if you want to add a second high frequency roll off near f_H, that's easy. But for the low fre-quency rolloff, the correct way to look at it is:

$$-V_O = 2\pi f_H \int V_{IN}\, dt \times \frac{(p \times fA_{VO} / 2\pi f_H)}{(1 + (p \times fA_{VO} / 2\pi f_H))}$$

The default value of gain when f gets very small becomes A_V = ~A_{VO}, as the other terms cancel out. But the low-frequency "break frequency" moves around as A_{VO} changes. It's $F_O = 2\pi f_H/A_{VO}$, and that's okay. The frequency-domain guys can analyze this any way they want to. The fuzzy-logic guys can analyze it any way they want to. But I have a bunch of friends who have sold several *billion* op amps, and we are *right*, and most frequency-domain guys are *wrong*, about how to describe an op amp.

If the dc gain goes up to 10 million, or more, that's not really bad. The f-3dB could fall to 0.1 Hz, or lower, but that does not mean that the amplifier's response will have long settling tails at 0.1 Hz—as I pointed out in "What's All This Output Impedance Stuff, Anyhow? (Part 2)" *(Aug. 28, 2008).*

Am I any expert on poles and zeros? Uh-uh. The frequency-domain guys have those tools. They like to use those to solve some problems that I would probably have trouble with. I prefer to solve those problems in the time domain. I like to use p = d/dt. The derivative operator. In linear systems, in the frequency domain, p = s = 2πj(f), but I won't waste much time with that. How can I sell you on the time domain? Where can you learn more? I dunno. More later.

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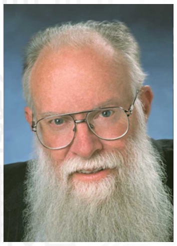


CHAPTER 7:

What's All This Multiplier-Divider Stuff, Anyhow?

BOB PEASE, Senior Analog Engineer

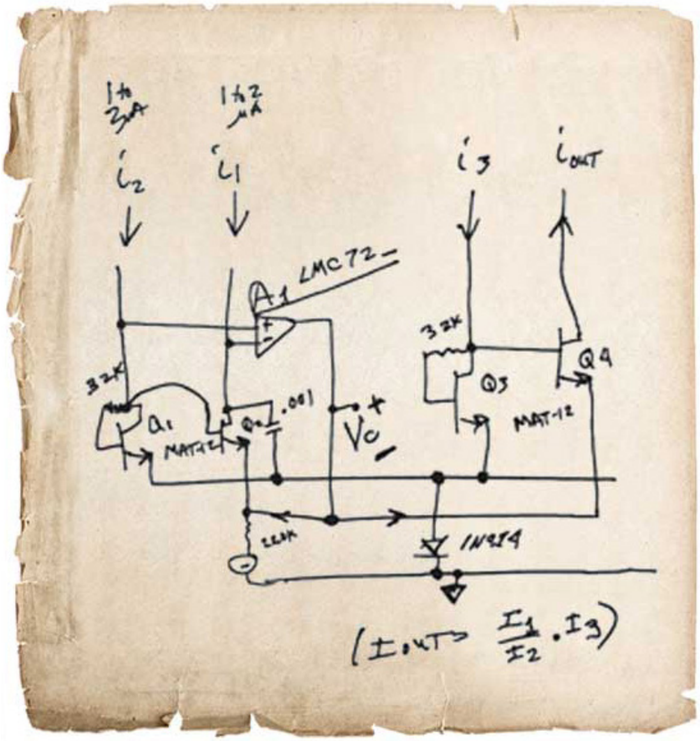
A few weeks ago, an engineer knocked on the front of my cubicle and asked, "Can you recommend a good design for a multiplier-divider?" I just happened to have my Linear Apps book open to AN-31, so I said, "Like this one?" (see http://cds.linear.com/docs/Application%20Note/an31.pdf)

She said, "No, not exactly. That runs on ±15-V supplies, and I need one to run on +4 V." Oh. So the engineer showed me the specs she needed: IOut = (I1/I2) × I3. The I1 and I2 would cover barely an octave, and I2 could be bigger than or smaller than I1. And I3 would be just a few µA.

