CHAPTER 10: PASSIVE COMPONENTS

▶ **BASIC LINEAR DESIGN**

CHAPTER 10: PASSIVE COMPONENTS

Introduction

When designing precision analog circuits, 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.

So, you've spent good money for a precision op amp or data converter, only to find that, when plugged into your 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 you blame the device, you should closely examine your 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 your circuit. And all too often these effects go unspecified (or underspecified) by passive component manufacturers.

In general, if you use data converters having 12 bits or more of resolution, or high precision op amps, 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 your problems either. Often, a *correct* 25-cent capacitor yields a better-performing, more cost-effective design than a premium-grade (expensive) part. With a few basics, understanding and analyzing passive components may prove rewarding, albeit not easy.

Notes:

SECTION 10.1: CAPACITORS

Basics

A capacitor is a passive electronic component that stores energy in the form of an electrostatic field. In its simplest form, a capacitor consists of two conducting plates separated by an insulating material called the dielectric. The capacitance is directly proportional to the surface areas of the plates, and is inversely proportional to the separation between the plates. Capacitance also depends on the dielectric constant of the substance separating the plates.

Capacitive reactance is defined as:

$$X_C = 1/\omega C = 1/2\pi fC$$
Eq. 10-1

where X_C is the capacitive reactance, ω is the angular frequency, f is the frequency in Hertz, and C is the capacitance.

Capacitive reactance is the negative imaginary component of impedance.

The complex impedance of an inductor is then given by:

$$Z = 1/j\omega C = 1/j2\pi fC$$
Eq. 10-2

where j is the imaginary number.

$$j = \sqrt{-1}$$
Eq. 10-3

Dielectric types

There are many different types of capacitors, and an understanding of their individual characteristics is absolutely mandatory to the design of practical circuits. A thumbnail sketch of capacitor characteristics is shown in the chart of Figure 10.1. Background and tutorial information on capacitors can be found in Reference 2 and many vendor catalogs.

With any dielectric, a major potential filter loss element is ESR (equivalent series resistance), the net parasitic resistance of the capacitor. ESR provides an ultimate limit to filter performance, and requires more than casual consideration, because it can vary both with frequency and temperature in some types. Another capacitor loss element is ESL (equivalent series inductance). ESL determines the frequency where the net impedance of the capacitor switches from a capacitive to inductive characteristic. This varies from as low as 10 kHz in some electrolytics to as high as 100 MHz or more in chip ceramic types. Both ESR and ESL are minimized when a leadless package is used, and all capacitor types discussed here are available in surface mount packages, which are preferable for high speed uses.

The *electrolytic* family provides an excellent, cost effective low-frequency filter component, because of the wide range of values, a high capacitance-to-volume ratio, and a broad range of working voltages. It includes *general-purpose aluminum electrolytic* types, available in working voltages from below 10 V up to about 500 V, and in size from 1 µF to several thousand µF (with proportional case sizes). All electrolytic capacitors are polarized, and thus cannot withstand more than a volt or so of reverse bias without damage. They have relatively high leakage currents (this can be tens of µA, but is strongly dependent upon specific family design, electrical size and voltage rating versus applied voltage). However, this is not likely to be a major factor for basic filtering applications.

Also included in the electrolytic family are *tantalum* types, which are generally limited to voltages of 100 V or less, with capacitance of 500 µF or less. In a given size, tantalums exhibit higher capacitance-to-volume ratios than do the general purpose electrolytics, and have both a higher frequency range and lower ESR. They are generally more expensive than standard electrolytics, and must be carefully applied with respect to surge and ripple currents.

A subset of aluminum electrolytic capacitors is the *switching* type, which is designed and specified for handling high pulse currents at frequencies up to several hundred kHz with low losses. This type of capacitor competes directly with the tantalum type in high frequency filtering applications, and has the advantage of a much broader range of available values.

More recently, high performance aluminum electrolytic capacitors using an organic semiconductor electrolyte have appeared. These *OS-CON* families of capacitors feature appreciably lower ESR and higher frequency range than do the other electrolytic types, with an additional feature of low low-temperature ESR degradation.

