Adjustable Cable Equalizer Combines Wideband Differential Receiver with Analog Switches

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Originally intended to carry LAN traffic, *category-5* (Cat-5) *unshielded twisted-pair* (UTP) cable has become an economical solution in many other signal-transmission applications, owing to its respectable performance and low cost. For instance, an application that has become popular is *keyboard-video-mouse* (KVM) networking, in which three of the four twisted pairs carry the *red, green, and blue* (RGB) video signals.

Like any transmission medium, Cat-5 imposes transmission losses on the signals it carries, manifested as signal dispersion and loss of high-frequency content. Unless something is done to compensate for these losses, they can render the cables useless for transmitting high-resolution video signals over reasonable distances. Presented here is a practical technique to compensate for Cat-5 losses by introducing an *equalizer* (EQ), with eleven (11) switchable cable-range settings, at the receiving end of the cable. Because each setting of the EQ provides the proper amount of frequency-dependent gain to make up for the cable losses, the EQ-cable combination becomes suitable for high-resolution video transmission.

The first step in the EQ design is to derive a model for the Cat-5 frequency response. It is well known that the frequency response of metallic cable follows a low-pass characteristic, with an exponential roll-off that depends on the square root of frequency. Figure 1 depicts this relationship for lengths of Cat-5 from 100 feet (30.48 m) through 1000 feet (304.8 m), in 100-foot increments. In this illustration, it should be evident that the *power* loss at a given frequency is characterized by a constant attenuation rate (expressed in dB/ft).

Table I presents the Cat-5 equivalent *voltage*-attenuation magnitudes as a function of frequency for the same cable lengths as shown in Figure 1.

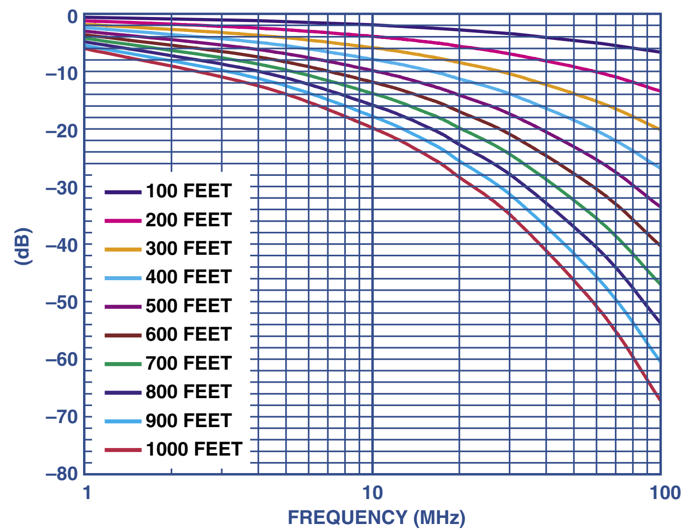


Figure 1. Frequency responses for various lengths of Cat-5 cable.

Using the data in Table I, the frequency response for each cable length can be approximated by a mathematical model based on a negative-real-axis pole-zero transfer function. Any one of the many available math software packages with the capability of least-squares polynomial curve fitting can be used to perform the approximation. Figure 1 suggests that, for long cables at high frequencies—because of the steepening slope, exceeding 20 dB/decade—consecutive negative-real-axis poles are required to obtain a close fit, while at low frequencies—to fit the nearly linear slope—alternating poles and zeros are required. As an extreme example, the frequency response for 1000 feet of cable at 100 MHz is rolling off approximately as $1/f^4$, which can only be attained by a model having multiple consecutive poles.

Equalization is achieved by passing the signal received over the cable through an equalizer whose transfer function is the reciprocal of the cable pole-zero model's transfer function. To neutralize the cable's frequency-dependency, the EQ has poles that are coincident with the zeros of the cable model and zeros that are coincident with the poles of the cable model.

Table I. Voltage-attenuation magnitude ratios of Cat-5 cable. For example, 500 feet of cable attenuates a 10-MHz, 1-V signal to 0.32 V, which corresponds to about –9.90 dB (Figure 1).