After pinning down the process and a few more details, I told her, "Go away, and see what I got for you tomorrow."

Time to design

Anyhow, I cobbled together this basic idea. It was still adding and subtracting logs, but it didn't need so many op amps—like one instead of four. A week later, I asked her how it was coming. She replied, "Oh, we decided we didn't really need that...." I explained, "Well, I gave that to you, and if you aren't going to use it, I want it back." So I took it back. And here it is (**see the figure**).



To take a ratio of I1 to I2, we just need to put those currents through a pair of well-matched NPNs, such as Q1 and Q2. I put in a servo amplifier A1 to establish a control voltage, VC, which will be 18 mV per octave at room temp. VC can be positive if I1 is bigger than I2—or smaller if vice versa. This VC is also applied to Q3 and Q4. The ratio factor established by VC will scale IOut as needed. Piece of cake.

Normally we would like to use the National LM394 for the matched pairs. But since some foolish people obsoleted them, we'll just the next best thing, the MAT-12.

Now, this multiplier as shown is designed around some tightly matched transistors and one real op amp. If this was being constructed in a biCMOS IC process, the discrete matched NPNs could easily be replaced by some well-matched common-centroid NPNs with (presumably) low offset, and the op amp could be replaced by a P-channel matched pair, running as a differential amplifier. Still, the idea is the same: A simple servo, to generate VC....

Have I built this? Not yet. But that's not a big deal. Circuits like this are very easy to test out with back-of-envelope Spice. It will work perfectly.

Bob's Mailbox: Steering Lock-Ups

Hi Bob,

I have a 1991 Chevy S-10 pickup truck with automatic transmission that locks up. When one turns the key all the way toward off and does not pull the key out, the steering wheel will still lock up. I tried this while driving and learned quickly how it behaves. Just to be sure, I went out and tried it in my driveway and it indeed locks up.

Full disclosure: the test while driving was with me driving and my daughter as a passenger. I wanted to demonstrate and make the point to her that if the engine stalls, you will lose power steering. But even though the steering wheel is harder to turn, it still works. To my surprise, it locked while doing 45 mph. I quickly turned it back on before I got into real trouble.

-Jim Sylivant

That was a first-class demonstration! I think your daughter got a fine education. One of the best.

But if you only turned it 20°, until the engine died, it wouldn't lock up, right? In a stick shift, just kick it out of gear when you turn off the key, and when you turn the key back, you'll be safe.

Best regards. /rap

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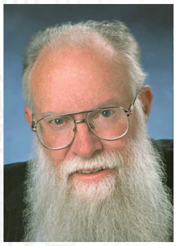


A focused
collection of
columns from
legendary
Electronic Design
author **Bob Pease**

CHAPTER 8:

What's All This Ball-On-Beam Balancing Stuff, Anyhow?

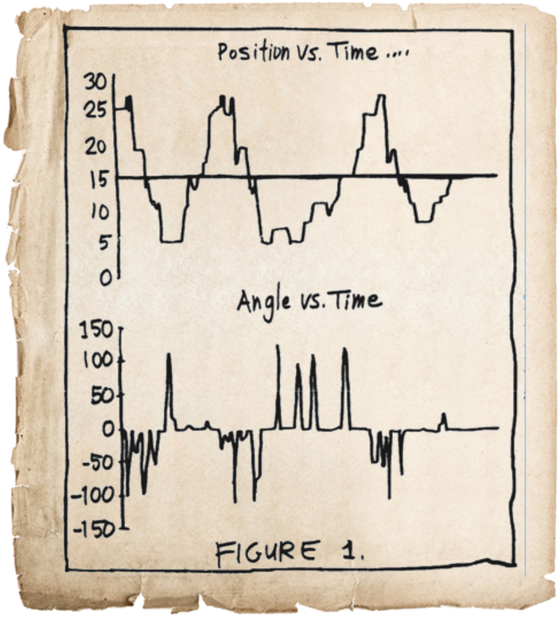
BOB PEASE, Senior Analog Engineer

Several months ago, a reader showed me an application of Fuzzy Logic, which claimed to be excellent. The IEEE Circuits and Devices Magazine (March 1994, pp. 30-35) published an article by Dr. Hua...