Film capacitors are available in very broad ranges of values and an array of dielectrics, including polyester, polycarbonate, polypropylene, and polystyrene. Because of the low dielectric constant of these films, their volumetric efficiency is quite low, and a 10 µF/50 V polyester capacitor (for example) is actually a handful. Metalized (as opposed to foil) electrodes does help to reduce size, but even the highest dielectric constant units among film types (polyester, polycarbonate) are still larger than any electrolytic, even using the thinnest films with the lowest voltage ratings (50 V). Where film types excel is in their low dielectric losses, a factor which may not necessarily be a practical advantage for filtering switchers. For example, ESR in film capacitors can be as low as 10 mΩ or less, and the behavior of films generally is very high in terms of Q. In fact, this can cause problems of spurious resonance in filters, requiring damping components.

Typically using a wound layer-type construction, film capacitors can be inductive, which can limit their effectiveness for high frequency filtering. Obviously, only noninductively made film caps are useful for switching regulator filters. One specific style which is noninductive is the *stacked-film* type, where the capacitor plates are cut as small overlapping linear sheet sections from a much larger wound drum of dielectric/plate

TYPE	TYPICAL DA	ADVANTAGES	DISADVANTAGES
Polystyrene	0.001% to 0.02%	Inexpensive Low DA Good stability (~120ppm/°C)	Damaged by temperatures >+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 temperatures >+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 Electrolyte	Very high	Large values High currents High voltages Small size	High leakage Usually polarized Poor stability, accuracy Inductive
Tantalum Electrolyte	Very high	Small size Large values Medium inductance	High leakage Usually polarized Expensive Poor stability, accuracy

Fig. 10.1: *Capacitor Comparison Chart*

material. This technique offers the low inductance attractiveness of a plate sheet style capacitor with conventional leads. Obviously, minimal lead length should be used for best high frequency effectiveness. Very high current polycarbonate film types are also available, specifically designed for switching power supplies, with a variety of low inductance terminations to minimize ESL.

Dependent upon their electrical and physical size, film capacitors can be useful at frequencies to well above 10 MHz. At the very high frequencies, stacked film types only should be considered. Some manufacturers are also supplying film types in leadless surface-mount packages, which eliminates the lead length inductance.

Ceramic is often the capacitor material of choice above a few MHz, due to its compact size and low loss. But the characteristics of ceramic dielectrics varies widely. Some types are better than others for various applications, especially power supply decoupling. Ceramic dielectric capacitors are available in values up to several µF in the high-K dielectric formulations of X7R and Z5U, at voltage ratings up to 200 V. NP0 (also called COG) types use a lower dielectric constant formulation, and have nominally zero TC, plus a low voltage coefficient (unlike the less stable high-K types). The NP0 types are limited in available values to 0.1 µF or less, with 0.01 µF representing a more practical upper limit.

Multilayer ceramic "chip caps" are increasingly popular for bypassing and filtering at 10 MHz or more, because their very low inductance design allows near optimum RF bypassing. In smaller values, ceramic chip caps have an operating frequency range to 1 GHz. For these and other capacitors for high frequency applications, a useful value can be ensured by selecting a value which has a self-resonant frequency *above* the highest frequency of interest.

All capacitors have some finite ESR. In some cases, the ESR may actually be helpful in reducing resonance peaks in filters, by supplying "free" damping. For example, in general purpose, tantalum and switching type electrolytics, a broad series resonance region is noted in an impedance versus frequency plot. This occurs where |Z| falls to a minimum level, which is nominally equal to the capacitor's ESR at that frequency. In an example below, this low Q resonance is noted to encompass quite a wide frequency range, several octaves in fact. Contrasted to the very high Q sharp resonances of film and ceramic caps, this low Q behavior can be useful in controlling resonant peaks.

In most electrolytic capacitors, ESR degrades noticeably at low temperature, by as much as a factor of 4 to 6 times at –55°C versus the room temperature value. For circuits where a high level of ESR is critical, this can lead to problems. Some specific electrolytic types do address this problem, for example within the HFQ switching types, the –10°C ESR at 100 kHz is no more than 2× that at room temperature. The OSCON electrolytics have an ESR versus temperature characteristic which is relatively flat.

