SECTION 7-1

Passive Components

James Bryant, Walt Jung, Walt Kester

Introduction

When designing with op amps and other precision analog devices, it is critical that users avoid the pitfall of poor passive component choice. In fact, the wrong passive component can derail even the best op amp or data converter application. This section includes discussion of some basic traps of choosing passive components for op amp applications.

So, good money has been spent for a precision op amp or data converter, only to find that, when plugged into the board, the device doesn't meet spec. Perhaps the circuit suffers from drift, poor frequency response, and oscillations—or simply doesn't achieve expected accuracy. Well, before blaming the device, closely examine the passive components—including capacitors, resistors, potentiometers and, yes, even the printed circuit boards. In these areas, subtle effects of tolerance, temperature, parasitics, aging, and user assembly procedures can unwittingly sink a circuit. All too often these effects go unspecified (or underspecified) by passive component manufacturers.

In general, if using data converters having 12 bits or more of resolution, or op amps that cost more than a few dollars, pay very close attention to passive components. Consider the case of a 12-bit DAC, where ½ LSB corresponds to 0.012% of full scale, or only 122 ppm. A host of passive component phenomena can accumulate errors far exceeding this. But, buying the most expensive passive components won't necessarily solve the problems either. Often, a *correct* 25-cent capacitor yields a better-performing, more cost-effective design than a premium-grade part. With a few basics, understanding and analyzing passive components may prove rewarding, albeit not easy.

Capacitors

Most designers are generally familiar with the range of capacitors available. But the mechanisms by which both static and dynamic errors can occur in precision circuit designs using capacitors are sometimes easy to forget, because of the tremendous variety of types available. These include dielectrics of glass, aluminum foil, solid tantalum and tantalum foil, silver mica, ceramic, Teflon, and the film capacitors, including polyester, polycarbonate, polystyrene, and polypropylene types. In addition to the traditional leaded packages, many of these are now also offered in surface-mount styles.

Figure 7-1 is a workable model of a nonideal capacitor. The nominal capacitance, C, is shunted by a resistance R_P, which represents *insulation resistance* or leakage. A second resistance, R_S—*equivalent series resistance*, or ESR—appears in series with the capacitor and represents the resistance of the capacitor leads and plates.

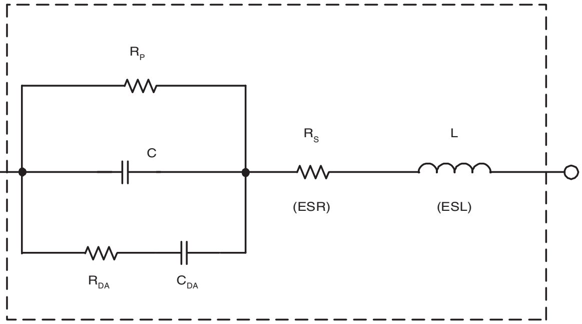


Figure 7-1: A nonideal capacitor equivalent circuit includes parasitic elements

Note that capacitor phenomena aren't that easy to separate out. The matching of phenomena and models is for convenience in explanation. Inductance, L—the *equivalent series inductance*, or ESL—models the inductance of the leads and plates. Finally, resistance R_{DA} and capacitance C_{DA} together form a simplified model of a phenomenon known as *dielectric absorption,* or DA. It can ruin fast and slow circuit dynamic performance. In a real capacitor, R_{DA} and C_{DA} extend to include multiple parallel sets. These parasitic RC elements can act to degrade timing circuits substantially, and the phenomenon is discussed further below.

Dielectric Absorption

Dielectric absorption, which is also known as "soakage" and sometimes as "dielectric hysteresis," is perhaps the least understood and potentially most damaging of various capacitor parasitic effects. Upon discharge, most capacitors are reluctant to give up all of their former charge, due to this memory consequence.

Figure 7-2 illustrates this effect. On the left of the diagram, after being charged to the source potential of V volts at time t_0, the capacitor is shorted by the switch S1 at time t_1, discharging it. At time t_2, the capacitor is then open-circuited; a residual voltage slowly builds up across its terminals and reaches a nearly constant value. This error voltage is due to DA, and is shown in the right figure, a time/voltage representation of the charge/discharge/recovery sequence. Note that the recovered voltage error is proportional to both the original charging voltage V, as well as the rated DA for the capacitor in use.

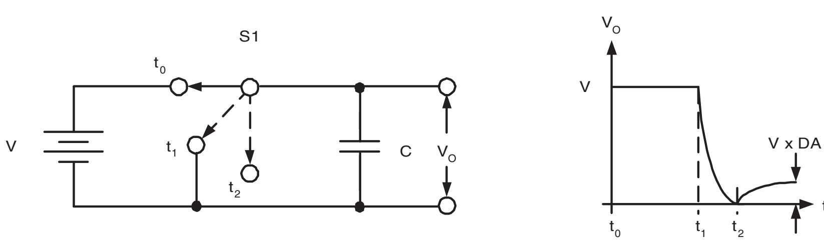


Figure 7-2: A residual open-circuit voltage after charge/discharge characterizes capacitor dielectric absorption

Standard techniques for specifying or measuring dielectric absorption are few and far between. Measured results are usually expressed as the percentage of the original charging voltage that reappears across the capacitor. Typically, the capacitor is charged for a long period, then shorted for a shorter established time. The capacitor is then allowed to recover for a specified period, and the residual voltage is then measured (see Reference 8 for details). While this explanation describes the basic phenomenon, it is important to note that real-world capacitors vary quite widely in their susceptibility to this error, with their rated DA ranging from well below to above 1%, the exact number being a function of the dielectric material used.

In practice, DA makes itself known in a variety of ways. Perhaps an integrator refuses to reset to zero, a voltage-to-frequency converter exhibits unexpected nonlinearity, or a sample-hold (SH) exhibits varying errors. This last manifestation can be particularly damaging in a data-acquisition system, where adjacent channels may be at voltages which differ by nearly full scale, as shown below.