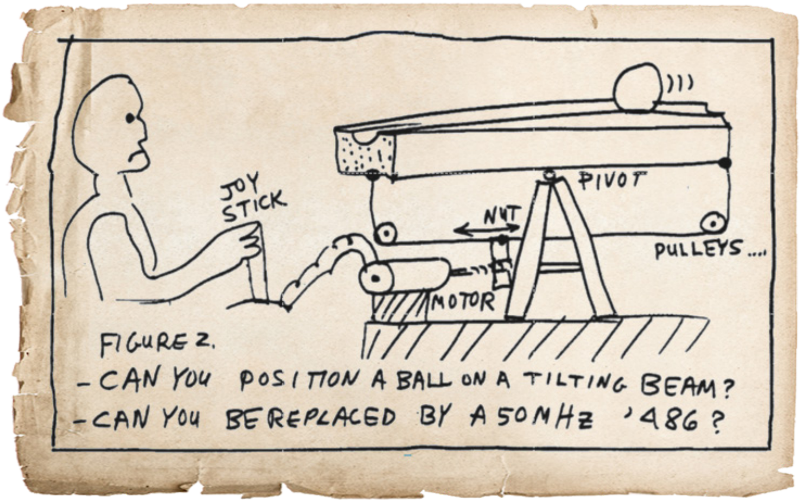
Several months ago, a reader showed me an application of Fuzzy Logic, which claimed to be excellent. The IEEE Circuits and Devices Magazine (March 1994, pp. 30-35) published an article by Dr. Hua Li and Dr. Yuandong Ji, showing how to balance a ball on a tilting beam. They showed how easy it was to use a little computer (well, actually, a 50-MHz '486-based PC) to get a ball to move to the center of a tilting beam. They claimed that Fuzzy Logic made a much quicker and smoother controller than a trained person. I present a copy of their results in **Figure 1**: The ball starts out, lurches along, crosses zero five times, and on the sixth try finally gets to stop at the center.

I asked the authors why it took six tries to stop, when any good controller would slow down and stop on the first pass. I asked them why the plot of the beam's tilt looked fishy. That's because sometimes when the beam tilted DOWN, the ball went DOWN. Other times when the beam tilted DOWN, the ball went UP. Sometimes the beam did not tilt at all, and the ball suddenly stopped. The Circuits and Devices Magazine published my questions in the Sept. 1994 issue, and also the authors' comments. Dr. Li's basic reply was to recommend that I ought to read a good book on Fuzzy Logic. And as for their results, Dr. Li could only respond, "I feel my professional practice and achievement respond louder an anything I can write".

After I researched this a little, and found almost no published information on a ball balanced on a beam, I felt challenged to make my own controller. After all, it isn't every day that I can replace a '486