Frequency	100 ft	200 ft	300 ft	400 ft	500 ft	600 ft	700 ft	800 ft	900 ft	1000 ft
1 MHz	0.932	0.869	0.8100	0.7550	0.7040	0.65600	0.6120	0.57000	0.53200	0.496000
4 MHz	0.866	0.750	0.6490	0.5620	0.4870	0.42200	0.3650	0.31600	0.27400	0.237000
10 MHz	0.796	0.634	0.5040	0.4020	0.3200	0.25400	0.2030	0.16100	0.12800	0.102000
16 MHz	0.750	0.562	0.4220	0.3160	0.2370	0.17800	0.1330	0.10000	0.07500	0.056300
20 MHz	0.722	0.521	0.3760	0.2710	0.1960	0.14100	0.1020	0.07350	0.05300	0.038300
31 MHz	0.663	0.440	0.2920	0.1940	0.1280	0.08510	0.0565	0.03750	0.02480	0.016500
63 MHz	0.551	0.303	0.1670	0.0920	0.0507	0.02790	0.0154	0.00846	0.00466	0.002570
100 MHz	0.462	0.214	0.0987	0.0456	0.0211	0.00973	0.0045	0.00208	0.00096	0.000444

One of the properties of passive RC networks is that the alternating poles and zeros of their driving-point impedances are restricted to the negative real axis. This property also holds for those operational-amplifier circuits having a transfer function determined by the simple ratio of feedback-impedance to gain-impedance (Z_f/Z_g), where these impedances are RC networks. (The property does not hold for other cases, such as *active* RC filter sections that synthesize conjugate pole-pairs.)

For a practical equalizer design, we prefer that an EQ be based on a single amplifier stage in order to keep its adjustability manageable and to minimize cost and complexity. The equalizer to be discussed here uses RC networks of the former type, described by Budak, with alternating poles and zeros; but such a design precludes the use of a single amplifier stage to realize the consecutive zeros required to compensate for consecutive poles in the cable model at all frequencies. As a compromise that will provide good equalization for all but long cables at high frequencies, the design chosen uses a single amplifier to realize two zeros and two poles, alternating on the negative real axis.

Because equalization requires increased gain at the high end of the band, a low-noise amplifier is required. To avoid introducing significant errors due to amplifier dynamics, a large gain-bandwidth product is needed. For the specific design requirements of this application, the amplifier must have the capacity to perform frequency-dependent differential-to-single-ended transformations with voltage gain. The Analog Devices AD8129, just such an amplifier, is the heart of the basic frequency-dependent gain stage in the EQ. Figure 2 shows the dual-differential-input architecture of the AD8129, and its standard closed-loop configuration for applications requiring voltage gain.

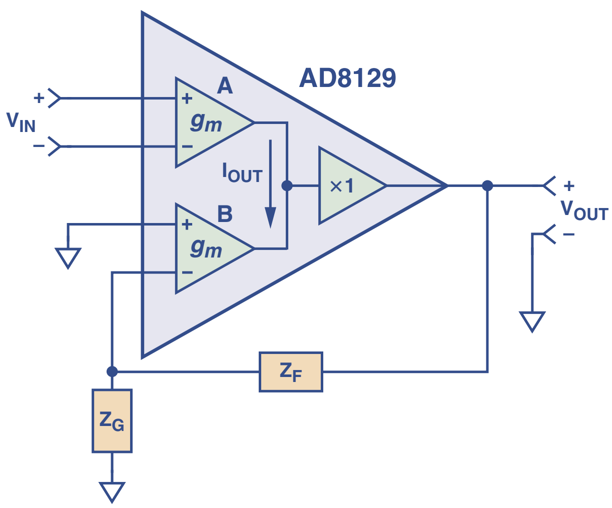


Figure 2. The AD8129 in a standard closed-loop gain configuration.