Figure 10.2 illustrates the high frequency impedance characteristics of a number of electrolytic capacitor types, using nominal 100 µF/20 V samples. In these plots, the impedance, |Z|, vs. frequency over the 20 Hz to 200 kHz range is displayed using a high resolution 4-terminal setup. Shown in this display are performance samples for a

100 µF/25 V general purpose aluminum unit, a 120 µF/25 V HFQ unit, a 100 µF/20 V tantalum bead type, and a 100 µF/20 V OS-CON unit (lowest curve @ right). While the HFQ and tantalum samples are close in 100 kHz impedance, the general purpose unit is about four times worse. The OS-CON unit is nearly an order of magnitude lower in 100 kHz impedance than the tantalum and switching electrolytic types.
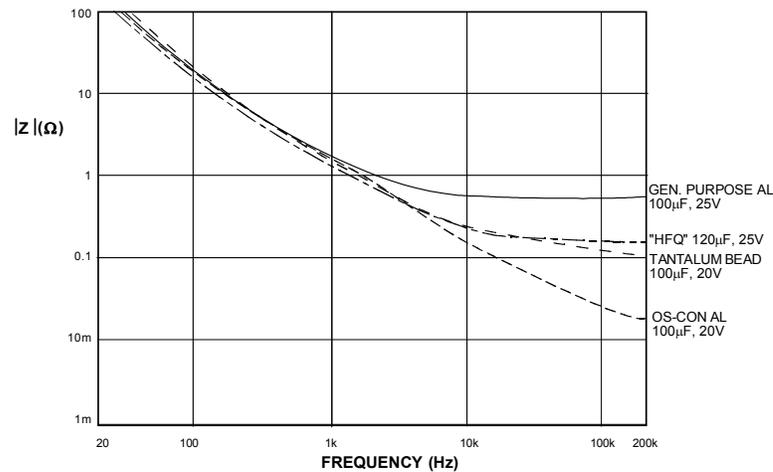
Figure 10.2: *Impedance Z(Ω) vs. Frequency for 100 µF*
Electrolytic Capacitors (AC Current = 50 mA RMS)

As noted above, all real capacitors have parasitic elements which limit their performance. As an insight into why the impedance curves of Figure 10.2 appear the way they do, a (simplified) model for a 100 µF/20 V tantalum capacitor is shown in Figure 10.3.

The electrical network representing this capacitor is shown, and it models the ESR and ESL components with simple R and L elements, plus a 1 MΩ shunt resistance. While this simple model ignores temperature and dielectric absorption effects which occur in the real capacitor, it is still sufficient for this discussion.
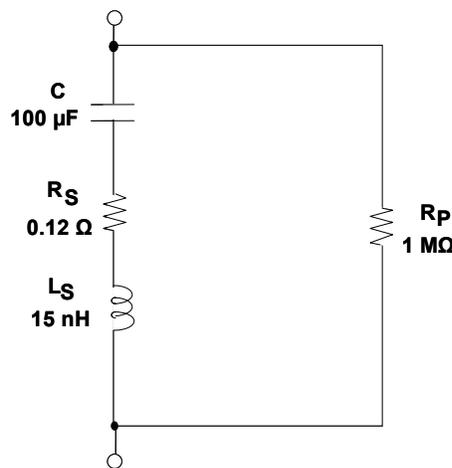
Figure 10.3: *Simplified Spice Model for a Leaded*
100 µF/20 V Tantalum Electrolytic Capacitor