1. They claimed that Fuzzy Logic made a much quicker and smoother controller than a trained person.



2. First I had to design a sensor.

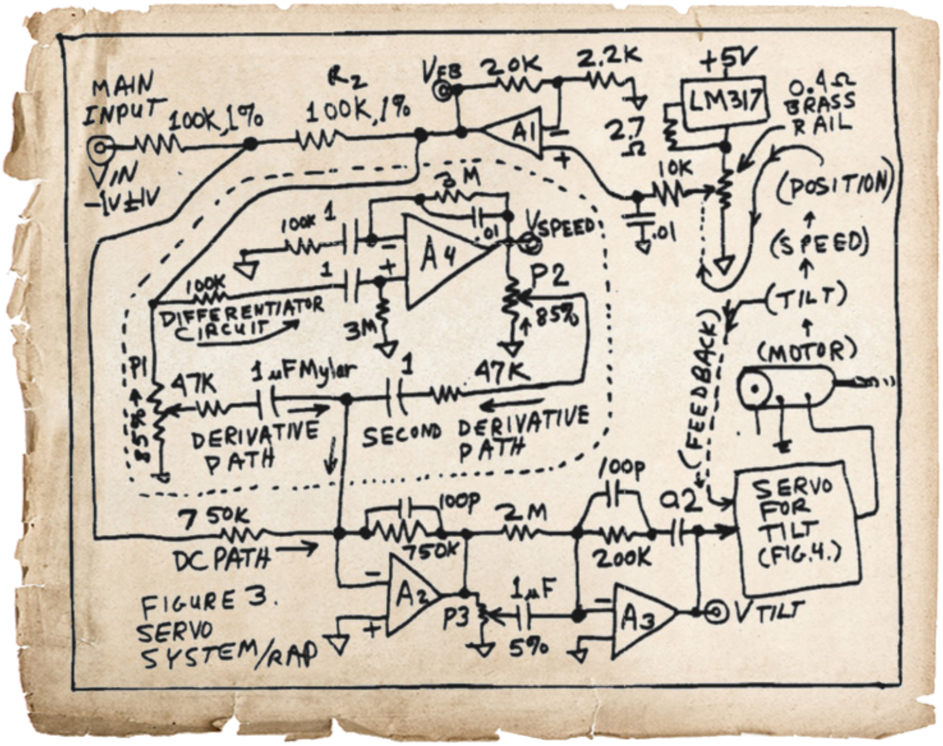
with a simple $0.85 quad op amp. So when Bob Milne asked what I plan to do for a nice "Analog" column to run in the Special Analog issue in November, I knew this was the right topic.

First I had to design a sensor. There's hardly anything simpler than V = I X R. I forced 0.50 A down a 3-ft. length of brass model-railroad rail (Zgauge) glued on top of a 1 in. x 1/4-in. X 3-ft. wooden beam (**Fig. 2**). As the metal ball rolled along on two rails, this voltage on the "hot" rail was transferred to the other rail-just like a wiper on a pot. The I X R drop was about 200 mV, so I fed that into a gain-of-plus-10 preamp to get 2 V full scale. (I mention this because Li and Ji, like most F.L. experts, do not like to mention their sensors.)

I bought several ball-bearings, 3/4-in., 7/8- in. and 1 3/32-in. diameter, at Performance Bearings on 3rd Street in San Francisco. I figured there might be some reason the larger balls would work better. Actually there was no difference. Professors Li and Ji did not disclose what they were using for a sensor, but they were apparently shining some kind of LIGHTS onto their ping-pong ball, and detecting the edge of the shadow. That seemed very awkward, as they appeared to have more than one light source. I figured my steel ball would make an excellent pot wiper. And I was right. Note, a ping-pong ball has more air-friction— it's easier to make it stop, but harder to get it going. A ball-bearing has very little friction, so it's a tough test to get it to stop.

To drive the beam and control its tilt, I used a surplus dc motor with a lead-screw. (I could have clamped a piece of threaded rod onto the shaft of any motor, but this was easy.) I drove this motor with an LM3876 power op-amp running on ±18 V to get good speed. I used dental-floss (less stretchy than nylon cord) and pulleys to couple the horizontal motion of the lead-screw to the ends of the beam. I mention this because Li and Ji, like most Fuzzy Logic experts, never talk about their power amplifier or output transducer.

The dc path for the control amplifier was easy. I took the voltage from the sensor preamp (AI) and fed it back through R2 to the main control op amp, A2. But, if I just hooked it up this way, it would oscillate like CRAZY,



3. If I just hooked it up this way, it would oscillate like CRAZY, because there are three integrators in the feedback path, not to mention the integrator action of A3.