Figure 7-3 illustrates the case of DA error in a simple SH. On the left, switches S1 and S2 represent an input multiplexer and SH switch, respectively. The multiplexer output voltage is V_X, and the sampled voltage held on C is V_Y, which is buffered by the op amp for presentation to an ADC. As can be noted by the timing diagram on the right, a DA error voltage, $\in$, appears in the hold mode, when the capacitor is effectively open circuit. This voltage is proportional to the difference of voltages V1 and V2, which, if at opposite extremes of the dynamic range, exacerbates the error. As a practical matter, the best solution for good performance in terms of DA in a SH is to use only the best capacitor.

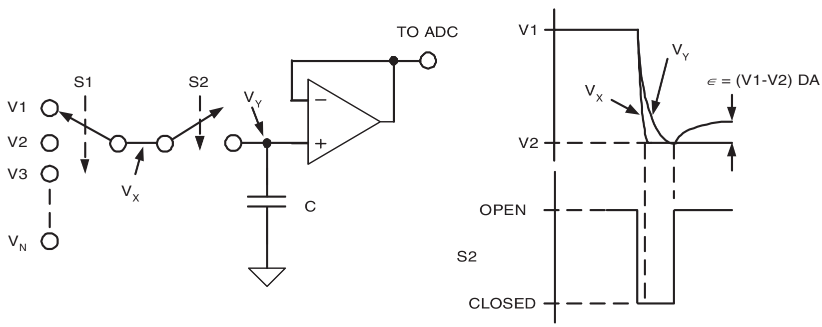


Figure 7-3: Dielectric absorption induces errors in SH applications

The DA phenomenon is a characteristic of the dielectric material itself, although inferior manufacturing processes or electrode materials can also affect it. DA is specified as a percentage of the charging voltage. It can range from a low of 0.02% for Teflon, polystyrene, and polypropylene capacitors, up to a high of 10% or more for some electrolytics. For some time frames, the DA of polystyrene can be as low as 0.002%.

Common high-K ceramics and polycarbonate capacitor types display typical DA on the order of 0.2%, it should be noted this corresponds to ½ LSB at only 8 bits. Silver mica, glass, and tantalum capacitors typically exhibit even larger DA, ranging from 1.0% to 5.0%, with those of polyester devices failing in the vicinity of 0.5%. As a rule, if the capacitor spec sheet doesn't specifically discuss DA *within your time frame and voltage range*, exercise caution. Another type with lower *specified* DA is likely a better choice.

DA can produce long tails in the transient response of fast-settling circuits, such as those found in high-pass active filters or ac amplifiers. In some devices used for such applications, Figure 7-1's R_{DA}-C_{DA} model of DA can have a time constant of milliseconds. Much longer time constants are also quite usual. In fact, several paralleled R_{DA}-C_{DA} circuit sections with a wide range of time constants can model some devices. In fast-charge, fast-discharge applications, the behavior of the DA mechanism resembles "analog memory"; the capacitor in effect tries to remember its previous voltage.

The effects of DA can be compensated for in some designs if it is simple and easily characterized, and the user is willing to do custom tweaking. In an integrator, for instance, the output signal can be fed back through a suitable compensation network, tailored to cancel the circuit equivalent of the DA by placing a negative impedance effectively in parallel. Such compensation has been shown to improve SH circuit performance by factors of 10 or more (Reference 6).

Capacitor Parasitics and Dissipation Factor

In Figure 7-1, a capacitor's leakage resistance, R_P, the effective series resistance, R_S, and effective series inductance, L, act as parasitic elements, which can degrade an external circuit's performance. The effects of these elements are often lumped together and defined as a dissipation factor, or D_F.

A capacitor's leakage is the small current that flows through the dielectric when a voltage is applied. Although modeled as a simple insulation resistance (R_P) in parallel with the capacitor, the leakage actually is nonlinear with voltage. Manufacturers often specify leakage as a megohm-microfarad product, which describes the dielectric's self-discharge time constant, in seconds. It ranges from a low of 1 s or less for high-leakage capacitors, such as electrolytic devices, to the hundreds of seconds for ceramic capacitors. Glass devices exhibit self-discharge time-constants of 1,000 or more; but the best leakage performance is shown by Teflon and the film devices (polystyrene, polypropylene), with time constants exceeding 1,000,000 megohm-microfarads. For such a device, external leakage paths—created by surface contamination of the device's case or in the associated wiring or physical assembly—can overshadow the internal dielectric-related leakage.

Effective series inductance, ESL (Figure 7-1) arises from the inductance of the capacitor leads and plates which, particularly at the higher frequencies, can turn a capacitor's normally capacitive reactance into an inductive reactance. Its magnitude strongly depends on construction details within the capacitor. Tubular wrapped-foil devices display significantly more lead inductance than molded radial-lead configurations. Multilayer ceramic and film-type devices typically exhibit the lowest series inductance, while ordinary tantalum and aluminum electrolytics typically exhibit the highest. Consequently, standard electrolytic types, if used alone, usually prove insufficient for *high speed* local bypassing applications. Note however that there also are more specialized aluminum and tantalum electrolytics available, which may be suitable for higher speed uses. These are the types generally designed for use in switch-mode power supplies, which are covered more completely in a following section.

Manufacturers of capacitors often specify effective series impedance by means of impedance-versus-frequency plots. Not surprisingly, these curves show graphically a predominantly capacitive reactance at low frequencies, with rising impedance at higher frequencies because of the effect of series inductance.

Effective series resistance, ESR (resistor R_S of Figure 7-1), is made up of the resistance of the leads and plates. As noted, many manufacturers lump the effects of ESR, ESL, and leakage into a single parameter called *dissipation factor*, or DF. Dissipation factor measures the basic inefficiency of the capacitor. Manufacturers define it as the ratio of the energy lost to energy stored per cycle by the capacitor. The ratio of ESR to total capacitive reactance—at a specified frequency—approximates the dissipation factor, which turns out to be equivalent to the reciprocal of the figure of merit, Q. Stated as an approximation, $Q \approx 1/DF$ (with DF in numeric terms). For example, a DF of 0.1% is equivalent to a fraction of 0.001; thus the inverse in terms of Q would be 1000.