When driven with a constant level of ac current swept from 10 Hz to 100 MHz, the voltage across this capacitor model is proportional to its net impedance, which is shown in Figure 10.4.
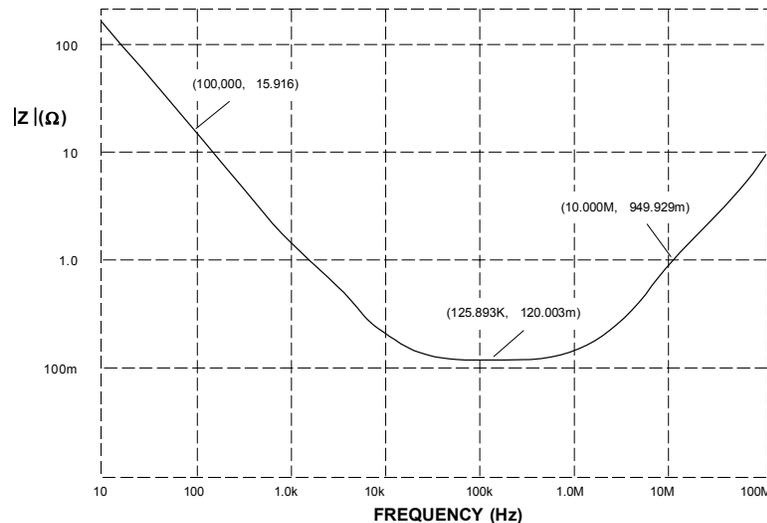
Figure 10.4: 100 µF/20 V Tantalum Capacitor Simplified Model
Impedance (Ω) vs. Frequency (Hz)

At low frequencies the net impedance is almost purely capacitive, as noted by the 100 Hz impedance of 15.9 Ω. At the bottom of this "bathtub" curve, the net impedance is determined by ESR, which is shown to be 0.12 Ω at 125 kHz. Above about 1 MHz this capacitor becomes inductive, and impedance is dominated by the effect of ESL. While this particular combination of capacitor characteristics have been chosen purposely to correspond to the tantalum sample used with Figure 10.4, it is also true that all electrolytics will display impedance curves which are similar in general shape. The minimum impedance will vary with the ESR, and the inductive region will vary with ESL (which in turn is strongly effected by package style). The simulation curve of Figure 10.4 can be considered as an extension of the 100 µF/20 V tantalum capacitor curve from Figure 10.2.

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 also 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 ppm/°C to 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 which 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.

Parasitics

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 10.5 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.
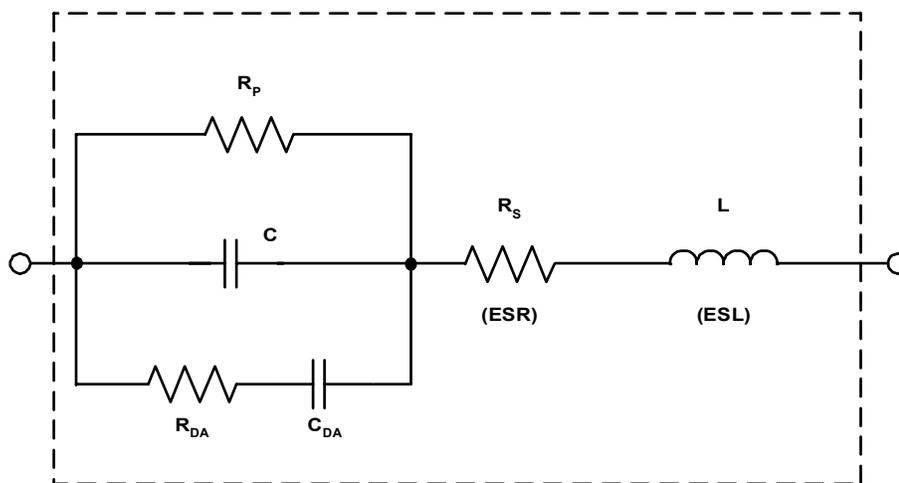