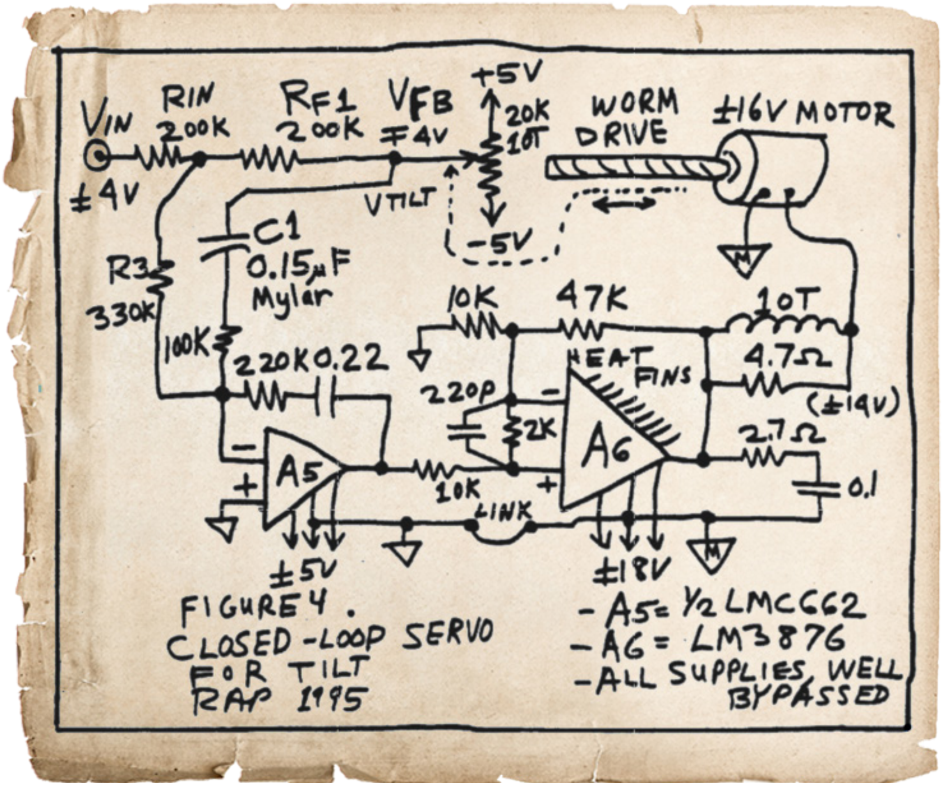
because there are three integrators in the feedback path, not to mention the integrator action of A3 (**Fig. 3**). After all, the position of the ball is the integral of its velocity—and its velocity will be the integral of the beam's tilt! The dc gain will be INFIINITE, and any roll-off of the gain with frequency will surely be at 18 dB per octave. Very bad phase shifts! This loop is sure to oscillate if I don't intervene and put in some special tricks to make this loop stable.

I knew I could not use just a simple PID (Proportional, Integral, Derivative) controller of the type I discussed in a recent column (June 26, 1995), because when there are two or three integrators in the loop, a simple PID controller can't control the loop. But my plan does look like a PID with the addition of extra circuits (differentiators).

First, I put the motor inside a feedback loop, so the motor's lag would be effectively decreased to a negligible amount (**Fig. 4**). Then we set up a pot to sense the tilt of the beam so we could close the loop. And then we built the tilt-servo. It worked the first time, but I had to make a couple changes:

- I had to add a heat sink for the power amplifier-the poor little amplifier kept going into thermal shutdown because I forgot to provide one!
- I had to decrease one capacitor's value (C1, cut 1 μF to 0.15 μF) since it was causing a long tail in settling. (If I had put a pot in there, I could have just turned the knob.)

Then we added a joystick (just a pot with a lever) so we could control the



4. I put the motor inside a feedback loop, so the motor's lag would be effectively decreased to a negligible amount.

beam's tilt manually. Making the ball move over from one spot to another spot is NOT a really easy loop to control by hand. It requires thought and planning. It's fun to try. And it's still possible to have it run smoother and faster than Li's Fuzzy Controller.