Dissipation factor often varies as a function of both temperature and frequency. Capacitors with mica and glass dielectrics generally have DF values from 0.03% to 1.0%. For ordinary ceramic devices, DF ranges from a low of 0.1% to as high as 2.5% at room temperature. And electrolytics usually exceed even this level. The film capacitors are the best as a group, with DFs of less than 0.1%. Stable-dielectric ceramics, notably the NP0 (also called COG) types, have DF specs comparable to films (more below).

Tolerance, Temperature, and Other Effects

In general, precision capacitors are expensive and—even then—not necessarily easy to buy. In fact, choice of capacitance is limited both by the range of available values and by tolerances. In terms of size, the better performing capacitors in the film families tend to be limited in practical terms to 10 µF or less (for dual reasons of size and expense). In terms of low value tolerance, ±1% is possible for NP0 ceramic and some film devices, but with possibly unacceptable delivery times. Many film capacitors can be made available with tolerances of less than ±1%, but on a special order basis only.

Most capacitors are sensitive to temperature variations. DF, DA, and capacitance value are all functions of temperature. For some capacitors, these parameters vary approximately linearly with temperature, in others they vary quite nonlinearly. Although it is usually not important for SH applications, an excessively large *temperature coefficient* (TC, measured in ppm/°C) can prove harmful to the performance of precision integrators, voltage-to-frequency converters, and oscillators. NP0 ceramic capacitors, with TCs as low as 30 ppm/°C, are the best for stability, with polystyrene and polypropylene next best, with TCs in the 100–200 ppm/°C range. On the other hand, when capacitance stability is important, one should stay away from types with TCs of more than a few hundred ppm/°C, or in fact any TC that is nonlinear.

A capacitor's maximum working temperature should also be considered, in light of the expected environment. Polystyrene capacitors, for instance, melt near 85°C, compared to Teflon's ability to survive temperatures up to 200°C.

Sensitivity of capacitance and DA to applied voltage, expressed as *voltage coefficient*, can also hurt capacitor performance within a circuit application. Although capacitor manufacturers don't always clearly specify voltage coefficients, the user should always consider the possible effects of such factors. For instance, when maximum voltages are applied, some high-K ceramic devices can experience a decrease in capacitance of 50% or more. This is an inherent distortion producer, making such types unsuitable for signal path filtering, for example, and better suited for supply bypassing. Interestingly, NP0 ceramics, the stable dielectric subset from the wide range of available ceramics, do offer good performance with respect to voltage coefficient.

Similarly, the capacitance, and dissipation factor of many types vary significantly with frequency, mainly as a result of a variation in dielectric constant. In this regard, the better dielectrics are polystyrene, polypropylene, and Teflon.

Assemble Critical Components Last

The designer's worries don't end with the design process. Some common printed circuit assembly techniques can prove ruinous to even the best designs. For instance, some commonly used cleaning solvents can infiltrate certain electrolytic capacitors—those with rubber end caps are particularly susceptible. Even worse, some of the film capacitors, polystyrene in particular, actually melt when contacted by some solvents. Rough handling of the leads can damage still other capacitors, creating random or even intermittent circuit problems. Etched-foil types are particularly delicate in this regard. To avoid these difficulties it may be advisable to mount especially critical components as the last step in the board assembly process—if possible.

Table 7-1
CAPACITOR COMPARISON CHART

TYPE	TYPICAL DA	ADVANTAGES	DISADVANTAGES
Polystyrene	0.001% to 0.02%	Inexpensive Low DA Good Stability (~120ppm/°C)	Damaged by Temperature >85°C Large High Inductance Vendors Limited
Polypropylene	0.001% to 0.02%	Inexpensive Low DA Stable (~200ppm/°C) Wide Range of Values	Damaged by Temperature >105°C Large High Inductance
Teflon	0.003% to 0.02%	Low DA Available Good Stability Operational Above 125°C Wide Range of Values	Expensive Large High Inductance
Polycarbonate	0.1%	Good Stability Low Cost Wide Temperature Range Wide Range of Values	Large DA Limits to 8-Bit Applications High Inductance
Polyester	0.3% to 0.5%	Moderate Stability Low Cost Wide Temperature Range Low Inductance (Stacked Film)	Large DA Limits to 8-Bit Applications High Inductance (Conventional)
NP0 Ceramic	<0.1%	Small Case Size Inexpensive, Many Vendors Good Stability (30ppm/°C) 1% Values Available Low Inductance (Chip)	DA Generally Low (May Not be Specified) Low Maximum Values (≤10nF)
Monolithic Ceramic (High K)	>0.2%	Low Inductance (Chip) Wide Range of Values	Poor Stability Poor DA High Voltage Coefficient
Mica	>0.003%	Low Loss at HF Low Inductance Good Stability 1% Values Available	Quite Large Low Maximum Values (≤10nF) Expensive
Aluminum Electrolytic	Very High	Large Values High Currents High Voltages Small Size	High Leakage Usually Polarized Poor Stability, Accuracy Inductive
Tantalum Electrolytic	Very High	Small Size Large Values Medium Inductance	High Leakage Usually Polarized Expensive Poor Stability, Accuracy

Table 7-1 summarizes selection criteria for various capacitor types, arranged roughly in order of decreasing DA performance. In a selection process, the general information of this table should be supplemented by consultation of current vendor's catalog information (see References at end of section).

Designers should also consider the natural failure mechanisms of capacitors. Metallized film devices, for instance, often self-heal. They initially fail due to conductive bridges that develop through small perforations in the dielectric film. But, the resulting fault currents can generate sufficient heat to destroy the bridge, thus returning the capacitor to normal operation (at a slightly lower capacitance). Of course, applications in high-impedance circuits may not develop sufficient current to clear the bridge, so the designer must be wary here.

Tantalum capacitors also exhibit a degree, of self-healing but, unlike film capacitors, the phenomenon depends on the temperature at the fault location rising slowly. Therefore, tantalum capacitors self-heal best in high impedance circuits which limit the surge in current through the capacitor's defect. Use caution therefore, when specifying tantalums for high-current applications.