As can be seen, the AD8129 circuitry and operation differ from those of the traditional op amp; principally, it provides the designer with a beneficial separation of circuitry between the differential input and the feedback network. The two input stages are high-impedance, high-common-mode-rejection (CMR), wideband, high-gain transconductance amplifiers with closely matched g_m. The output currents of the two transconductance amplifiers are summed (at high impedance), and the voltage at the summing node is buffered to provide a low-impedance output. The output current of amplifier A equals the negative of the output current of amplifier B, and their transconductances are closely matched, so negative feedback applied around amplifier B drives v_{out} to the level that forces the input voltage of amplifier B to equal the negative of the input voltage of amplifier A. From the above discussion, the closed-loop voltage gain for the ideal case can be expressed as:

$$\frac{v_{out}}{v_{in}} = 1 + \frac{Z_f}{Z_g} \equiv A_V \qquad (1)$$

The EQ is designed using this gain equation, with RC networks for Z_f and Z_g. Its canonical circuit is depicted in Figure 3, which represents an EQ designed to compensate for a given length of cable.

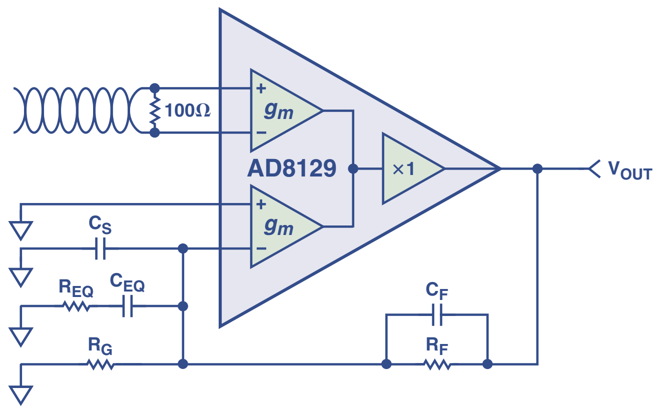


Figure 3. Canonical circuit of the equalizer.

In Figure 3, the high-differential input impedance of the upper transconductance amplifier facilitates provision of a good impedance match for the signal to be received over the Cat-5 cable; the lower amplifier provides the negative feedback circuit that implements the frequency-dependent gain. The Bode plot for the circuit has a high-pass characteristic, as shown in Figure 4. Z_n and P_n are the respective zeros and poles of the equalizer.

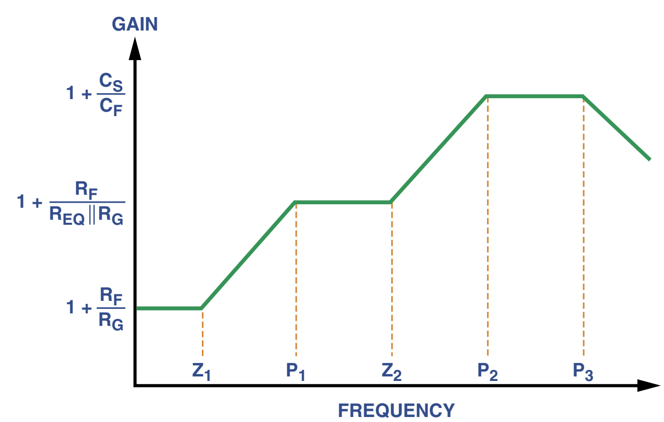


Figure 4. Bode plot of the canonical equalizer circuit.

In the following analysis, where the pole-zero pairs in Figure 4 are sufficiently separated, the capacitors can be approximated as short- or open circuits. The pole- and zero frequencies are expressed in radians per second. At low frequencies, all capacitors are open circuits, and the gain is simply

$$1 + \frac{R_f}{R_g}$$

This gain, set to compensate for flat (i.e., dc) losses, includes any loss due to matching and the cable's low-frequency flat loss. It also provides the flat gain required to stabilize the AD8129 when equalizing short cables (to be covered in greater depth below).

Moving up in frequency, the lowest-frequency pole-zero EQ section, comprising the series-connected R_{EQ} and C_{EQ}, starts to take effect, producing Z_1 and P_1. By approximating C_f and C_S as open circuits, the following equations can be written:

$$Z_1 = \frac{1}{\left(R_g + R_{EQ}\right) C_{EQ}} \tag{2}$$

$$P_1 = \frac{1}{R_{EQ}\, C_{EQ}} \tag{3}$$

The magnitude of the frequency response asymptotically approaches

$$1 + \frac{R_f}{R_{EQ} \parallel R_g}$$

as C_{EQ} approaches a short circuit.

As frequency increases, C_S begins to take effect, introducing another zero, Z_2. The primary function of C_f is to keep the amplifier stable by compensating for C_S. By initially approximating C_f as an open circuit ($C_f << C_S$), and presuming that Z_2 is sufficiently far in frequency from P_1 that C_{EQ} can be considered as having negligible impedance compared to R_{EQ}, the approximate expression for Z_2 can be written:

$$Z_2 = \frac{1}{\left(R_f \parallel R_{EQ} \parallel R_g\right)\left(C_S + C_f\right)} \approx \frac{1}{\left(R_f \parallel R_{EQ} \parallel R_g\right) C_S} \tag{4}$$

Finally, P_2 can be expressed as:

$$P_2 = \frac{1}{R_f\, C_f} \tag{5}$$

Between P_2 and P_3, the magnitude of the frequency response asymptotically approaches the closed-loop gain produced by the capacitive divider formed by C_f and C_S,

$$1 + \frac{C_S}{C_f}$$

This is the closed-loop gain at frequencies leading up to P_3, so P_3, which is due to the amplifier's dominant-pole roll-off, can be approximated as:

$$P_3 = \left[1 + A_O\left(\frac{C_f}{C_f + C_S}\right)\right]\omega_p \tag{6}$$

where A_O is the amplifier's dc open-loop gain, and ω_p is the amplifier's dominant pole. This result follows directly from standard op-amp gain-bandwidth analysis. P_3 is imposed by the gain-bandwidth product of the amplifier, and sets the approximate upper frequency limit of the equalizer. Using the above results, along with the cable's pole-zero model, an EQ can be designed for any practical length of cable that can be modeled by two alternating pole-zero pairs, provided that the amplifier has a sufficiently high-gain-bandwidth product.

In order for the EQ to be useful over a wide range of cable lengths, it must be adjustable. A simple means of adding adjustability is to switch different RC networks between the feedback pin of the AD8129 and ground. This scheme is illustrated in Figure 5.

Each EQ section in Figure 5 is appropriate for a range of cable lengths. Section EQ0 covers 0 to 50 feet, and Section EQ10 covers 950 to 1000 feet. The other sections are centered on 100 feet, 200 feet, etc., and cover ±50 feet from their centers. This resolution is sufficient for most RGB applications.

Practical Matters

The AD8129 is stable for gains greater than 10 V/V, where it has a nominal phase margin of 56°, but if care is taken with regard to layout and parasitic capacitance, it can be successfully operated with a gain of 8, where it has approximately 45° of phase margin.

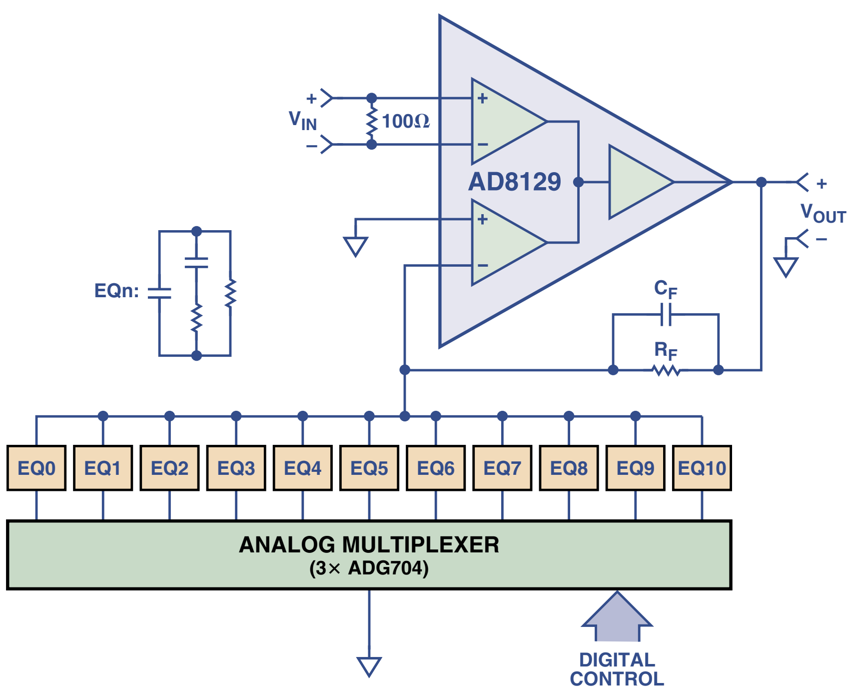


Figure 5. Equalizer with switchable sections.

This gain is required at high frequencies. For the longer cable lengths, sufficient high-frequency gain is provided by the high-pass nature of the equalizer. For cable lengths between 0 and 300 feet, however, excess flat gain is required in order to keep the AD8129 stable. Because the excess gain is flat, it can be easily inserted by adjusting the R_f/R_g ratio, and removed by switching in the same amount of flat attenuation after the equalizer.

The AD8129's input stage has a limited linear dynamic range (±0.5-V operating range). For optimum performance, it is best to attenuate the 700-mV RGB video signals by a factor of *four* before applying them to the AD8129 inputs. Sometimes the video signals are already attenuated by a factor of two before transmission over the cable. (This is not the matching loss—which is normally accounted for by using a cable *driver* with a gain of 2.) In this case, an additional factor-of-two attenuation can be inserted at the input to the AD8129 to produce an end-to-end flat attenuation factor of four. A buffer with a flat gain of four, placed after the EQ, is used to compensate for this attenuation (the AD8001 is an excellent choice for this stage). The buffer also simplifies the switched attenuator at the EQ output, which can be a simple L-pad.

The parasitic capacitance of each *off* channel in the ADG704 analog multiplexers used to select the EQ sections is 9 pF. The parasitic capacitance of the sum all of the unselected EQ sections is therefore quite large; it adds to the C_S value of the selected EQ section. For the EQ sections from 400 to 1000 feet, this parasitic capacitance can usually be absorbed into C_S. For the shorter sections, the excess closed-loop gain described above is used to compensate for the peaking caused by the parasitic capacitance. As a general rule, it is best to scale the impedances used in the EQ sections in such a way as to maximize the capacitance values, thus allowing absorption of as much parasitic capacitance into C_S as possible. This can't be carried too far however, since it reduces the associated resistances. The scaling is also limited by the parasitic inductance in the traces that connect the EQ sections. Small resistances provide little damping; if the resistance levels are too small, a moderate-Q tank circuit, resulting from the parasitic trace inductances and switch capacitances, can cause instability in the AD8129.

Optimizing the EQ PCB layout is of paramount importance. The major part of all power- and ground-plane copper must be removed from all layers under the traces that connect to the AD8129 summing node. Small ground-plane strips can be strategically placed as needed in these areas to provide low-Z return current paths, while minimizing stray capacitance at the summing node. The AD8129s and ADG704s should be in μSOIC packages, and the AD8001 should be in the SOT-5 package. Trace inductance in the EQ sections must be kept to an absolute minimum to avoid instability in the AD8129, so 0402 packages should be used for the resistors and capacitors, and the EQ sections should be laid out in such a way as to minimize trace lengths.

After the RC values that are based on the cable model have been determined, and the parasitic effects have been taken into account, a final tuning process in the time domain is required for RGB video applications. This is because one of the most important performance metrics for RGB video circuits is the step response; the *step response* of the cable and EQ combination must be tuned so as to exhibit fast rise time, minimum overshoot and ringing, and short settling time. C_S has the greatest effect on overshoot and ringing, and the series connection of R_{EQ} and C_{EQ} has the greatest effect on the long-term settling time. The positions of the pole and the zero produced by the series connection of R_{EQ} and C_{EQ} can be altered somewhat without changing the frequency response a great deal, because they are placed where the cable's frequency response has a rather gradual roll-off. This means that the equalized frequency response can appear to be quite good, while the positions of the pole and zero can be suboptimal from a step-response standpoint. It is therefore best to fine-tune the values of C_S, R_{EQ}, and C_{EQ} in the time domain by adjusting their values to produce a step response with the shortest settling time.

Since the equalizer must interface with long differential cables with no ground reference, the received signal may contain large common-mode voltage swings with respect to the power supply voltages at the receiver. It is therefore best to use dual power supplies of at least ±5 V. This also allows the output signal to swing to 0 V, which is generally required for video signals.

CONCLUSION

The equalizer presented here can stably compensate for lengths of Cat-5 cable from 0 to 1000 feet at frequencies to greater than 100 MHz at short cable lengths and to 25 MHz at 1000 feet, making it suitable for KVM networking and other high-resolution video transmission applications. ▶

REFERENCE

Budak, *Passive and Active Network Analysis and Synthesis*, Houghton Mifflin, 1974.

This article can be found at http://www.analog.com/library/analogDialogue/archives/38-07/equalizer.html, with a link to a PDF.

A READER NOTES—REACTANT FLOW SENSOR

Jason Dugas, an electrical engineer at NASA's Johnson Space Center recently contacted the editors about an article published in *Analog Dialogue* in 1971. The article, entitled *Measuring Air Flow Using a Self-Balancing Bridge* (see below), details how an operational amplifier can be applied in the design of hot wire anemometers used to monitor flow rates. An excerpt of Jason's letter follows.

> The article was of interest because I wanted to learn more about the history of a similarly-designed flow meter that uses this architecture and is still flying on the space shuttle fleet today! The flow meters are used to monitor oxidizer and fuel flow rates at the inlets of the shuttle orbiter's fuel cell stacks.
>
> Analyzing the flow meter has been a lesson in history, as much as in electrical engineering, as the circuit for which I'm responsible was designed in 1971 (I wasn't born until 1978). The reactant flow sensors use the old "Royer oscillator" for the internal power supply. Both +24 and –12-V dc supplies are required for the circuit's operation; the internal supply produces these voltages from a 28-V bus. The power supply consists of two sections: (1) a dc-dc converter that takes the 28 V from the shuttle's power bus and produces regulated +20 V, and (2) a self-oscillating power oscillator that converts the +20 V into a square wave, which is then coupled via two transformer windings to bridge rectifier circuits that produce the isolated +24-V and –12-V supply voltages.
>
> The dc-dc converter is based upon a LM105 voltage regulator. The LM105—developed in the late 1960's and now obsoleted by newer designs/topologies—can be used as both a linear and a switching regulator. Refer to National Semiconductor application notes AN-1, AN-2, AN-8, and AN-23 (if you can find them by now—Ha!).
>
> Once the output of the dc/dc converter has ramped up, the oscillator starts to function. Start-up of the oscillator depends on a mismatch between two BJTs, which results in one of them reaching saturation first, while the other one is driven to cutoff.
>
> The other main functional area of the RFS is the circuitry that interfaces with the temperature sensors. The flow sensor works on thermal convection. The basic topology is a hot wire anemometer. Hot wire anemometers determine flow by measuring the change in temperature of an element due to convective heat loss. Traditionally, the sensor operates from a constant current source. This increases sensitivity but presents a problem when there is a wide range of reactant flow. At low flow rates, the element will overheat and potentially fail. The RFS avoids this problem by using a slightly different approach, called a self-balancing bridge. In this design, the temperature of the sensor is held constant by varying the excitation voltage of the bridge. The flow rate is inferred from the change in excitation voltage.
>
> In order to become smarter and familiar with the older architectures, I consulted the now historical literature, including early volumes of *Analog Dialogue* and the *Switching and Linear Power Supply, Power Converter Design* by Pressman, written in 1977—one year before my birth.

APPLICATION BRIEF

Reprinted from *Analog Dialogue* Volume 5, Number 1, January 1971

Measuring Air Flow Using a Self-Balancing Bridge

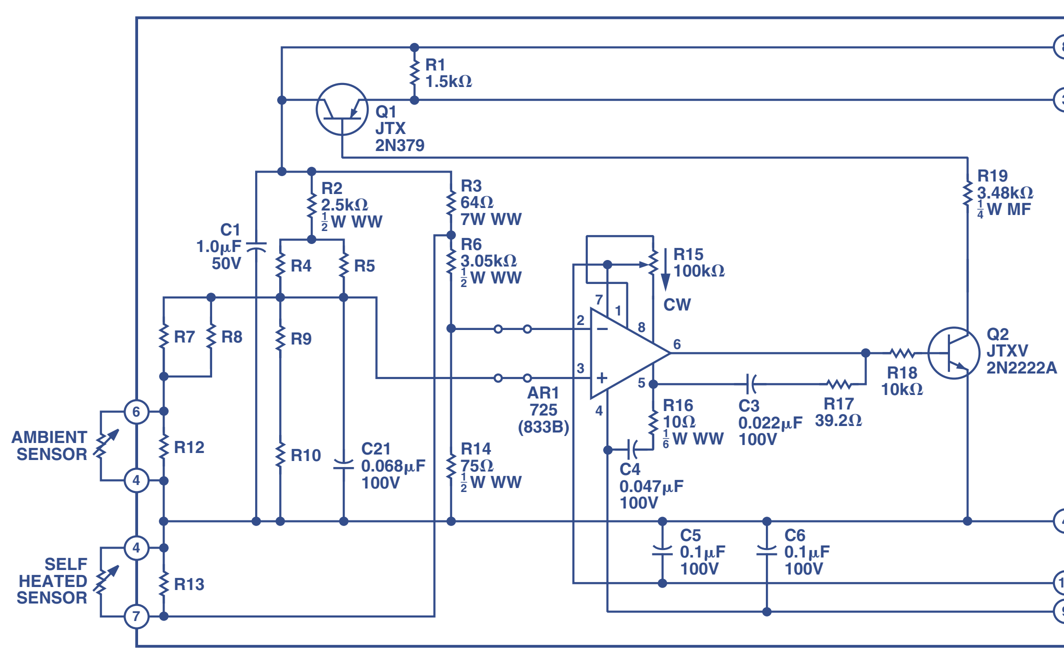
The design of a hot wire anemometer presents an interesting application for operational amplifiers.

The purpose of the instrument is to measure air speed by its cooling effect upon an electrically heated platinum filament that exhibits a high positive temperature coefficient of resistivity.

The filament characteristic is shown in Figure 1, for two values of air speed. Two classical ways of operating it would be at constant voltage or constant current.

With constant current applied (say, 0.6 amperes), the sensitivity is reasonably good—about 0.4V change for $\Delta s = s_2 - s_1$.

However, there is every prospect that in still air the filament has the potential of burning itself up (resistance increases as temperature increases, due to lack of air flow, which causes the voltage to increase [constant current], increasing the dissipation, the temperature, and again the resistance, etc.).



This bridge amplifier, used to measure oxidizer- and fuel flow rates on the space shuttle, was derived from the circuit described in a 1971 *Analog Dialogue* article.

With constant voltage applied, the increase of resistance with temperature causes operation to be quite safe, but also relatively insensitive, especially at low air speeds.

Another factor that makes both constant-current and constant-voltage operation unsatisfactory is the necessity for the temperature of the filament to change to detect a change in air speed; this necessarily causes delay. If the air speed indicator is part of a control loop, the measurement delay could cause slow response or instability of the control loop.

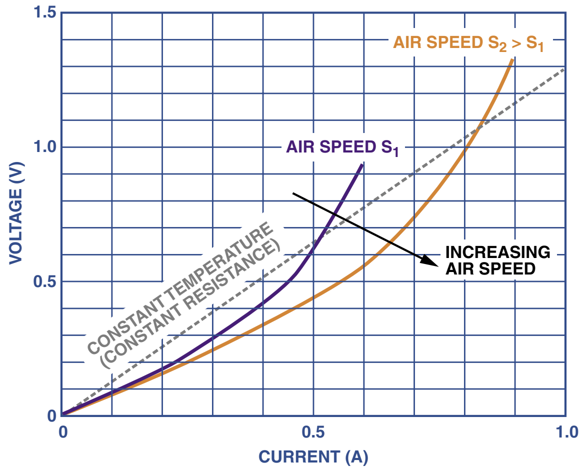


Figure 1. Volt-ampere characteristics of 7/16"L × 0.002"D straight filament of pure platinum in the presence of moving air, for two values of air speed.

An approach that answers all of these objections satisfactorily is to operate the filament as though it had constant resistance (i.e., constant temperature).

If, for example, the resistance were maintained at 1.3 Ω, as indicated by the dashed line, one could obtain a current change of 0.3 A and a voltage change of 0.4 V, with no danger of overheating in normal operation. Response would be quite speedy, since temperature changes are momentary and small.

The basic circuit for achieving constant temperature operation is the feedback circuit of Figure 2, consisting of a bridge, an op amp, and a power amplifier. The operational amplifier continuously adjusts the flow of current (through the power transistor) to maintain its two inputs equal. This can be done only by keeping the voltage across the filament equal to that across R2, and the filament current equal to the current through R1. However, since the current through R1 is proportional to the current through R0, (which has the same voltage drop as R1) and the current through R0 is determined by the voltage drop across R2, it can be seen that the resistance of the filament, RF, must be equal to that of R2, multiplied by the ratio of R_1 to R_0.

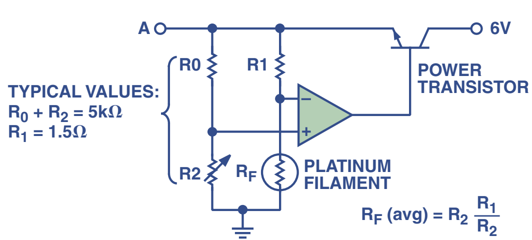


Figure 2. Basic circuit of the temperature-controlled bridge.

THE AUTHOR

In 1971, Ing. José Miyara was a partner in the firm TELEGUARD, in Rosario, Argentina. Their principal activities are in Industrial Electronics and Automatic Control.

Suppose now, that, starting from a given operating equilibrium point, the air flow increases. This will take heat away from R_F, causing its voltage to tend to drop. The amplifier's output voltage increases, which increases the current through the power transistor, and thus makes more power available for the filament to dissipate to maintain its temperature (and hence its resistance) constant.

The output voltage is measured at terminal "A", which provides an amplified version of the filament voltage, at an impedance level low enough to operate even the crudest of meter movements.

The zero-air-speed voltage is backed off by means of an auxiliary constant voltage, and the readings can be displayed with a moving coil meter. The scale is a nonlinear function of air speed, actually expanding toward the lowest values. LOW air speeds can be read with high sensitivity; in fact, the device can virtually detect a whisper several feet away.

CIRCUIT NOTES

For practical realization, the following points must be considered:

1. A voltage offset must be deliberately introduced into the operational amplifier (or elsewhere) to insure that the output goes positive with zero differential input; otherwise, the circuit might remain dead when turned on.
2. The power transistor must have ample current-handling capacity; the filament requires several hundred mA
3. Depending on the physical layout, especially when the filament is away at the end of a twisted pair, wild high-frequency oscillations are possible. Though not visible with low-frequency readout devices (however, rectification can cause voltage offsets), they look nasty on an oscilloscope screen. A 0.1 µF capacitor between the base and collector of the power transistor can often serve to stabilize these oscillations. A small resistor in series with the base may also be helpful.
4. The filament is a physical device with thermal lag. Although the circuit is fast enough to prevent loop oscillations when used in a larger control loop, it is itself a process control loop and may require the usual compensation techniques to maintain its own internal stability.
5. R0 and R2 form a trim potentiometer to set the operating temperature (e.g. resistance) of the filament. If R2 is a variable resistance, you start with $R_2 = 0$, and increase it until the filament just starts to glow, then back down a little. This will give optimal sensitivity.

APPLICATIONS

The device has been used in a commercially-produced apparatus to trip out equipment when the air speed in a forced draft duct falls below a preset value, but the approach is suggestive of a number of other applications in instrumentation.

In gas chromatographs, it could be used to monitor the minute gas flows required, and also to assure optimum sensitivity from thermal-conductivity filaments, while preventing their burnout in faulty operation.

Another application could be for constant-temperature ovens for crystals, differential pairs, etc., using copper or a thermally sensitive alloy such as "Balco" for the combined heater/temperature-sensor function. In such an arrangement, the control- led temperature could be slaved to another (arbitrarily) variable temperature (e.g., to insure a constant temperature difference) if R2 were a platinum temperature bulb with the circuit so dimensioned as to avoid causing it to introduce errors due to its own self-heating.