Figure 10.5: A Nonideal Capacitor Equivalent Circuit

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 10.6 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.
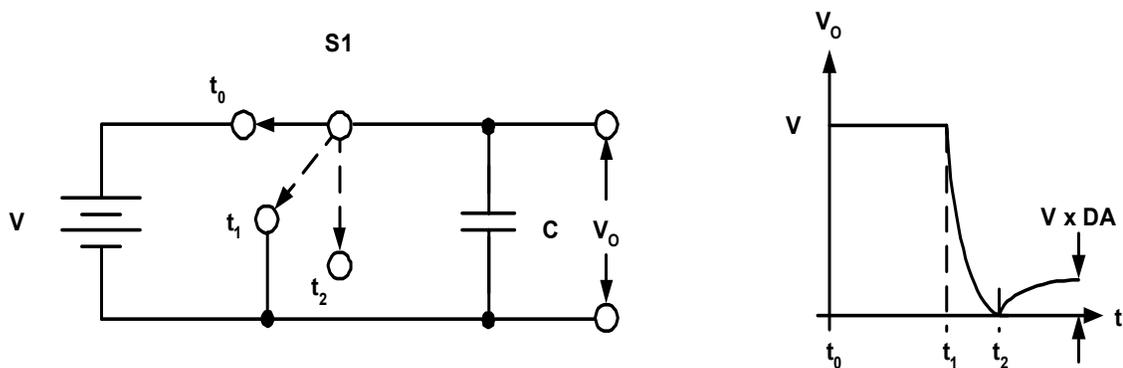
Figure 10.6: *A Residual Open-Circuit Voltage After Charge/Discharge Characterizes Capacitor Dielectric Absorption*

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 10.7 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, ϵ, 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.

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%.
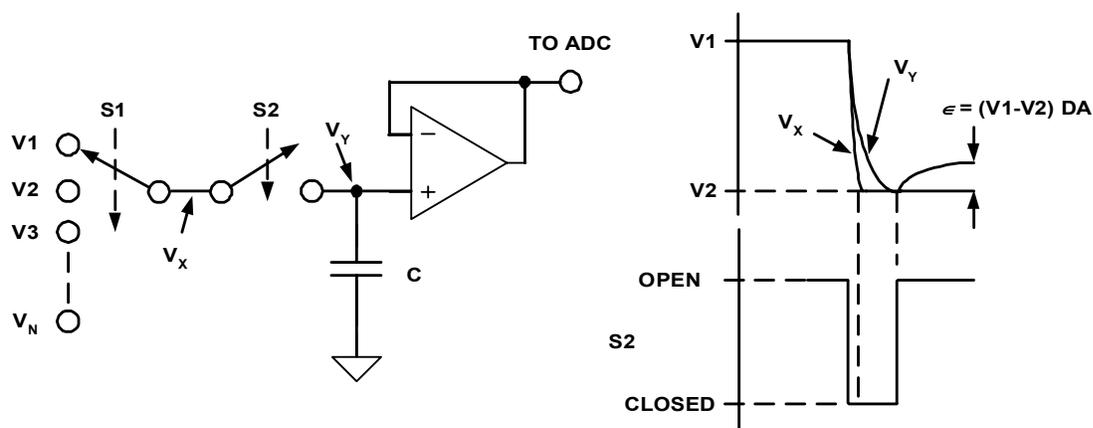
Figure 10.7: *Dielectric Absorption Induces Errors in SH Applications*

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 10.5'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.

In some designs, you can compensate for the effects of DA if it is simple and easily characterized, and you are 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 10.5, 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 DF.

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 megaohm-microfarad product, which describes the dielectric's self-discharge time constant, in seconds. It ranges from a low of 1s or less for high-leakage capacitors, such as electrolytic devices, to the 100's of seconds for ceramic capacitors. Glass devices exhibit self-discharge time-constants of 1000 or more; but the best leakage performance is shown by Teflon and the film devices (polystyrene, polypropylene), with time constants exceeding 1,000,000 megaohm-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.

Equivalent series inductance, ESL (Figure 10.5) 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 10.5), 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).

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.

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.

SECTION 10.2: RESISTORS AND POTENTIOMETERS

BASICS

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.

$$G = 1 + \frac{R1}{R2} = 100$$

R1 = 9.9kΩ, 1/4 W
TC = +25ppm/°C

R2 = 100Ω, 1/4 W
TC = +50ppm/°C
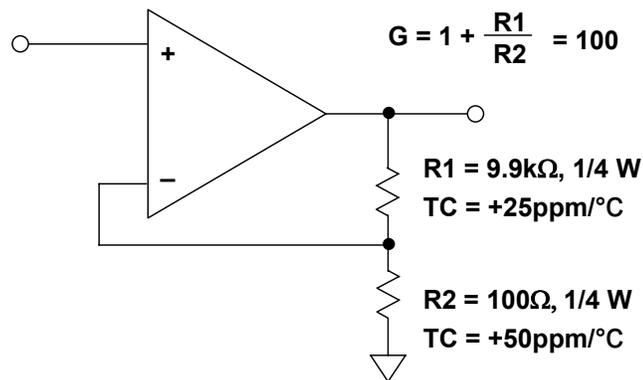
◆ Temperature change of 10°C causes gain change of 250ppm