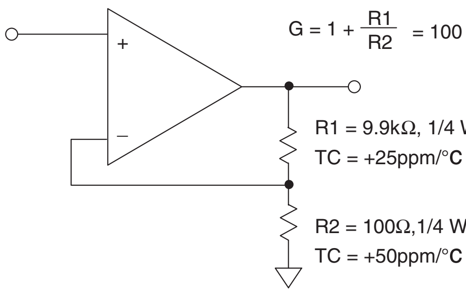
Electrolytic capacitor life often depends on the rate at which capacitor fluids seep through end caps. Epoxy end seals perform better than rubber seals, but an epoxy sealed capacitor can explode under severe reverse-voltage or overvoltage conditions. Finally, *all* polarized capacitors must be protected from exposure to voltages outside their specifications.

Resistors and Potentiometers

Designers have a broad range of resistor technologies to choose from, including carbon composition, carbon film, bulk metal, metal film, and both inductive and noninductive wire-wound types. As perhaps the most basic—and presumably most trouble-free—of components, resistors are often overlooked as error sources in high performance circuits.

An improperly selected resistor can subvert the accuracy of a 12-bit design by developing errors well in excess of 122 ppm (½ LSB).

Consider the simple circuit of Figure 7-4, showing a noninverting op amp where the 100× gain is set by R1 and R2. The TCs of these two resistors are a somewhat obvious source of error. Assume the op amp gain errors to be negligible, and that the resistors are perfectly matched to a 99/1 ratio at 25ºC. If, as noted, the resistor TCs differ by only 25 ppm/ºC, the gain of the amplifier changes by 250 ppm for a 10ºC temperature change. This is about a 1 LSB error in a 12-bit system, and a major disaster in a 16-bit system. Temperature changes, however, can limit the accuracy of the Figure 7-4 amplifier in several ways. In this circuit (as well as many op amp circuits with component-ratio defined gains), the *absolute* TC of the resistors is less



Temperature change of 10°C causes gain change of 250ppm

This is 1LSB in a 12-bit system and a disaster in a 16-bit system

Figure 7-4: Mismatched resistor TCs can induce temperature-related gain errors

important—*as long as they track one another in ratio*. But even so, some resistor types simply aren't suitable for precise work. For example, *carbon composition* units—with TCs of approximately 1,500 ppm/°C, won't work. Even if the TCs could be matched to an unlikely 1%, the resulting 15 ppm/°C differential still proves inadequate—an 8°C shift creates a 120 ppm error.

Many manufacturers offer metal film and bulk metal resistors, with absolute TCs ranging between ±1 and ±100 ppm/°C. Be aware, though; TCs can vary a great deal, particularly among discrete resistors from different batches. To avoid this problem, more expensive matched resistor pairs are offered by some manufacturers, with temperature coefficients that track one another to within 2 to 10 ppm/°C. Low priced thin-film networks have good relative performance and are widely used.

Suppose, as shown in Figure 7-5, R1 and R2 are ¼W resistors with identical 25 ppm/ºC TCs. Even when the TCs are identical, there can still be significant errors. When the signal input is zero, the resistors dissipate no heat. But, if it is 100 mV, there is 9.9 V across R1, which then dissipates 9.9 mW. It will experience a temperature rise of 1.24ºC (due to a 125ºC/W ¼W resistor thermal resistance). This 1.24ºC rise causes a resistance change of 31 ppm, and thus a corresponding gain change. But R2, with only 100mV across it, is only heated a negligible 0.0125ºC. The resulting 31 ppm net gain error represents a full-scale error of ½ LSB at 14 bits, and is a disaster for a 16-bit system.

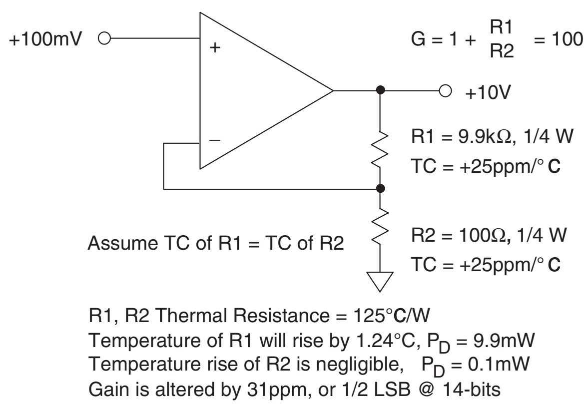


Figure 7-5: Uneven power dissipation between resistors with identical TCs can also introduce temperature-related gain errors

Even worse, the effects of this resistor self-heating also create easily calculable *nonlinearity errors*. In the Figure 7-5 example, with one-half the voltage input, the resulting self-heating error is only 15 ppm. In other words, the stage gain is not constant at ½ and full scale (nor is it so at other points), as long as uneven temperature shifts exist between the gain-determining resistors. This is by no means a worst-case example; physically smaller resistors would give worse results, due to higher associated thermal resistance.

These, and similar errors, are avoided by selecting critical resistors that are accurately matched for both value and TC, are well derated for power, and have tight thermal coupling between those resistors where matching is important. This is best achieved by using a resistor network on a single substrate—such a network may either be within an IC, or a separately packaged thin-film resistor network.

When the circuit resistances are very low (≤10 Ω), *interconnection stability* also becomes important. For example, while often overlooked as an error, the resistance TC of typical copper wire or printed circuit traces can add errors. The TC of copper is typically ~3,900 ppm/°C. Thus a precision 10 Ω, 10 ppm/°C wirewound resistor with 0. 1 Ω of copper interconnect effectively becomes a 10.1 Ω resistor with a TC of nearly 50 ppm/°C.

One final consideration applies mainly to designs that see widely varying ambient temperatures: a phenomenon known as *temperature retrace* describes the change in resistance which occurs after a specified number of cycles of exposure to low and high ambients with constant internal dissipation. Temperature retrace can exceed 10 ppm/°C, even for some of the better thin-film components.

In summary, to design resistance-based circuits for minimum temperature-related errors, consider the points noted in Figure 7-6 (along with their cost).