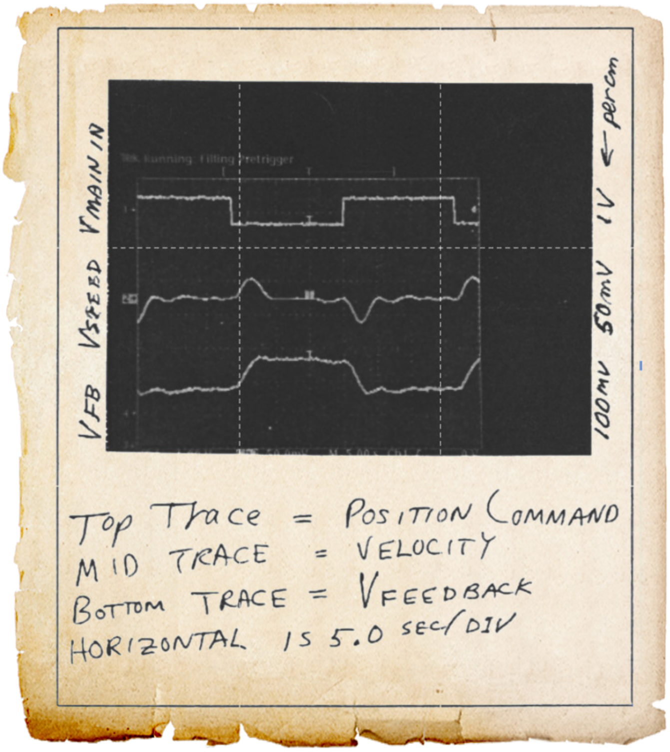
Then we built the main loop, with the dc path and a couple of differentiators added in (**Fig. 3, again**). These signals help us anticipate that the ball is approaching the right place, to help the controller slow it down and stop. It's just like the Differentiator effect in a PID loop, but we have a Derivative and a Second Derivative. You might call this a PID2 Controller.

We fired it up and it worked pretty well the first time, with most of the pots set at maximum damping. But it didn't damp out too well until I fixed two more items:

- I had to cut down some friction in the pulleys.
- One amplifier was wired wrong in the differentiator.

After that, it ran pretty well. The ball moves smoothly and only overshoots a few percent. It rarely lurches or jitters. I can use it to position the ball at any point on the beam. As shown in the scope photo (**Fig. 5**), the voltage that represents the ball position settles smoothly—just as the ball does.

I still don't know why my scheme is so slow; I want to figure out how to make it faster. There are several ideas I'd like to try out, and I've had several suggestions on making it servo and settle much faster. Maybe the integrator in the main servo



5. The ball moves smoothly and only overshoots a few percent. It rarely lurches or jitters.

loop is adding too much dc gain. I may be able to decrease friction with a dither, or a preload.

When I do get this working really well, I'll send a videotape to Dr. Li and Dr. Ji - just to show them how a beam balancer can work well. They claimed they got their best results because they didn't use any models. Mine runs well because I was able to use simple models: a rolling ball is a double integrator. They claimed their Fuzzy Logic was able to triumph over the nonlinearities of a tilted beam. Well, the approximation that sine of tilt angle = tilt is only nonlinear by a few percent. They still can't servo the ball to any place on the beam as fast as my PID2 controller.

If I get some time, I may be able to try out a deterministic solution: Bang the motor ON for a while, then when the ball has moved about half way, bang the motor to tilt the beam the other way. When the ball is nearly at its goal, then I'll servo the beam to be flat. This will require a bunch of tricky circuits, but if you're really in a hurry, this is surely the right way to do it.

P.S. Thanks for Jay Friedman's and Kevin Thompson's help.

PPS. In case the position sensor got flaky, I was prepared to buy a tube of carbon-loaded lubricant that's made by Planned Products of Santa Cruz. But we never had any trouble with that, so there was no need to buy that stuff. It's a little pricey, about $13 at Fry's, but it's nice to know it's there.

Originally published in Electronic Design, November 20, 1995

RAP's 1997 comments: My BOBB keeps on plugging. When I published this in '95, 1 had gotten the settling time down to 8 seconds. I added an R-C network from -Vin to the first summing point, and it didn't help. I added it again later, and it worked just fine, and helped me get the settling time down to 4 seconds, and 3.2 on a good day. How am I going to get it down to less than 2 seconds? More on this, later. I have seen a couple recent technical reports on BOBB using Fuzzy Logic. They are not bad at all. They seem rational. They are not hyped. And, surprisingly, they make no reference to the seminal Ji/Li article. More on this later.

Here I'll list three References from the magazine: IEEE Transactions on Fuzzy Systems. (Isn't it funny that dozens of F.L. enthusiasts gather there to say good things about F.L., and I'm the only protection you have from them?)