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

Figure 10.8: *Mismatched Resistor TCs Can Induce Temperature-Related Gain Errors*

Yet, an improperly selected resistor can subvert the accuracy of a 12-bit design by developing errors well in excess of 122 ppm (½ LSB). When did you last read a resistor data sheet? You'd be surprised what can be learned from an informed review of data.
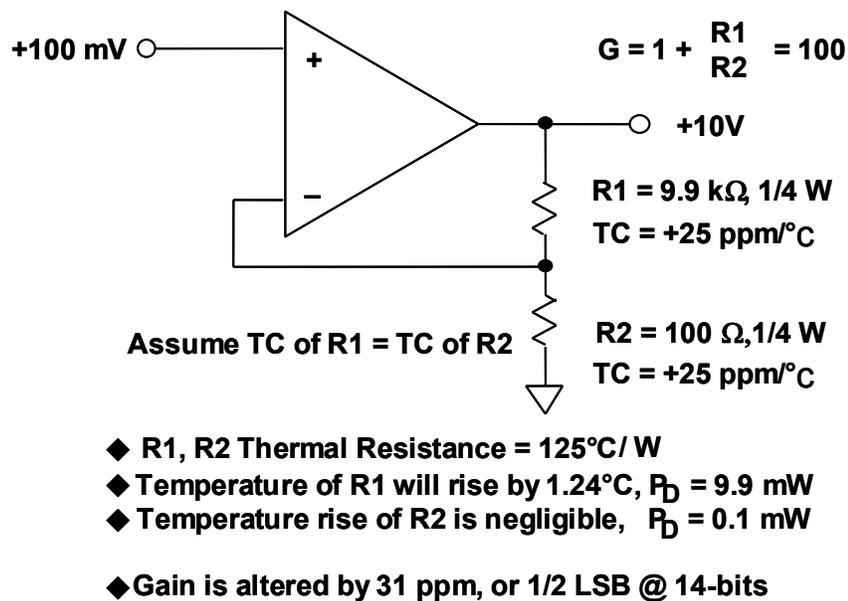
Consider the simple circuit of Figure 10.8, showing a noninverting op amp where the 100× gain is set by R1 and R2. The TCs of these two resistors are a somewhat obvious source of error. Assume the op amp gain errors to be negligible, and that the resistors are perfectly matched to a 99/1 ratio at +25ºC. 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 10.8 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 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 ppm/°C and ±100 ppm/°C. Beware, 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 ppm/°C to 10 ppm/°C. Low priced thin-film networks have good relative performance and are widely used.

$$G = 1 + \frac{R1}{R2} = 100$$

+100 mV

+10V

R1 = 9.9 kΩ, 1/4 W
TC = +25 ppm/°C

Assume TC of R1 = TC of R2

R2 = 100 Ω, 1/4 W
TC = +25 ppm/°C

◆ R1, R2 Thermal Resistance = 125°C/ W
◆ Temperature of R1 will rise by 1.24°C, P_D = 9.9 mW
◆ Temperature rise of R2 is negligible, P_D = 0.1 mW

◆ Gain is altered by 31 ppm, or 1/2 LSB @ 14-bits

Figure 10.9: Uneven Power Dissipation Between Resistors with Identical TCs
Can Also Introduce Temperature-Related Gain Errors

Suppose, as shown in Figure 10.9, R1 and R2 are ¼W resistors with identical 25 ppm/°C 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 1.24°C (due to a 125°C/W ¼W resistor thermal resistance). This 1.24°C rise causes a resistance change of 31 ppm, and thus a corresponding gain change. But R2, with only 100 mV across it, is only heated a negligible 0.0125°C. 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.

Even worse, the effects of this resistor self-heating also create easily calculable *nonlinearity errors*. In the Figure 10.9 example, with ½ 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 were 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 it may be 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.