- Closely match resistance TCs
- Use resistors with low absolute TCs
- Use resistors with low thermal resistance (higher power ratings, larger cases)
- Tightly couple matched resistors thermally (use standard common-substrate networks)
- For large ratios consider using stepped attenuators

Figure 7-6: A number of points are important towards minimizing temperature- related errors in resistors

Resistor Parasitics

Resistors can exhibit significant levels of parasitic inductance or capacitance, especially at high frequencies. Manufacturers often specify these parasitic effects as a reactance error, in % or ppm, based on the ratio of the difference between the impedance magnitude and the dc resistance, to the resistance, at one or more frequencies.

Wirewound resistors are especially susceptible to difficulties. Although resistor manufacturers offer wirewound components in either normal or noninductively wound form, even noninductively wound resistors create headaches for designers. These resistors still appear slightly inductive (of the order of 20 µH) for values below 10 kΩ. Above 10 kΩ the same style resistors actually exhibit 5 pF of shunt capacitance.

These parasitic effects can raise havoc in dynamic circuit applications. Of particular concern are applications using wirewound resistors with values both greater than 10 kΩ. Here it isn't uncommon to see peaking, or even oscillation. These effects become more evident at low kHz frequency ranges.

Even in low-frequency circuit applications, parasitic effects in wirewound resistors can create difficulties. Exponential settling to 1 ppm may take 20 time constants or more. The parasitic effects associated with wirewound resistors can significantly increase net circuit settling time to beyond the length of the basic time constants.

Unacceptable amounts of parasitic reactance are often found even in resistors that aren't wirewound. For instance, some metal-film types have significant interlead capacitance, which shows up at high frequencies. In contrast, when considering this end-end capacitance, carbon resistors do the best at high frequencies.

Thermoelectric Effects

Another more subtle problem with resistors is the *thermocouple effect*, also sometimes referred to as *thermal EMF*. Wherever there is a junction between two different metallic conductors, a thermoelectric voltage results. The thermocouple effect is widely used to measure temperature, as described in detail within Chapter 4. However, in any low level precision op amp circuit it is also a potential source of inaccuracy, since wherever two different conductors meet, a thermocouple is formed (whether we like it or not). In fact, in many cases, it can easily produce the dominant error within an otherwise precision circuit design.

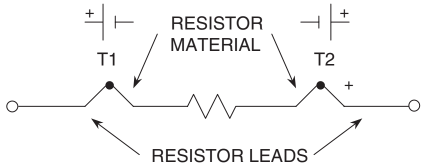
Parasitic thermocouples will cause errors when and if the various junctions forming the parasitic thermocouples are at different temperatures. With two junctions present on each side of the signal being processed within a circuit, by definition at least one thermocouple pair is formed. If the two junctions of this thermocouple pair are at different temperatures, there will be a net temperature dependent error voltage produced. Conversely, if the two junctions of a parasitic thermocouple pair are kept at an identical temperature, then the net error produced will be zero, as the voltages of the two thermocouples effectively will be canceled.

This is a critically important point, since in practice we cannot avoid connecting dissimilar metals together to build an electronic circuit. But, what we can do is carefully control temperature differentials across the circuit, so such that the undesired thermocouple errors cancel one another.

The effect of such parasitics is very hard to avoid. To understand this, consider a case of making connections *with copper wire only*. In this case, even a junction formed by different copper wire alloys can have a thermoelectric voltage that is a small fraction of 1 μV/ºC. And, taking things a step further, even such apparently benign components as resistors contain parasitic thermocouples, with potentially even stronger effects.

For example, consider the resistor model shown in Figure 7-7. The two connections between the resistor material and the leads form thermocouple junctions, T1 and T2. This thermocouple EMF can be as high as 400 μV/ºC for some carbon composition resistors, and as low as 0.05 μV/ºC for specially constructed resistors (see Reference 15). Ordinary metal film resistors (RN-types) are typically about 20 μV/ºC.

Note that these thermocouple effects are relatively unimportant for ac signals. Even for dc-only signals, they will nicely cancel one another if, as noted above, the entire resistor is at a uniform temperature. However, if there is significant power dissipation in a resistor, or if its orientation with respect to a heat source is nonsymmetrical, this can cause one of its ends to be warmer than the other, causing a net thermocouple



TYPICAL RESISTOR THERMOCOUPLE EMFs

- CARBON COMPOSITION ≈ 400 μV/ °C
- METAL FILM ≈ 20 μV/ °C
- EVENOHM OR MANGANIN WIREWOUND ≈ 2 μV/ °C
- RCD Components HP-Series ≈ 0.05 μV/ °C

Figure 7-7: Every resistor contains two thermocouples, formed between the leads and resistance element

error voltage. Using ordinary metal film resistors, an end-to-end temperature differential of 1ºC causes a thermocouple voltage of about 20 µV. This error level is quite significant compared to the offset voltage drift of a precision op amp like the OP177, and extremely significant when compared to chopper-stabilized op amps, with their drifts of <1 µV/°C.

Figure 7-8 shows how resistor orientation can make a difference in the net thermocouple voltage. In the left diagram, standing the resistor on end in order to conserve board space will invariably cause a temperature gradient across the resistor, especially if it is dissipating any significant power. In contrast, placing the resistor flat on the PC board as shown at the right will generally eliminate the gradient. An exception might occur, if there is end-to-end resistor airflow. For such cases, orienting the resistor axis perpendicular to the airflow will minimize this source of error, since this tends to force the resistor ends to the same temperature.

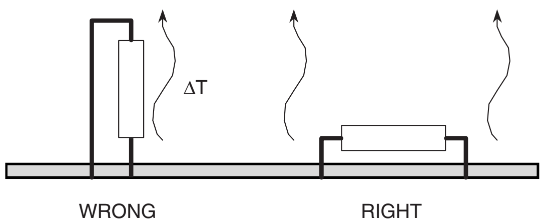


Figure 7-8: The effects of thermocouple EMFs generated by resistors can be minimized by orientation that normalizes the end temperatures

Note that this line of thinking should be extended to include orientation of resistors on a vertically mounted PC board. In such cases, natural convection air currents tend to flow upward across the board. Again, the resistor thermal axis should be perpendicular to convection, to minimize thermocouple effects. With tiny surface-mount resistors, the thermocouple effects can be less problematic, due to tighter thermal coupling between the resistor ends.