1. *"Designing Fuzzy Controllers from a Variable Structures Standpoint,"* J. Glower and J. Munighan, North Dakota State University, Fargo, N. Dak. pp. 138-144, Feb. 1997. They simulated a response of about 4 seconds, but had not built a model at the time of publication. Honest, realistic work. I'll add more comments later. They agree that adding P plus D terms before they are sent into the F.L. controller can improve and simplify the system, as I proposed above in my PID column.
2. *"How to Design a Discrete Supervisory Controller for Real-Time Fuzzy Control Systems,"* N. Muskinja et al., University of Maribor, Slovenia, May 1997, pages 161- 166. They show some curves with some not-entirely linear results, with settling in the range 4 to 8 seconds. They don't explain a lot about their analog-digital interface. But they use pulleys like I do.
3. *"Adaptive Fuzzy Control: Experimental and Comparative Analyses."* R. Ordonez et al., Ohio State University, May 1997, pp. 167-188. They show comparisons between computer simulations and the systems they BUILT. Very honest guys. They actually wrote about their sensors and their interfaces. Very honest. They got results in the 4 to 6 second range. More on this, later.-rap

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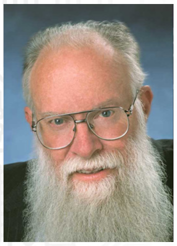


CHAPTER 9:

What's All This "Error Budget" Stuff, Anyhow?

BOB PEASE, Senior Analog Engineer

I love to recommend amplifiers with high CMRR but depending on cheap 1% resistors can hurt your "error budget" a lot more than you'd suspect.

I was just on the phone explaining how to do an "error budget" analysis on some fairly simple circuits to a young engineer. Later, I mentioned this while I was visiting with my friend Martin, and he said he had been quite surprised when he found that many engineers in Europe were quite unfamiliar with the concept of an "error budget." How can you design a good circuit without being aware of which components will hurt your accuracy?

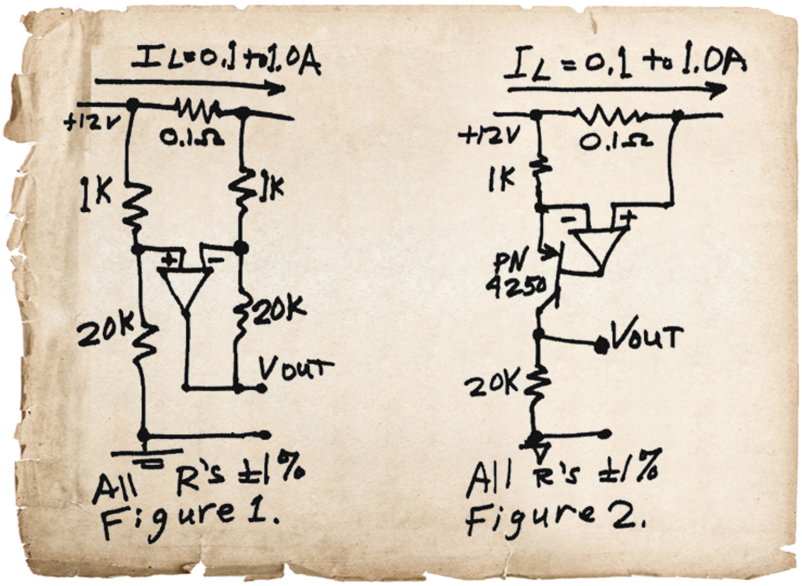
When I was a kid engineer back in 1962, my boss George Philbrick gave me a book on differential amplifiers by Dr. R. David Middlebrook, and he asked me to do a book review. I studied the book, and it was full of hundreds of partial differential equations. If you wanted the output of a circuit with 14 components, you could see a complete analysis of how each component would affect the output offset and gain. Each equation filled up a whole page. It did this several times.

Yet it didn't offer any insights into what's important. I mean, is b × d(R1) more important than R1 × d(b)? In retrospect, I'm glad I didn't submit any critique of that book. I woulda done more harm than good. Such a mess! Even now, it would be hard to write a critique on a book that was so true, but so unhelpful.

Things are much simpler now that people are mostly (but not entirely) designing with op amps. The best thing is that the output offset and dc gain and ac gain errors are largely orthogonal. An "operational" amplifier does perform, largely, an " operation" based on what task you ask it to perform when you "program it" with Rs and Cs. If the offset varies, the gain does not, and vice versa. We all agree that it's very helpful that you can compute what the performance will be with almost no interaction. No partial derivatives.

Now, let's take a look at a couple of applications—real circuits—and their tolerances within an error budget. Here is an amplifier to magnify the I × R drop of current through a 0.1-Ω resistor and bring it back down to ground. **Figure 1** shows a conventional differential amplifier, with the common mode up at +12 V. The gain of −20 will bring the 1.0 A × 0.1Ω signal down to a ground level. If the current is 0.1 A, the output will be 0.2 V, "small-scale." A full-scale current of 1 A will bring the output up to 2.0 V, which is suitable to send to a detector or analog-to-digital converter.

Let's select an op amp like the LMC6482B, with low offset voltage less than 1.0 mV. (There are other versions of this amplifier with less than 0.35 mV, but let's select an intermediate model.) This 1 mV does cause 21 mV of output error. This op amp has less than 20 pA of I_B at all temperatures, so at least that's negligible.



(Bipolar op amps might have small I_B errors, but you'd have to check it.)

Now let's see what the resistors add. Assuming all Rs have a 1% tolerance, the gain of (2.0 V per A) has a tolerance of ±3%. This would cause ±60 mV at full scale, but only ±6 mV at "small scale" (0.1 A). This may be acceptable.

Then let's consider the common-mode errors. If R4 has a 1% tolerance, and it has 11.4 V across it, the 1% tolerance could cause a 114-mV error. By symmetry, a 1% error of each of R1, R2, R3 can cause another 114 mV! Added together, the common mode could cause an output error of 456 mV! That's about ±1/4 of full scale—even for small signals. That doesn't look so good to me!

It's true that if adjacent 1-kΩ resistors are inserted, they're likely to match within ±1/2% so the probable error between the pair might cause ±60 mV, and the ±1/2% matching between the 20 kΩ would cause another 60 mV. That added to the 21 mV from the VOS would add to 141 mV.

Some textbooks teach you that you should add these errors arithmetically to 141 mV. Others point out that they could be added in an RMS way, so that 60 + 60 + 21 mV = 87 mV. Typically, this might be true. But the worst case of 141 or 456 mV might be more realistic. I mean, if you're going to build 1000 circuits, and most of them are better than 141 mV, what are you going to do with the 400 circuits that are worse than 141 mV? And, that's still 7% of full-scale....

You could go shopping for 0.1% resistors, but they aren't cheap. You could put in a trim-pot to trim the error to (no offset error) for small signals. But as you may have noticed, a trim-pot has to be properly trimmed. And if that pot is accessible, it could someday be mistrimmed, and it would have to be corrected, in some awkward calibration cycle. Most people want to avoid that trim-pot. Before we decide that this 141 mV is unacceptable, let's look at another circuit.

Figure 2 shows an alternative circuit with the same gain, 2.0 V per A, using a PN4250 or 2N4250, a high-beta pnp transistor. What does the error budget look like? The same op amp causes just 20 mV of output error. The 1% resistor tolerances cause the same gain error, 60 mV at full scale, or 6 mV at "small scale."

The newly added transistor adds (?1/3%) max from its alpha, or less than 7 mV, at full scale.

What is the offset error due to common-mode rejection ratio (CMRR), or due to resistor mismatch? Nothing. Zero. The transistor doesn't care about the voltage across it. There are no resistors with 12 V across them.

So the offset error is ±20 mV, due primarily to the amplifier's VOS (which could be reduced), not ±400 mV. This little circuit has greatly reduced errors compared to Figure 1, even if Figure 1 had a couple bucks of 0.1% resistors. This may be acceptable. Even the offset errors could be reduced to 7 mV by selecting the LMC6482A.

So we have seen that circuits with similar functions can have completely different "error budgets." I love to recommend amplifiers with high CMRR. But depending on cheap 1% resistors can hurt your "error budget" a lot more than you'd suspect.

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