◆ **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 10.10: Important Points for Minimizing Temperature-Related Errors in Resistors

In summary, to design resistance-based circuits for minimum temperature-related errors, consider the points noted in Figure 10.10 (along with their cost):

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. 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.
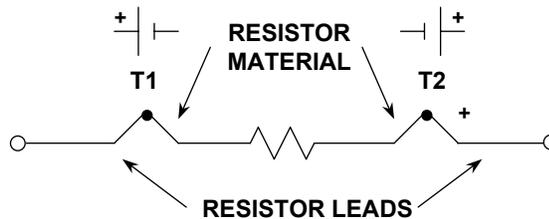
Parasitic thermocouples will cause errors when and if the various junctions forming the parasitic thermocouples are at different temperatures. With two junctions present on each side of the signal being processed within a circuit, by definition we have formed at least one thermocouple pair. If the two 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 which is a small fraction of 1µV/°C! 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 10.11. 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/°C for some carbon composition resistors, and as low as 0.05 µV/°C for specially constructed resistors. Ordinary metal film resistors (RN-types) are typically about 20 µV/°C.

Figure 10.11: Every resistor contains two thermocouples, formed between the leads and resistance element

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 error voltage. Using ordinary metal film resistors, an end-to-end temperature differential of 1°C 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 10.12 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.

WRONG   **RIGHT**
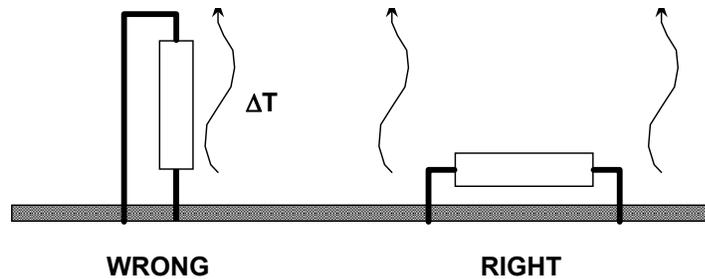
Figure 10.12: *The Effects of Thermocouple EMFs Generated by Resistors Can Be Minimized by Orientation that Normalizes the End Temperatures*

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 megaohm 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 ppm/year 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 ppm or 200 ppm. Metal film resistors may need 4000 to 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, which 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.

	TYPE	ADVANTAGES	DISADVANTAGES
DISCRETE	Carbon Composition	Lowest Cost High Power/Small Case Size Wide Range of Values	Poor Tolerance (5%) Poor Temperature Coefficient (1500ppm/°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 (<0.1%) 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

Figure 10.13: Resistor Comparison Chart

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 trim potentiometers 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 trim potentiometers 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 trim potentiometers user. Rule 1: Use infinite care and infinitesimal adjustment range to avoid infinite frustration when applying manual trim potentiometerss. 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 trim potentiometers. There are also many low cost multichannel voltage output DACs expressly designed for system voltage trimming.

SECTION 10.3: INDUCTORS

Basics

An inductor is a passive electronic component that stores energy in the form of a magnetic field. In its simplest form, an inductor consists of a wire loop or coil. The inductance is directly proportional to the number of turns in the coil. Inductance also depends on the radius of the coil and on the type of material around which the coil is wound.

Inductive reactance is defined as:

$$X_L = \omega L = 2\pi f L$$ Eq. 10-4

where X_L is the inductive reactance, ω is the angular frequency, f is the frequency in Hertz, and L is the inductance.

Inductive reactance is the positive imaginary component of impedance.

The complex impedance of an inductor is then given by:

$$Z = j\omega L = j2\pi f L$$ Eq. 10-5

where j is the imaginary number.

$$j = \sqrt{-1}$$ Eq.10-6

Inductors had been rare in circuit design, especially at lower frequencies, due to the physical size. Since inductance is the inverse mathematical function of capacitance, inductors are sometimes synthesized by placing a capacitor in the feedback network of an op amp. See figure 10.14. This technique is obviously limited to frequencies where the open-loop gain of the op amp is sufficient to support this operation. It also may not be practical at high current levels.

Inductors are popular in RF circuits since the small inductance values make them physically small. Passive LC filters are used at RF frequencies since active filters are not practical.