In general, designers should strive to avoid thermal gradients on or around critical circuit boards. Often this means thermally isolating components that dissipate significant amounts of power. Thermal turbulence created by large temperature gradients can also result in dynamic noise-like low frequency errors.

Voltage Sensitivity, Failure Mechanisms, and Aging

Resistors are also plagued by changes in value as a function of applied voltage. The deposited-oxide high megohm type components are especially sensitive, with voltage coefficients ranging from 1 ppm/V to more than 200 ppm/V. This is another reason to exercise caution in such precision applications as high-voltage dividers.

The normal failure mechanism of a resistor can also create circuit difficulties, if not carefully considered beforehand. For example, carbon-composition resistors fail safely, by turning into open circuits. Consequently, in some applications, these components can play a useful secondary role, as a fuse. Replacing such a resistor with a carbon-film type can possibly lead to trouble, since carbon-films can fail as short circuits. (Metal-film components usually fail as open circuits.)

All resistors tend to change slightly in value with age. Manufacturers specify long-term stability in terms of change—ppm/year. Values of 50 or 75 ppm/year are not uncommon among metal film resistors. For critical applications, metal-film devices should be burned in for at least one week at rated power. During burn-in, resistance values can shift by up to 100 or 200 ppm. Metal film resistors may need 4–5000 operational hours for full stabilization, especially if deprived of a burn-in period.

Resistor Excess Noise

Most designers have some familiarity with thermal, or Johnson, noise that occurs in resistors. But a less widely recognized secondary noise phenomenon is associated with resistors, and it is called *excess noise*. It can prove particularly troublesome in precision op amp and converter circuits, as it is evident only when current passes through a resistor.

To review briefly, thermal noise results from thermally induced random vibration of charge resistor carriers. Although the average current from the vibrations remains zero, instantaneous charge motions result in an instantaneous voltage across the terminals.

Excess noise on the other hand, occurs primarily when dc flows in a discontinuous medium—for example the conductive particles of a carbon composition resistor. The current flows unevenly through the compressed carbon granules, creating microscopic particle-to-particle "arcing." This phenomenon gives rise to a 1/f noise-power spectrum, in addition to the thermal noise spectrum. In other words, the excess spot noise voltage increases as the inverse square-root of frequency.

Excess noise often surprises the unwary designer. Resistor thermal noise and op amp input noise set the noise floor in typical op amp circuits. Only when voltages appear across input resistors and causes current to flow does the excess noise become a significant—and often dominant—factor. In general, carbon composition resistors generate the most excess noise. As the conductive medium becomes more uniform, excess noise becomes less significant. Carbon film resistors do better, with metal film, wirewound and bulk-metal-film resistors doing better yet.

Manufacturers specify excess noise in terms of a noise index—the number of microvolts of rms noise in the resistor in each decade of frequency per volt of dc drop across the resistor. The index can rise to 10 dB (3 microvolts per dc volt per decade of bandwidth) or more. Excess noise is most significant at low frequencies, while above 100 kHz thermal noise predominates.

Potentiometers

Trimming potentiometers (trimpots) can suffer from most of the phenomena that plague fixed resistors. In addition, users must also remain vigilant against some hazards unique to these components.

For instance, many trimpots aren't sealed, and can be severely damaged by board washing solvents, and even by excessive humidity. Vibration—or simply extensive use—can damage the resistive element and wiper terminations. Contact noise, TCs, parasitic effects, and limitations on adjustable range can all hamper trimpot circuit operation. Furthermore, the limited resolution of wirewound types and the hidden limits to resolution in cermet and plastic types (hysteresis, incompatible material TCs, slack) make obtaining and maintaining precise circuit settings anything but an "infinite resolution" process. Given this background, two rules are suggested for the potential trimpot user. Rule 1: Use infinite care and infinitesimal adjustment range to avoid infinite frustration when applying manual trimpots. Rule 2: *Consider the elimination of manual trimming potentiometers altogether, if possible.* A number of digitally addressable potentiometers (RDACs) are now available for direct application in similar circuit functions as classic trimpots (see Reference 17). There are also many low cost multi-channel voltage output DACs expressly designed for system voltage trimming.

Table 7-2 summarizes selection criteria for various fixed resistor types, both in discrete form and as part of networks. In a selection process, the general information of this table should be supplemented by consultation of current vendor's catalog information (see References at end of section).

Table 7-2
RESISTOR COMPARISON CHART

	TYPE	ADVANTAGES	DISADVANTAGES
DISCRETE	Carbon Composition	Lowest Cost High Power/Small Case Size Wide Range of Values	Poor Tolerance (5%) Poor Temperature Coefficient (1500 ppm/°C)
	Wirewound	Excellent Tolerance (0.01%) Excellent TC (1ppm/°C) High Power	Reactance is a Problem Large Case Size Most Expensive
	Metal Film	Good Tolerance (0.1%) Good TC (<1 to 100ppm/°C) Moderate Cost Wide Range of Values Low Voltage Coefficient	Must be Stabilized with Burn-In Low Power
	Bulk Metal or Metal Foil	Excellent Tolerance (to 0.005%) Excellent TC (to <1ppm/°C) Low Reactance Low Voltage Coefficient	Low Power Very Expensive
	High Megohm	Very High Values (10^8 to $10^{14}\Omega$) Only Choice for Some Circuits	High Voltage Coefficient (200ppm/V) Fragile Glass Case (Needs Special Handling) Expensive
NETWORKS	Thick Film	Low Cost High Power Laser-Trimmable Readily Available	Fair Matching (0.1%) Poor TC (>100ppm/°C) Poor Tracking TC (10ppm/°C)
	Thin Film	Good Matching (<0.01%) Good TC (<100ppm/°C) Good Tracking TC (2ppm/°C) Moderate Cost Laser-Trimmable Low Capacitance Suitable for Hybrid IC Substrate	Often Large Geometry Limited Values and Configurations

Inductance

Stray Inductance

All conductors are inductive, and at high frequencies, the inductance of even quite short pieces of wire or printed circuit traces may be important. The inductance of a straight wire of length L mm and circular cross-section with radius R mm in free space is given by the first equation shown in Figure 7-9.

The inductance of a strip conductor (an approximation to a PC track) of width W mm and thickness H mm in free space is also given by the second equation in Figure 7-9.

In real systems, these formulas both turn out to be approximate, but they do give some idea of the order of magnitude of inductance involved. They tell us that 1 cm of 0.5 mm o.d. wire has an inductance of 7.26 nH, and 1 cm of 0.25 mm PC track has an inductance of 9.59 nH. These figures are reasonably close to measured results.

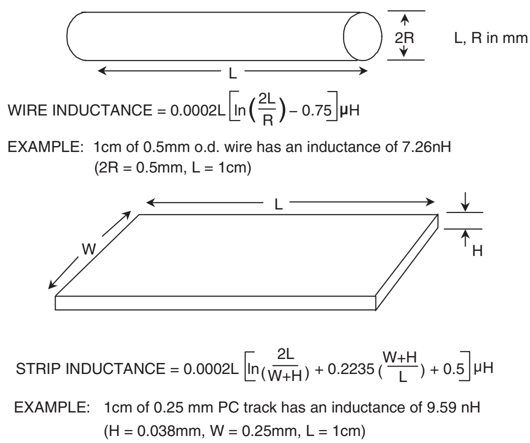


Figure 7-9: Wire and strip inductance calculations

At 10 MHz, an inductance of 7.26 nH has an impedance of 0.46 Ω, and so can give rise to 1% error in a 50 Ω system.

Mutual Inductance

Another consideration regarding inductance is the separation of outward and return currents. As we shall discuss in more detail later, Kirchoff's Law tells us that current flows in closed paths—there is always an outward and return path. The whole path forms a single-turn inductor.

This principle is illustrated by the contrasting signal trace routing arrangements of Figure 7-10. If the area enclosed within the turn is relatively large, as in the upper "nonideal" picture, the inductance (and hence the ac impedance) will also be large.

On the other hand, if the outward and return paths are closer together, as in the lower "improved" picture, the inductance will be much smaller.

Note that the nonideal signal routing case of Figure 7-10 has other drawbacks—the large area enclosed within the conductors produces extensive external magnetic fields, which may interact with other circuits,

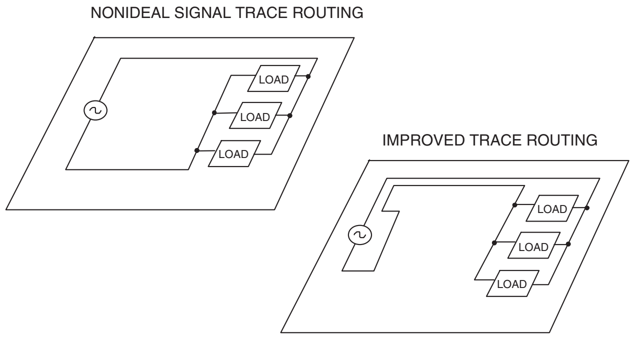


Figure 7-10: Nonideal and improved signal trace routing

causing unwanted coupling. Similarly, the large area is more vulnerable to interaction with external magnetic fields, which can induce unwanted signals in the loop.

The basic principle is illustrated in Figure 7-11, and is a common mechanism for the transfer of unwanted signals (noise) between two circuits.

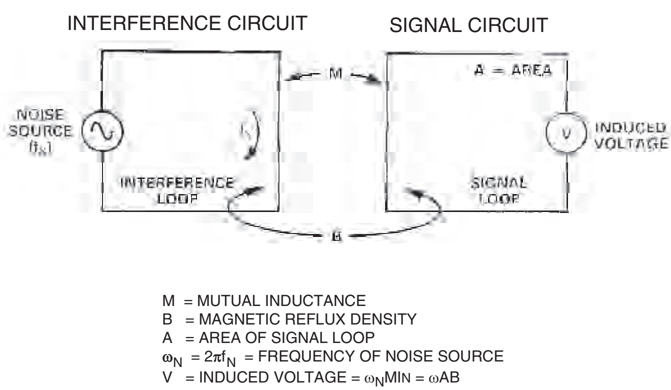


Figure 7-11: Basic principles of inductive coupling

As with most other noise sources, as soon as we define the working principle, we can see ways of reducing the effect. In this case, reducing any or all of the terms in the equations in Figure 7-11 reduces the coupling. Reducing the frequency or amplitude of the current causing the interference may be impracticable, but it is frequently possible to reduce the mutual inductance between the interfering and interfered with circuits by reducing loop areas on one or both sides and, possibly, increasing the distance between them.

A layout solution is illustrated by Figure 7-12. Here two circuits, shown as Z1 and Z2, are minimized for coupling by keeping each of the loop areas as small as is practical.

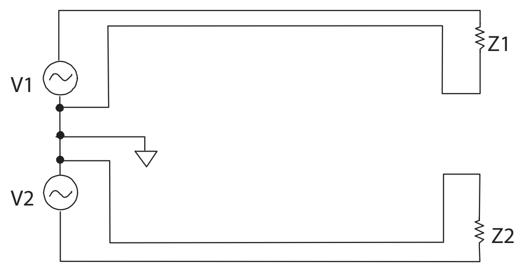


Figure 7-12: Proper signal routing and layout can reduce inductive coupling

As also illustrated in Figure 7-13, mutual inductance can be a problem in signals transmitted on cables. Mutual inductance is high in ribbon cables, especially when a single return is common to several signal circuits (top). Separate, dedicated signal and return lines for each signal circuit reduces the problem (middle). Using a cable with twisted pairs for each signal circuit as in the bottom picture is even better (but is more expensive and often unnecessary).

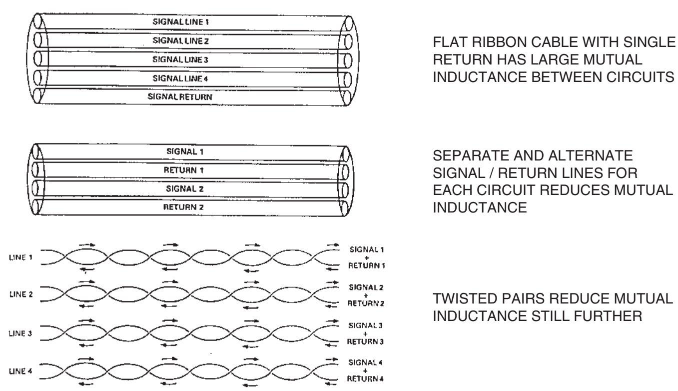


Figure 7-13: Mutual inductance and coupling within signal cabling

Shielding of magnetic fields to reduce mutual inductance is sometimes possible, but is by no means as easy as shielding an electric field with a Faraday shield (following section). HF magnetic fields are blocked by conductive material provided the skin depth in the conductor at the frequency to be screened is much less than the thickness of the conductor, and the screen has no holes (Faraday shields can tolerate small holes, magnetic screens cannot). LF and dc fields may be screened by a shield made of mu-metal sheet. Mu-metal is an alloy having very high permeability, but it is expensive, its magnetic properties are damaged by mechanical stress, and it will saturate if exposed to too high fields. Its use, therefore, should be avoided where possible.

Ringing

An inductor in series or parallel with a capacitor forms a resonant, or "tuned," circuit, whose key feature is that it shows marked change in impedance over a small range of frequency. Just how sharp the effect is depends on the relative Q of the tuned circuit. The effect is widely used to define the frequency response of narrow-band circuitry, but can also be a potential problem source.

If stray inductance and capacitance (which may or may not be stray) in a circuit should form a tuned circuit, that tuned circuit may be excited by signals in the circuit, and ring at its resonant frequency.

An example is shown in Figure 7-14, where the resonant circuit formed by an inductive power line and its decoupling capacitor may possibly be excited by fast pulse currents drawn by the powered IC.

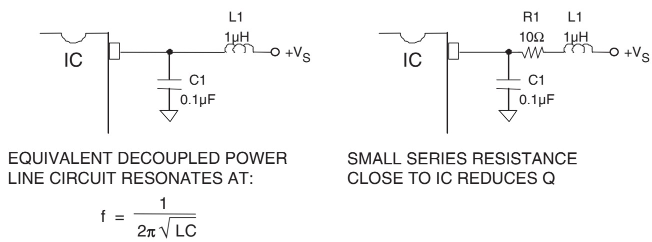


Figure 7-14: Resonant circuit formed by power line decoupling

While normal trace inductance and typical decoupling capacitances of 0.01 µF–0.1 µF will resonate well above a few MHz, an example 0.1 µF capacitor and 1 CH of inductance resonates at 500 kHz. Left unchecked, this could present a resonance problem, as shown in the left case. Should an undesired power line resonance be present, the effect may be minimized by lowering the Q of the inductance. This is most easily done by inserting a small resistance (~10 Ω) in the power line close to the IC, as shown in the right case.

Parasitic Effects in Inductors

Although inductance is one of the fundamental properties of an electronic circuit, inductors are far less common as components than are resistors and capacitors. As for precision components, they are even more rare. This is because they are harder to manufacture, less stable, and less physically robust than resistors and capacitors. It is relatively easy to manufacture stable precision inductors with inductances from nH to tens or hundreds of µH, but larger valued devices tend to be less stable, and large.

As we might expect in these circumstances, circuits are designed, where possible, to avoid the use of precision inductors. We find that stable precision inductors are relatively rarely used in precision analog circuitry, except in tuned circuits for high frequency narrow band applications. Of course, they are widely used in power filters, switching power supplies and other applications where lack of precision is unimportant (more on this in a following section).

The important features of inductors used in such applications are their current carrying and saturation characteristics, and their Q. If an inductor consists of a coil of wire with an air core, its inductance will essentially be unaffected by the current it is carrying. On the other hand, if it is wound on a core of a magnetic material (magnetic alloy or ferrite), its inductance will be nonlinear since, at high currents, the core will start to saturate. The effects of such saturation will reduce the efficiency of the circuitry employing the inductor and is liable to increase noise and harmonic generation.

As mentioned above, inductors and capacitors together form tuned circuits. Since all inductors will also have some stray capacity, all inductors will have a resonant frequency (which will normally be published on their data sheet), and should only be used as precision inductors at frequencies well below this.

Q or "Quality Factor"

The other characteristic of inductors is their Q (or "Quality Factor"), which is the ratio of the reactive impedance to the resistance, as indicated in Figure 7-15.

- $Q = 2\pi f L/R$
- The Q of an inductor or resonant circuit is a measure of the ratio of its reactance to its resistance.
- The resistance is the HF and NOT the DC value.
- The 3 dB bandwidth of a single tuned circuit is Fc/Q where Fc is the center frequency.

Figure 7-15: Inductor Q or quality factor

It is rarely possible to calculate the Q of an inductor from its dc resistance, since skin effect (and core losses if the inductor has a magnetic core) ensure that the Q of an inductor at high frequencies is always lower than that predicted from dc values.

Q is also a characteristic of tuned circuits (and of capacitors—but capacitors generally have such high Q values that it may be disregarded, in practice). The Q of a tuned circuit, which is generally very similar to the Q of its inductor (unless it is deliberately lowered by the use of an additional resistor), is a measure of its bandwidth around resonance. LC tuned circuits rarely have Q of much more than 100 (3 dB bandwidth of 1%), but ceramic resonators may have a Q of thousands, and quartz crystals tens of thousands.

Don't Overlook Anything

Remember, if a precision op amp or data-converter-based design does not meet specification, try not to overlook anything in trying to find the error sources. Analyze both active *and* passive components, trying to identify and challenge any assumptions or preconceived notions that may obscure to the facts. Take nothing for granted.

For example, when not tied down to prevent motion, cable conductors, moving within their surrounding dielectrics, can create significant static charge buildups that cause errors, especially when connected to high-impedance circuits. Rigid cables, or even costly low noise Teflon-insulated cables, are expensive alternative solutions.

As more and more high precision op amps become available, and system designs call for higher speed and increased accuracy, a thorough understanding of the error sources described in this section (as well those following) becomes more important.

Some additional discussions of passive components within a succeeding power supply filtering section complements this one. In addition, the very next section on PCB design issues also complements many points within this section. Similar comments apply to the section on EMI/RFI.

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