Switch mode power supplies are probably the common place to find inductors today. This is covered in depth in the section on switch mode regulators in the power and ground section.

It is only relatively recently that manufacturers developed the ability to fabricate inductors in monolithic semiconductor processes. This is beyond the scope of this book though.

$$L = \frac{C\,R1\,R3\,R5}{R2}$$
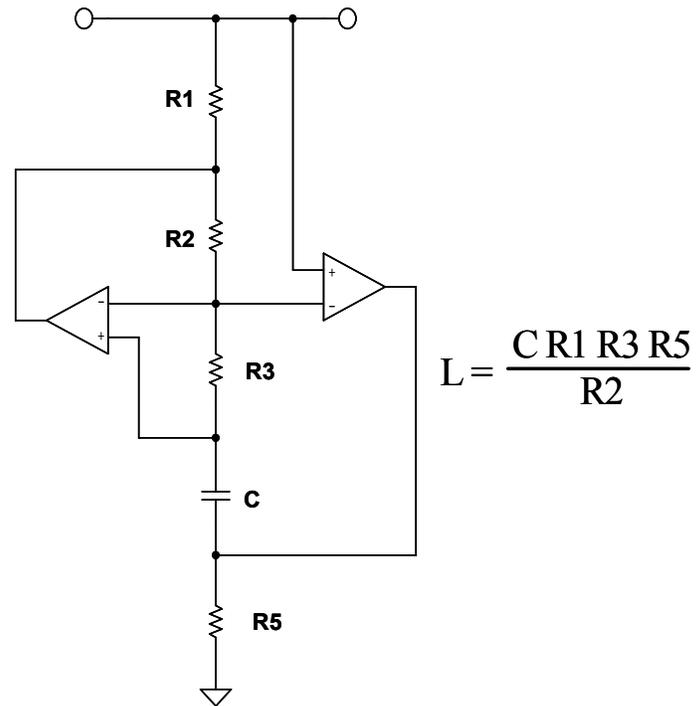
Fig 10.14:   A Synthetic Inductor.

Ferrites

Ferrites (nonconductive ceramics manufactured from the oxides of nickel, zinc, manganese, or other compounds) are extremely useful for decoupling in power supply filters. At low frequencies (<100 kHz), ferrites are inductive; thus they are useful in low-pass LC filters. Above 100 kHz, ferrites becomes resistive, an important characteristic in high-frequency filter designs. Ferrite impedance is a function of material, operating frequency range, dc bias current, number of turns, size, shape, and temperature.

Several ferrite manufacturers offer a wide selection of ferrite materials from which to choose, as well as a variety of packaging styles for the finished network. The most simple form is the *bead* of ferrite material, a cylinder of the ferrite which is simply slipped over the power supply lead to the stage being decoupled. Alternately, the *leaded ferrite bead* is the same bead, mounted by adhesive on a length of wire, and used simply as a component. More complex beads offer multiple holes through the cylinder for increased decoupling, plus other variations. Surface mount bead styles are also available.

Recently, PSpice ferrite models for Fair-Rite materials have become available that allow ferrite impedance to be estimated [Reference 12]. These models have been designed to match measured impedances rather than theoretical impedances.

A ferrite's impedance is dependent upon a number of inter-dependent variables, and is difficult to quantify analytically, thus selecting the proper ferrite is not straightforward. However, knowing the following system characteristics will make selection easier. First, determine the frequency range of the noise to be filtered. A spectrum analyzer is useful here. Second, the expected temperature range of the filter should be known, because ferrite impedance varies with temperature. Third, the dc bias current flowing through the ferrite must be known, to ensure that the ferrite does not saturate. Although models and other analytical tools may prove useful, the general guidelines given above, coupled with some experimentation with the actual filter connected to the supply output under system load conditions, should ultimately lead to the selection of the proper ferrite.

Sizing of the ferrite for current is especially important. If the saturation current of the ferrite is reached, it loses its inductive properties. This will obviously limit the ferrites usefulness in decoupling applications. This saturation current should be sized for the peak current requirements of the circuit (with some added margin, of course).

Notes:

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Notes: