SECTION 1

PRECISION SENSOR SIGNAL CONDITIONING AND TRANSMISSION

- PRECISION SIGNALS AND SYSTEM ERROR SOURCES
- SENSOR OUTPUT SIGNAL CONDITIONING EXAMPLES
- REMOTE SENSOR APPLICATION PROBLEMS:
  SHIELDING LONG CABLES EFFECTIVELY AGAINST NOISE PICKUP, FILTERING AND PROTECTION AGAINST EMI/RFI
- INPUT-STAGE RFI RECTIFICATION SENSITIVITY
- RFI RECTIFICATION: AN ANALYTICAL APPROACH
- INPUT OVERVOLTAGE PROTECTION
- TRANSMITTING PRECISION SIGNALS IN NOISY ENVIRONMENTS
- ISOLATING TRANSDUCERS

SECTION 1

PRECISION SENSOR SIGNAL CONDITIONING AND TRANSMISSION

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PRECISION SIGNALS AND SYSTEM ERROR SOURCES

Managing the total error budget in a complex high resolution precision system is extremely difficult. Every component in the system must be examined as a potential source of error.

Figure 1.1 shows accuracy (resolution) expressed as a percentage, ppm, and voltage (assuming a 10V fullscale signal range).

LEAST SIGNIFICANT BIT WEIGHTS FOR VARIOUS RESOLUTIONS

RESOLUTION	% FULLSCALE	PPM FULLSCALE	LSB WEIGHT (10V FS)
12-bits	0.0244%	244 ppm	2.44mV
16-bits	0.0015%	15 ppm	152μV
18-bits	0.00038%	3.8 ppm	38μV
20-bits	0.000095%	0.95 ppm	9.5μV
22-bits	0.000024%	0.238 ppm	2.38μV

Figure 1.1

Consider the industry-standard 741 or LM348 op amps which have an input offset voltage specification of 6mV. In a 12-bit system, this corresponds to over 2 LSBs absolute error (assuming no system calibration).

The precision OP-07 has an offset voltage specification of 150μV which is equivalent to 1 LSB absolute error in a 16-bit system. However, the OP-07 offset drift over temperature of 1.8μV/°C produces an additional 108μV error over a 60°C temperature change (25°C to 85°C). The resulting total error over temperature becomes almost 2 LSBs.

For 18-bit accuracy, even the ultra-low offset voltage OP-177 (its best grade has 10μV initial offset and 18μV total Vos drift), barely makes the 1 LSB error limit of 38μV.

For absolute accuracy at the 18-bit level and beyond, chopper stabilized amplifiers may be used to meet the DC offset and drift requirements at the expense of increased noise levels due to the chopping action. Additional filtering must be employed to take full advantage of choppers, and they are generally more noisy than bipolar op amps for bandwidths above 0.1Hz.

More and more data acquisition systems which operate at resolutions of 16-bits or greater are being designed with auto-zeroing circuits or with periodic self-calibration in order to maintain absolute accuracy.

EFFECTS OF OP AMP OFFSET AND DRIFT ON SYSTEM ACCURACY

■ 741 Amplifer Has Offset Voltage of 6mV:	2 LSBs at 12-bits
■ OP-07 Has 25°C Offset Voltage of 150μV:	1 LSB at 16-bits
■ OP-07 Has Tempco of 1.8μV/°C:	1 LSB at 16-bits, $\Delta T = 60°C$
■ OP-177 (Best Grade) Has 25°C Offset of 10μV:	1/4 LSB at 18-bits
■ OP-177 (Best Grade) Has Total Vos Drift of 18μV:	1/2 LSB at 18-bits

Figure 1.2

Besides evaluating the effects of electrical components on total error, mechanical and environmental issues must also be considered when designing precision systems. For example, the parasitic thermo-electric (bi-metallic) junction between copper and Kovar has a 18μV/°C temperature coefficient. Therefore, a 1°C temperature gradient across a PC board can generate as much as 2 LSBs error in a 20-bit system.

OTHER CONSIDERATIONS FOR HIGH ACCURACY SYSTEMS

- Chopper Stabilized Op Amps
- Auto-Zero and/or Self-Calibration
- Noise
- Mechanical Parasitic Thermoelectric Junctions

Figure 1.3

The effects of noise must also be considered in high-accuracy systems. Performance at the 22-bit level is often limited by noise. For example, the noise of the OP-177 in a 100Hz bandwidth is approximately 1μV p-p, 1/2 LSB at 22-bits. Indeed, even the tempco of the world's best reference (1ppm/°C) introduces a temperature drift of 4 LSBs/°C at 22-bits.

Clearly some extraordinary methods are necessary to achieve high levels of absolute accuracy. Some form of auto-calibration may be correct offset and gain scaling errors. Averaging data over long intervals minimizes random noise. However, these techniques often result in added cost and complexity.

In any event, all error sources (component-related and physical) must be well understood in order to successfully design precision systems. Figure 1.4 shows a list of common error sources and the approximate levels at which they become significant.

ERROR SOURCES AFFECT ABSOLUTE ACCURACY IN PRECISION SYSTEMS

ERROR SOURCE	12 Bits	14 Bits	16 Bits	18 Bits	20 Bits	22 Bits
Vos	X	X	X	X	X	X
Vos Drift			X	X	X	X
Ios			X	X	X	X
Ios Drift			X	X	X	X
Avol		X	X	X	X	X
CMRR				X	X	X
PSRR				X	X	X
Amplifier Noise				X	X	X
Resistor Tolerance	X	X	X	X	X	X
Resistor Tempco			X	X	X	X
Parasitic Tempco				X	X	X
Ground Noise	X	X	X	X	X	X
Layout Effects	X	X	X	X	X	X
Long Term Drift					X	X
Circuit Self-Heating			X	X	X	X

Figure 1.4

In a high precision system, all sources of error must be understood and evaluated. In the following sections of this seminar, we shall examine the various elements in the signal path from the transducer to the ADC. The error sources associated with each portion will be examined and appropriate remedies for them will be suggested.

SENSOR OUTPUT SIGNAL CONDITIONING EXAMPLES

There are resistive, capacitive, piezoelectric, and other types of sensors for different applications. There is no one universal interface design that can accommodate all sensor types. Within the resistive class of sensors, for example, resistances vary over a wide range, and biasing requirements differ depending on the individual sensor.

LOW IMPEDANCE, LOW VOLTAGE TRANSDUCERS

- RTD: 100Ω
- Thermocouple: Low Resistance, 10µV/°C to 50µV/°C Output Tempco
- Load-Cell, Strain Gauge: 350Ω to 10kΩ Bridge. Up to 100mV Fullscale Output
- Dynamic Microphone: 150Ω to 1500Ω, up to 50 - 100mV Fullscale Output

Figure 1.5

HIGH IMPEDANCE TRANSDUCERS

- Piezoelectric: 50kΩ to >10MΩ, µV to mV Output
- Photo-detector: >10MΩ, nA to µA Output
- Capacitive: >100MΩ

Figure 1.6

The following examples illustrate some very low level applications.

A PRECISION WEIGHT-SCALE AMPLIFIER

A precision weigh-scale transducer is usually configured as a 350Ω bridge. Figure 1.7 shows a load-cell amplifier that is powered from a single supply. The excitation voltage to the bridge must be precise and stable, otherwise it introduces an error in the measurement. In this circuit, a precision 5V reference is used as the bridge drive. The REF-195 reference can supply more than 30mA to a load, so it can drive the 350Ω bridge without the need of a buffer.

PRECISION LOAD-CELL AMPLIFIER

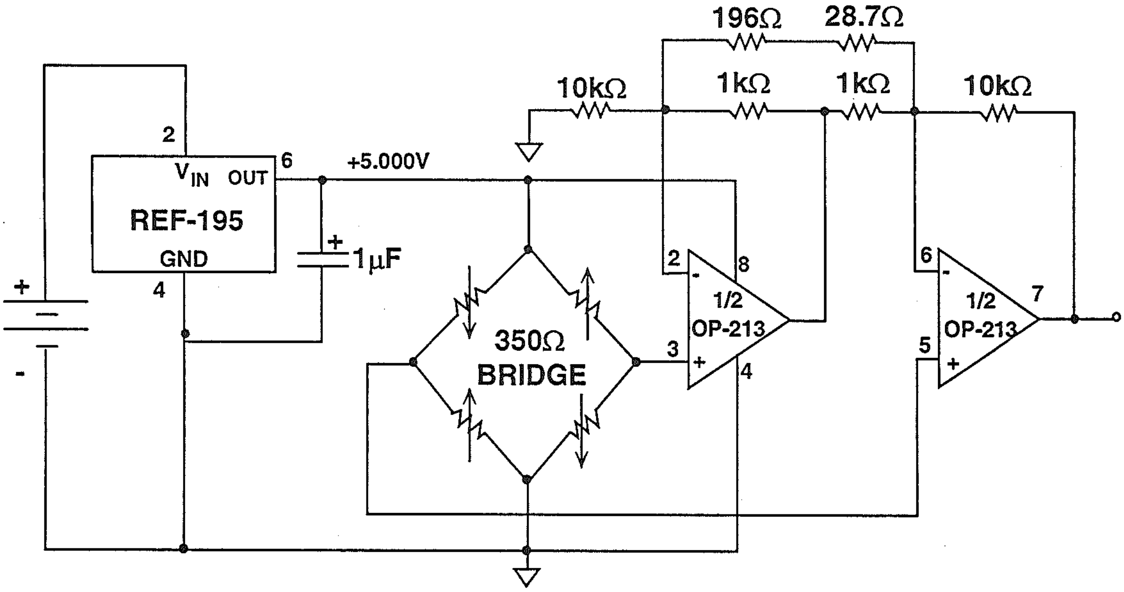


Figure 1.7

The bridge signal is amplified by the two OP-213 op amps connected as an instrumentation amplifier. The resistor network sets the gain according to the formula:

$$G = 1 + \frac{10k\Omega}{1k\Omega} + \frac{20k\Omega}{196\Omega + 28.7\Omega} = 100$$

For optimum common-mode rejection, the resistor ratios must be precise. High tolerance resistors (±0.5% or better) should be used.

For zero bridge signal, the amplifier will swing to within 2.5mV of 0V. This is the minimum output limit of the OP-213. Therefore, if an offset adjustment is required, one should start the adjustment from a positive voltage and adjust downward until the output stops changing. This is the point where the amplifier limits the swing. Because of the single supply design, the amplifier cannot sense signals which have negative polarity.

THE TMP-01 TEMPERATURE SENSOR/CONTROLLER

The TMP-01 is a new device that simplifies the job of designing a temperature controller circuit. As the block diagram in Figure 1.8 shows, the TMP-01 combines a temperature sensor with two comparators in order to provide a voltage output that is proportional to temperature, and also to include open collector outputs which signal if the temperature has exceeded a high set point or dropped below a low set point. In addition, a built in hysteresis generator adjusts the comparators' trip points by a programmable amount to ensure that the outputs do not oscillate. The core of the TMP-01 is a temperature sensor and 2.5V bandgap reference. The temperature sensor output is 5mV/K, resulting in an output voltage of 1.49V at 25°C. The amount of hysteresis and the comparator trip points are determined by three external resistors, R1, R2, and R3. The formulae for choosing these resistors are given in the TMP-01 data sheet.

TMP-01 TEMPERATURE SENSOR/CONTROLLER

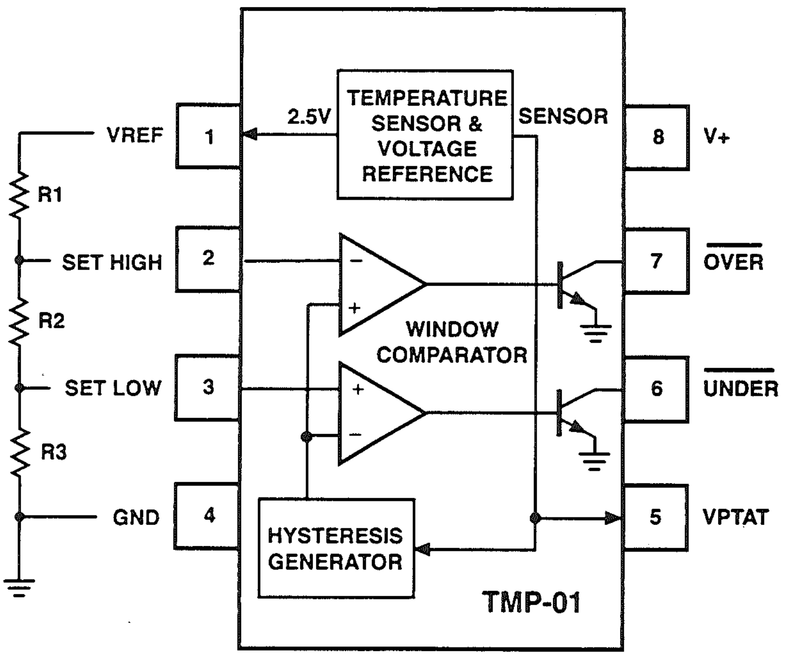
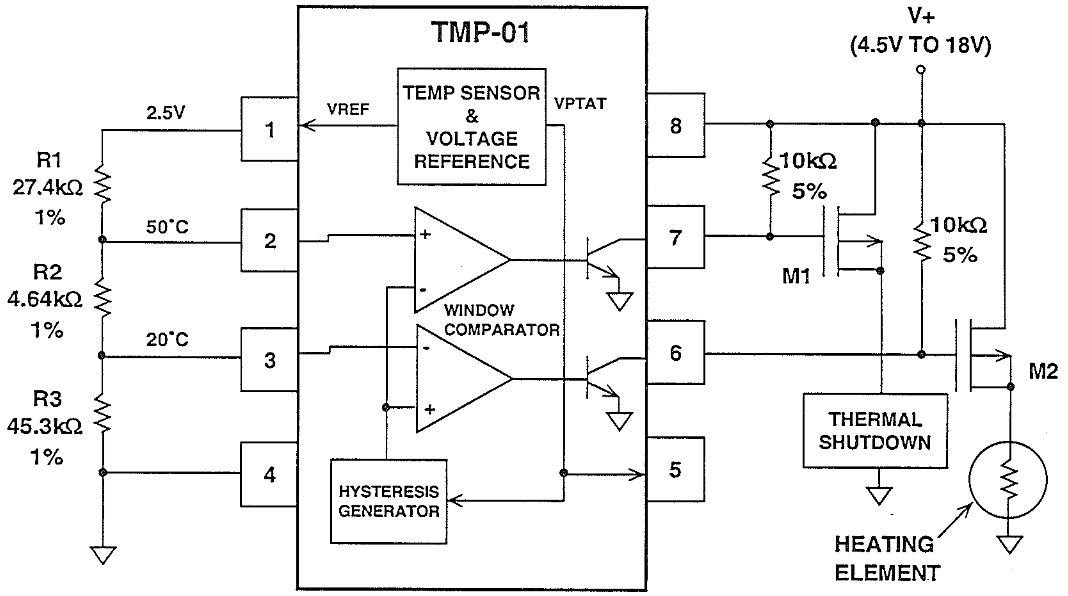


Figure 1.8

The versatility of the TMP-01 is illustrated in the temperature controller shown in Figure 1.9. In this case, the open collector outputs control a heating element and a thermal shutdown circuit to ensure that a piece of equipment is maintained within a predetermined temperature range. In this case, R1-R3 are chosen to give a low set point of 20°C and a high set point of 50°C, and 5°C of hysteresis. Thus when the temperature of the TMP-01 drops below 20°C, the negative input of the comparator has a higher voltage than the positive input, causing the open collector output at pin 6 to turn on. This brings the gate of the p-channel MOSFET, M2, low, turning on the transistor and allowing current to flow through the heating element. The heating element stays on until the temperature rises above 25°C (the set point plus the hysteresis), when the output transistor turns off M2 off. The same process occurs when the temperature rises above the high set point, except that here some type of thermal shutdown or cooling circuit is activated. Additionally, the temperature output pin of the TMP-01 can be connected to an ADC to continually monitor the system temperature. Overall, the TMP-01 combines a high degree of functionality with an easily used, space saving 8-pin dip or SOIC package.

HIGH/LOW SETPOINTS CONTROL CIRCUIT TEMPERATURE



- 5°C Hysteresis
- M1, M2: IRFR9022 or Equivalent P-Channel MOSFETs

Figure 1.9

SINGLE SUPPLY LINEARIZED RTD AMPLIFIER

Linearization of transducer outputs is more difficult when there is only one supply available. Figure 1.10 shows a single supply circuit to correct the nonlinear behavior of a Platinum Resistor Temperature Detector (RTD) device. The RTD operates in one leg of a full bridge circuit that is excited by a constant current source established by 1/2 of an OP-295 dual op amp. The bridge current is regulated by servoing the bridge current flowing into resistor RSENSE and comparing with a 200μV reference voltage derived from the REF-43.

PRECISION SINGLE SUPPLY RTD AMPLIFIER WITH LINEARIZATION

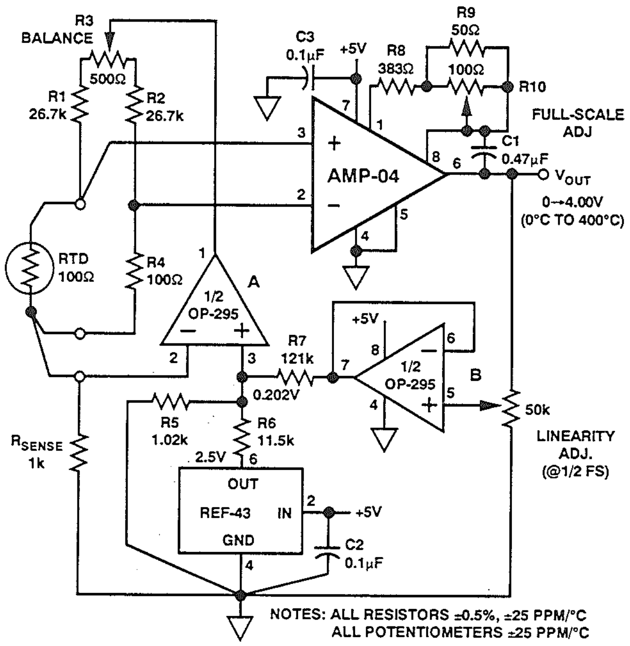


Figure 1.10

The temperature-dependent resistance change is amplified by the instrumentation amplifier AMP-04. Scaling is such that for 0°C the amplifier output is 0V, and at 400°C the amplifier output is +4.00V.

The RTD has an inherent curvature in its resistance-versus-temperature transfer function. If uncorrected, the sensor's nonlinearity would produce a 20°C error over the 400°C temperature range. This nonlinearity can be corrected by providing a small amount of positive feedback to the reference voltage, which increases the bridge current at high temperatures. The amount of positive feedback is sufficiently small as not to cause a stability problem in the circuit.

Calibration is an interactive three-step procedure. First set the FULL-SCALE and LINEARITY potentiometers to the middle of their adjustment ranges. The first calibration is made with the zero adjust, and it should be made at a voltage other than zero since zero-volts is the negative voltage limit of the circuit and is indistinguishable from a zero-volt signal. A convenient zero calibration point is +5°C. The procedure is to substitute a known, stable 101.95Ω resistor in place of the RTD and then adjust the ZERO ADJUST potentiometer for 0.050V at the output.

Next substitute the full-scale (400°C) equivalent RTD resistance of 247.04Ω. Adjust the FULL-SCALE ADJUST pot for a 4.000V at the output. Then substitute the resistance corresponding to half-scale (200°C), or 175.84Ω. Adjust the LINEARITY ADJUST pot for a 2.000V output. Since the FULL-SCALE and LINEARITY adjustments are interactive it is necessary to repeat the calibration routine once or twice until no further adjustment is necessary.

Once calibrated, the amplifier is accurate to better than ±0.5°C within the 0°C to 400°C measurement range. If a higher supply voltage is available, the measurement range can be increased.

A SINGLE SUPPLY, PRECISION PHOTODETECTOR

A common photodetector circuit is shown in Figure 1.11. The AD820 is used as the current to voltage converter. The photodiode is configured in the photovoltaic mode to prevent "dark current." The photovoltaic mode is used when accuracy is more important than speed. In this case, the output current of the photodiode can be well below 100pA. To resolve such low amounts of current, the op amp should not contribute a significant amount of bias current. compared to the photodiode. For this reason JFET op amps are commonly used as the current to voltage converters in such applications.

JFET amplifiers are usually designed to operate with dual supplies. As a result, to build a single supply photodetector, the inputs need to be biased within the common mode range of the amplifier. Doing so requires additional components, which adds cost as well as introducing possible error sources. The AD820 is a JFET amplifier designed to operate on a single supply over the range of +3V to +36V, which makes designing a single supply photodetector circuit much easier. In the circuit in Figure 1.11, the AD820's negative supply pin is connected to ground, and the inputs are biased at ground. The AD820 uses p-channel JFETs, which allows the gate to operate at the negative supply potential while still maintaining the input stage in its linear region. This particular circuit uses a 100MΩ resistor to convert the current output of the photodiode to a voltage. Care must be taken to minimize leakage paths at the inverting input node of the op amp. (Section 3 of Reference 1.)

SINGLE SUPPLY, PRECISION PHOTODETECTOR

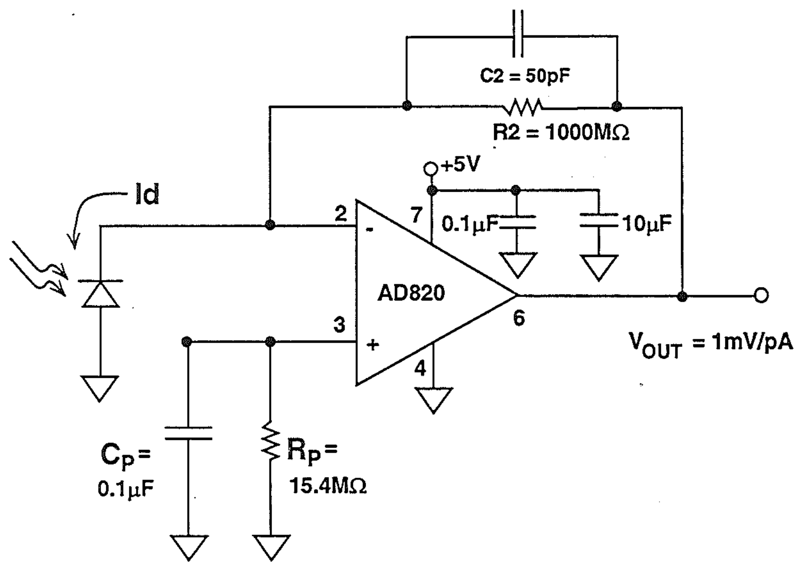


Figure 1.11

AD820 SINGLE SUPPLY FET INPUT OP AMP FEATURES

- Single Supply Operation: +3V to +36V
- Dual Supply Operation: ±1.5V to ±18V
- Output Swings Rail to Rail
- Low Power: 600µA Supply Current Maximum
- Input Offset Voltage: 200µV Maximum
- Input Bias Current: 25pA Maximum
- Low Noise: 13nV/√Hz at 10kHz
- Unity Gain Bandwidth: 1.8MHz
- Slew Rate: 3V/µs

Figure 1.12

Other features of the AD820 are listed in Figure 1.12. Another advantage of the AD820 is that its output stage can swing from rail-to-rail. In other words, the output can swing as low as the negative supply and as high as the positive supply. This feature provides the maximum amount of dynamic range for a given power supply voltage. For example, in a single +5V data acquisition system, the ADC probably has a 5V input range. In order to utilize the full input swing, the input amplifier needs to swing from ground to +5V, which is exactly what the AD820 can do. The low end accuracy of the photodetector circuit is limited mainly by the bias current of the amplifier. The AD820 has a very low 25pA maximum at room temperature. In addition, the low noise and low offset help maximize the low end resolution of this circuit, all of which makes the AD820 an ideal amplifier for single supply, precision photodetectors.

For further discussion of photodiode circuits, see Chapter 3 of Reference 1.

REMOTE SENSOR APPLICATION PROBLEMS

When a sensor is located an appreciable distance away from its electronics, noise pickup or ground loop problems frequently occur. The measurement system's circuits may be exposed to potentially damaging electrical levels. Protection circuits must be provided to prevent catastrophic failures. It is important to understand how these problems arise in order to deal with them effectively. This chapter discusses various shielding, filtering, and input protection techniques which can prove helpful in the design of a system.

SHIELDING LONG CABLES AGAINST NOISE PICKUP

A cable that is more than a few inches long may act as an antenna, picking up extraneous noise from radio frequency (RF) sources or other electromagnetic (EM) sources. In a high noise environment, it is often necessary to shield the cable. However, shielding alone is not sufficient to eliminate noise pickup. One must also ground the shield properly.

It is not always clear where to ground the shield. Grounding at the wrong place not only renders the shield ineffective, it may also introduce unwanted ground current loops.

To study this problem, the precision RTD amplifier circuit shown in Figure 1.15 was used as the basis for a series

REMOTE SENSING: AN ENGINEER'S NIGHTMARE

- RF Noise Pickup
- EMI Noise Pickup
- Where to Ground Shields to Avoid Ground Loops
- Wire Resistance Introduces Errors

Figure 1.13

AN IMPROPERLY GROUNDED CABLE SHIELD IS A POTENTIAL NOISE GENERATOR

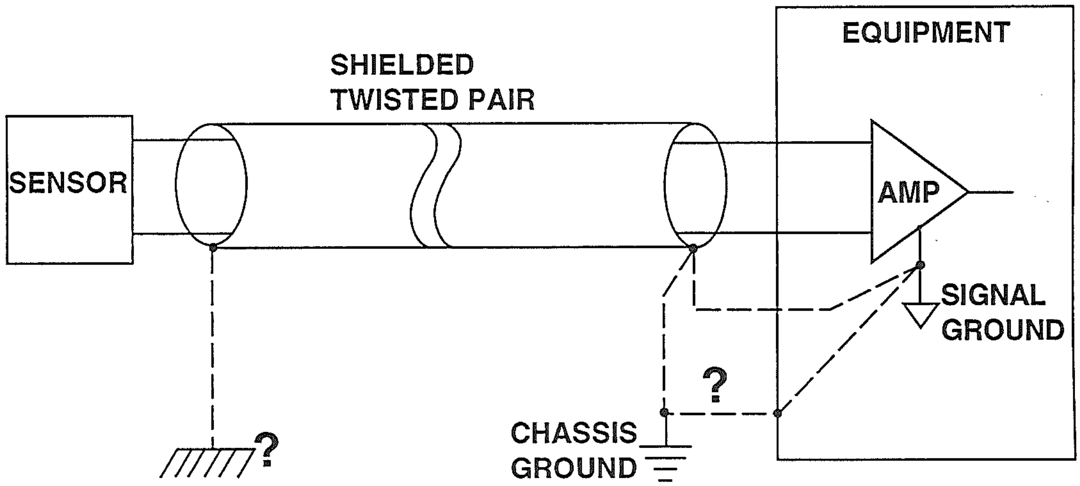


Figure 1.14

EXPERIMENT: A 100Ω RTD LOCATED 10 FEET FROM INSTRUMENT

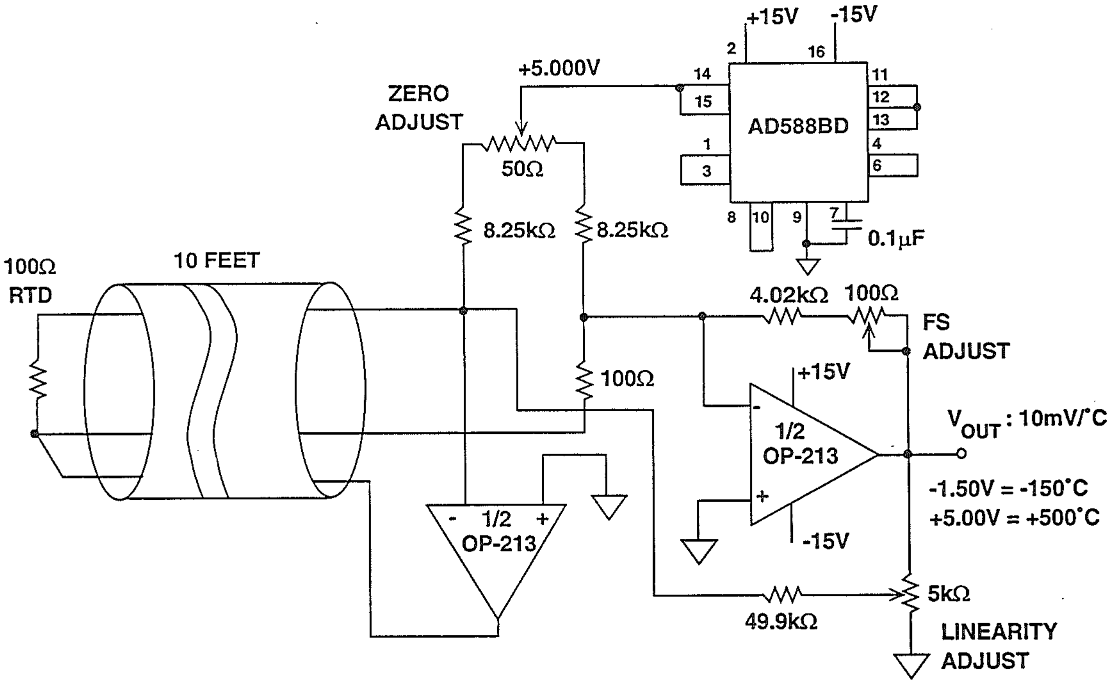
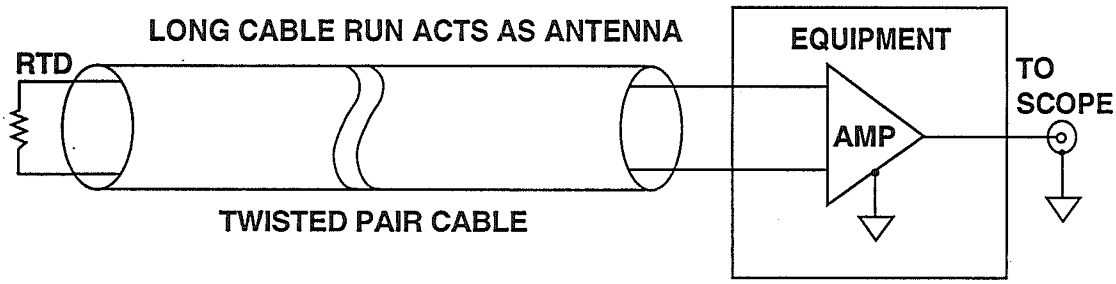


Figure 1.15

of experiments. The remote 100Ω RTD was connected to the amplifier circuit using 10 feet of four-wire cable (2 twisted pairs twisted inside a shield). Measurements were made with the shield grounded at various places along the length of the cable.

EXPERIMENT 1: UNSHIELDED, OR SHIELD LEFT FLOATING



VERTICAL SCALE: 2mV/div.
HORIZONTAL SCALE: 10ms/div.

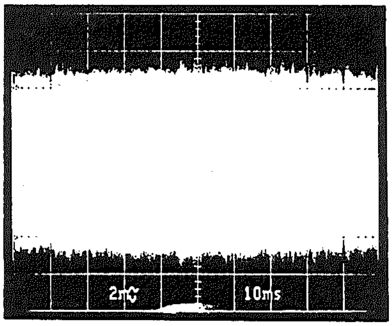


Figure 1.16

FLOATING SHIELD COUPLES ELECTROMAGNETIC ENERGY TO CENTER CONDUCTORS

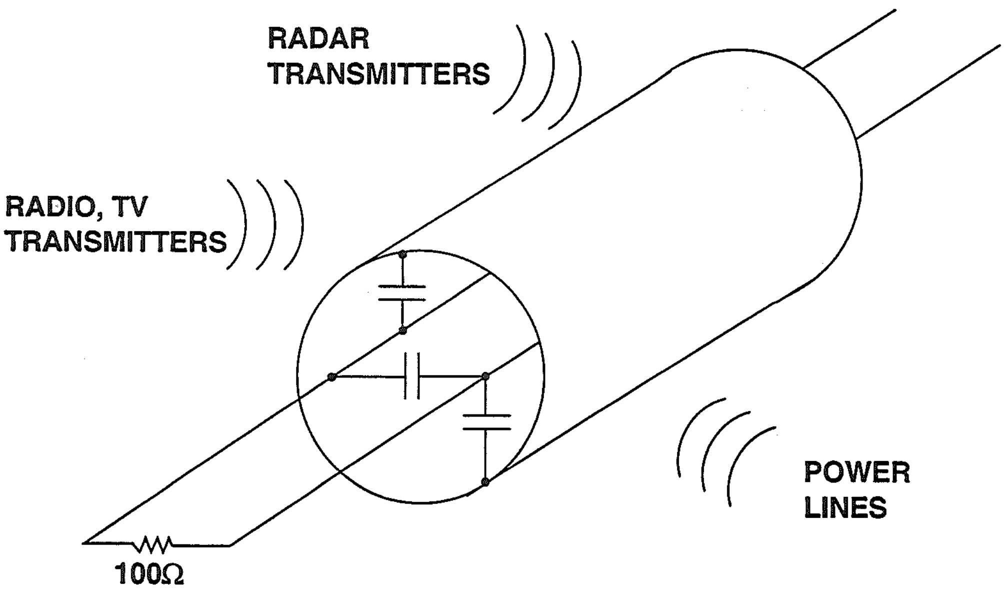
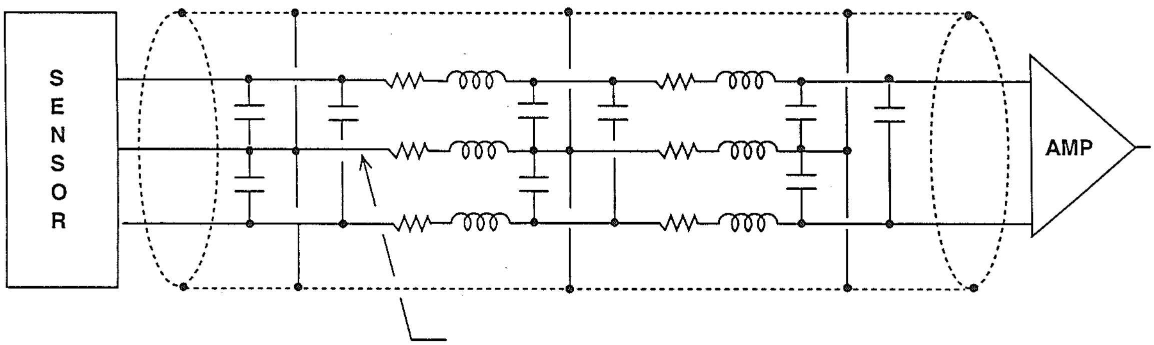


Figure 1.17

In the first experiment, the shield was left floating. A large amount of noise was recorded at the output of the amplifier (refer to Figure 1.16). Even though twisted pair wires are generally much better than non-twisted ones, floating the shield is still a poor practice. This is because the floating shield acts as an antenna, picking up radiated EM energy which is coupled to the inner conductors. Figure 1.17 shows how the shield forms a capacitance which couples the signals to the center conductors.

Figure 1.18 models the equivalent coupling capacitances that are distributed throughout the length of the cable, thereby forming a distributed network.

NOISE COUPLING IS DISTRIBUTED
ALONG THE ENTIRE LENGTH OF THE CABLE



LINE REPRESENTING THE
PARASITIC IMPEDANCES OF
THE SCREEN OF THE CABLE

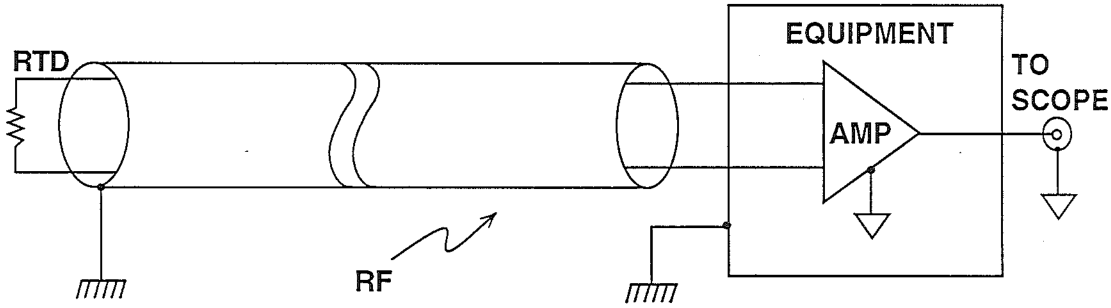
Figure 1.18

In the second experiment (shown in Figure 1.19), grounding the far end (transducer end) of the shield helps reduce noise pickup but does not remove it entirely. It does, however, cause the circuit to oscillate. Notice the small amount of high-frequency noise superimposed on the lower frequency oscillation.

Experiment 3 shows the shield grounded at the instrument end. This effectively shunts all radiated noise in the conductor to ground before it reaches the preamplifier (see Figure 1.20). Notice that although most of the high frequency noise is eliminated the oscillation remains.

EXPERIMENT 2: GROUNDING SHIELD AT FAR END

Problem: Oscillation and RF Noise Coupling at Near End



NOTE RF NOISE

VERTICAL SCALE: 200mV/div.
HORIZONTAL SCALE: 1μs/div.

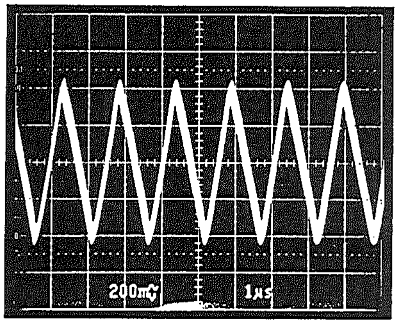
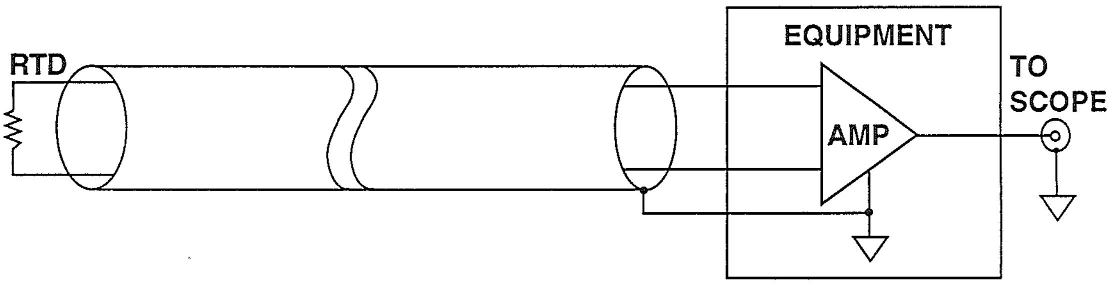


Figure 1.19

EXPERIMENT 3: SHIELD GROUNDED AT NEAR END

Result: RF Noise Eliminated, but Cable Capacitance Induces Oscillation



VERTICAL SCALE: 50mV/div.
HORIZONTAL SCALE: 200ns/div.

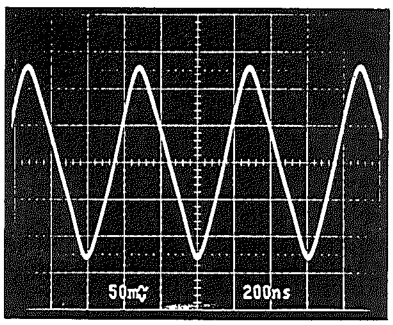


Figure 1.20

The oscillation is caused by cable capacitance which appears at the summing junction of the bridge biasing amplifier (refer to Figure 1.21) and increases the phase-lag. Consequently the amplifier has no phase margin. Placing a phase-lead compensation network around the feedback of the bridge biasing amplifier restores the phase margin of the amplifier and therefore stabilizes it. The result is lower noise and restored stability as shown in Figure 1.22.

OSCILLATION PROBLEM: CABLE CAPACITANCE DEGRADES AMPLIFIER PHASE MARGIN

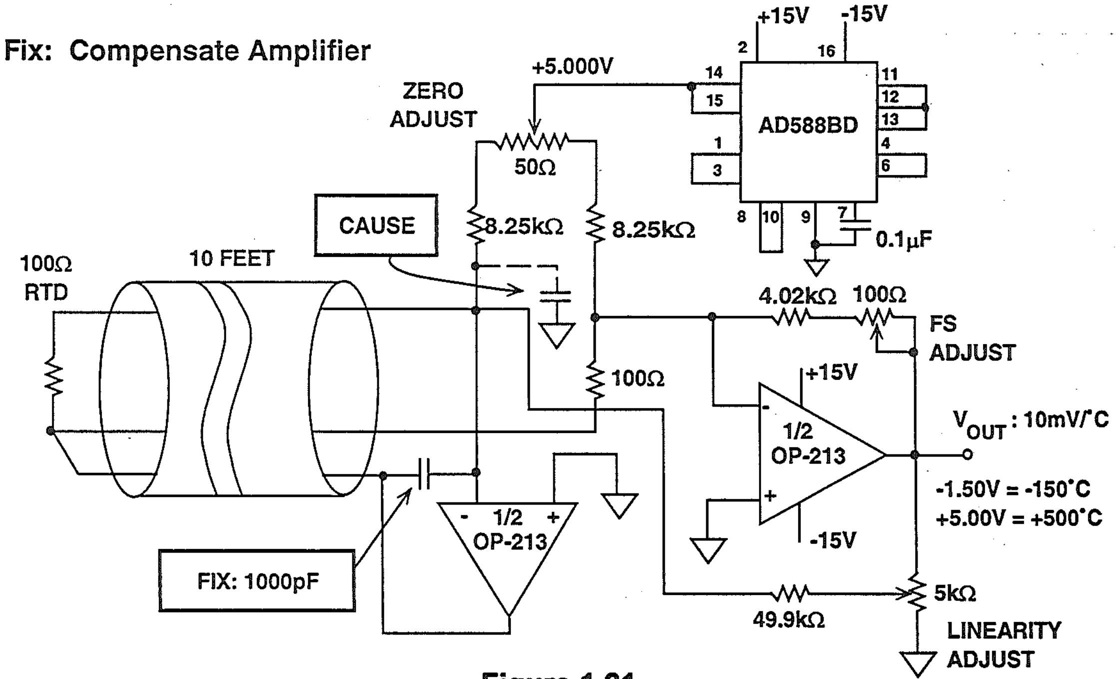
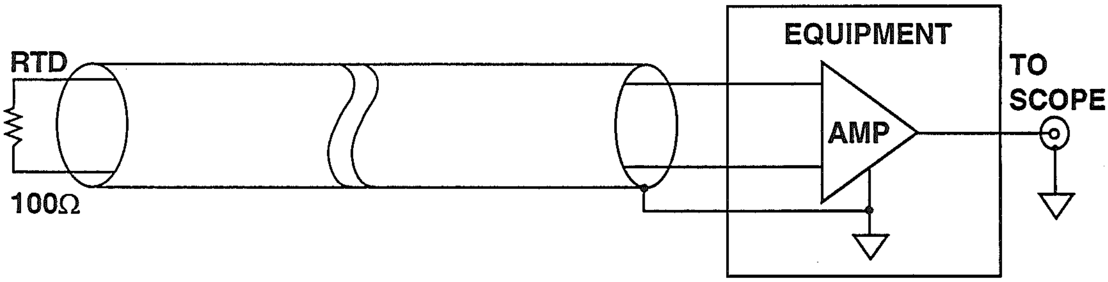


Figure 1.21

EXPERIMENT 4: GROUNDING SHIELD AT NEAR END SHUNTS RF/EMI NOISE EFFECTIVELY BEFORE REACHING AMPLIFIER



VERTICAL SCALE: 2mV/div.
HORIZONTAL SCALE: 10ms/div.

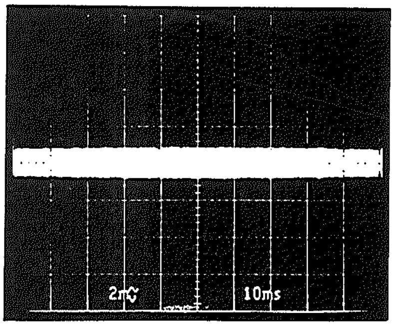
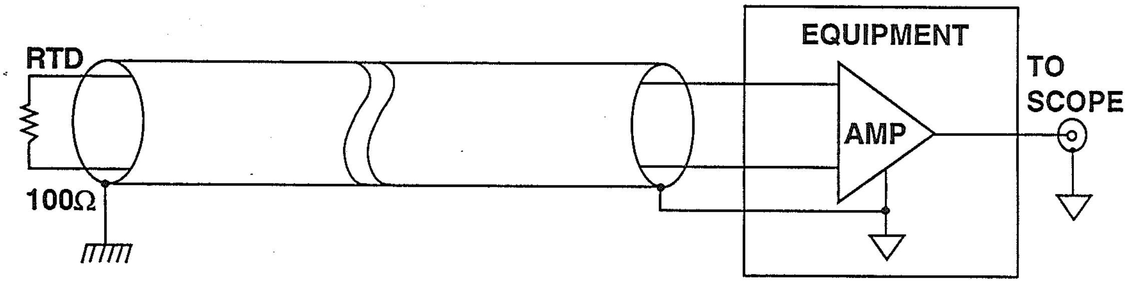


Figure 1.22

EXPERIMENT 5: AVOID DIFFERENT GROUNDS AT OPPOSITE ENDS OF THE SHIELD TO PREVENT GROUND LOOPS



VERTICAL SCALE: 2mV/div.
HORIZONTAL SCALE: 10ms/div.

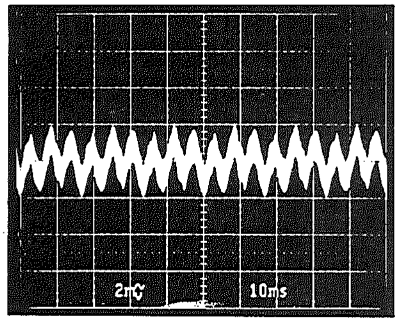


Figure 1.23

SIGNAL GROUND AND EARTH GROUND HAVE DIFFERENT POTENTIALS WHICH MAY INDUCE GROUND LOOP CURRENT

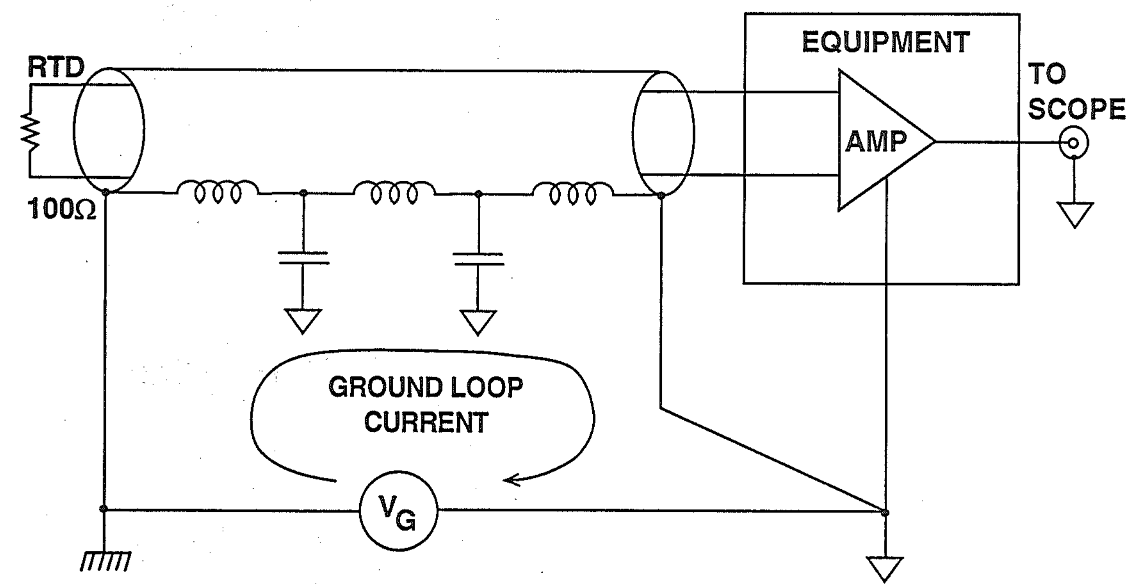


Figure 1.24

Without care in shield design, ground-loop induced noise can be a problem. For example, Figure 1.23 shows the effects of grounding opposite ends of the shield to different grounds. Ground loop current now flows in the shield which induce a noise voltage in the center conductors.

There are often significant AC and DC voltage differences between various ground points in a system's chassis. Extreme care is necessary when grounding a cable shield at both ends. Even a few millivolts difference in the two grounds may affect the DC and AC system performance.

In experiment 6, if the two ends of the shield can be connected to equi-potential (DC and AC) grounds, virtually all the noise issues are solved (refer to Figure 1.25), but such connection is rarely possible, and long wires are *not* a solution since they form loops which can pick up AC noise by induction.

In practical system designs, therefore, it is not usually practical to find equi-potential points which can be used to ground both ends of the shield. In such instances, grounding the near-end of the shield, as in Figure 1.26, is the best possibility to minimize noise pickup and EMI.

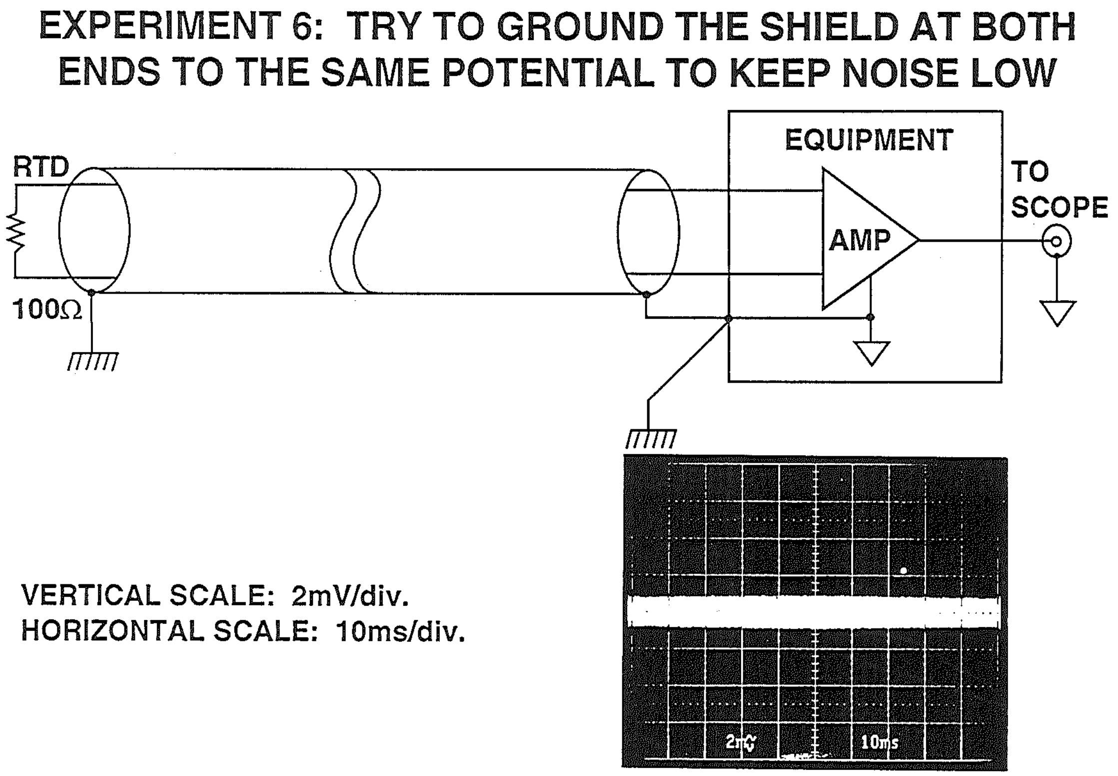
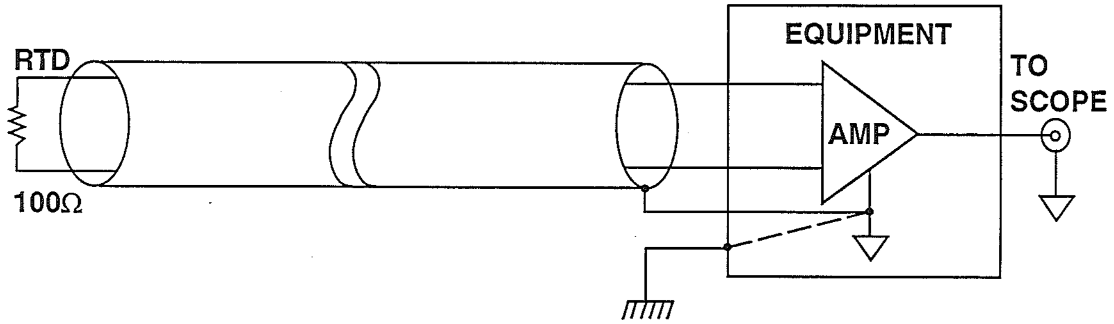


Figure 1.25

EXPERIMENT 7: WHERE EQUI-POTENTIAL GROUNDS ARE NOT POSSIBLE, GROUND SHIELD ONLY AT THE NEAR END



VERTICAL SCALE: 2mV/div.
HORIZONTAL SCALE: 10ms/div.

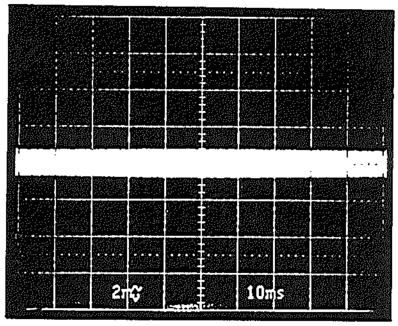


Figure 1.26

FILTERING AND PROTECTION AGAINST EMI/RFI

Circuit accuracy is affected by nearby electrical activity. This electrical activity may generate noise and is referred to as EMI (electromagnetic interference) or RFI (radio frequency interference). In this section, EMI will refer to both electromagnetic and radio frequency interference. EMI can be caused by:

1. Interference due to conduction (common-impedance),
2. Interference due to capacitive or inductive coupling (near-field interference), and
3. Electromagnetic radiation (far-field interference).

There are countless ways in which noise can couple into a sensitive circuit to ruin its accuracy. There are many types of noise sources, and some are listed in Figure 1.27.

Noise Coupling Mechanisms

RF energy may enter wherever there is an impedance mismatch or discontinuity in a system. In general this occurs at the interface where cables carrying sensitive analog signals are connected to PC boards, and through power supply leads. Improperly connected cables or poor supply filtering schemes are perfect conduits for interference. Figure 1.28 shows some of the ways noise can enter into a circuit.

FILTERING AND PROTECTION AGAINST EMI / RFI

- Noise (EMI/RFI) energy can couple into a circuit from *anywhere.*
- Sources of externally generated noise:
  - Radio and TV Broadcasts
  - Mobile Radio Communications
  - Ignition
  - Lightning
  - 50/60Hz Power Lines
  - Electric Motors
  - Computers
  - Garage Door Openers
  - Telemetry Equipment

Figure 1.27

HOW IS INTERFERENCE COUPLED?

- Impedance mismatches and discontinuities
- Common-mode impedance mismatches → Differential Signals
- Capacitively Coupled (Electric Field Interference)
  - dV/dt → Mutual Capacitance → Noise Current
  - Example: 1V/ns produces 1mA/pF
- Inductively Coupled (Magnetic Field)
  - di/dt → Mutual Inductance → Noise Voltage
  - Example: 1mA/ns produces 1mV/nH

Figure 1.28

Conducted noise may be encountered when two or more currents share a common path (impedance). This common path is often a high impedance "ground" connection. If two circuits share this path, noise currents from one will produce noise voltages in the other. In Reference 5, A. Rich outlines steps which may be taken to identify potential sources of interference.

There is a capacitance between any two conductors separated by a dielectric (air and vacuum are dielectrics, as well as all solid or liquid insulators). If there is a change of voltage on one conductor there will be change of charge on the other, and a "displacement current" will flow in the dielectric. Where capacitance or dV/dT is high, noise is easily coupled by this mechanism. (1V/ns gives rise to displacement currents of 1 mA/pF.)

If changing magnetic flux from current flowing in one circuit threads another circuit, it will induce an emf in the second circuit. Such "mutual inductance" can be a troublesome source of noise coupling from circuits with high values of dI/dT. (In a mutual inductance of 1nH, a changing current of 1A/ns will induce an emf of 1V.)

Reducing Common-Impedance Noise

Steps to be taken to eliminate or reduce noise due to the sharing of impedances are outlined in Figure 1.29.

SOLUTIONS TO EXTERNALLY-INDUCED INTERFERENCE

- Common-Impedance Noise
  - Decouple IC power supply leads at LF and HF
  - Reduce the common impedance
  - Eliminate shared paths
- Techniques
  - Tantalum electrolytic (LF) and ceramic (HF) bypass capacitors
  - Ground and Power Planes
  - Reconfigure the system design

Figure 1.29

Power supplies are an example of impedance shared among several circuits. Voltage sources may exhibit low output impedances or may not — especially over frequency. Furthermore the wiring used to distribute the power is inductive and resistive and may also form a ground loop.

The use of power and ground planes also reduces the impedance of power distribution circuits. These dedicated layers of conductors in a PCB are continuous and offer the lowest practical resistance and inductance. However, they are not perfect conductors, and there are some applications where reducing the common impedance using this method is not sufficient, and others where it is uneconomic.

Power supply impedance at individual ICs should be reduced by proper AC decoupling. A capacitor is connected between the supply pins of the IC which has low reactance at all frequencies present in the IC's supply current. Monolithic ceramic capacitors with short (and therefore low-inductance) leads are excellent for HF decoupling, but do not have sufficiently low reactance at lower frequencies, while tantalum electrolytic capacitors have excellent LF and MF performance but are somewhat inductive at HF. Every supply pin of every IC should be decoupled at HF with a 10-100 nF ceramic capacitor within 1-2 mm of the supply pin, but it may only be necessary to use one 10-22 μF tantalum capacitor for every few ICs, provided the length of conductor to it is no more than 3-5 cm. For more details on decoupling see Paul Brokaw's articles (References 4 & 7).

In some applications where low-level signals encounter high levels of common-impedance noise it will not be possible to prevent interference and the system architecture may need to be changed. Possible changes include:

1. Transmitting signals in differential form
2. Amplifying signals to higher levels
3. Converting signals into currents
4. Converting signals directly into digital form.

NOISE INDUCED BY NEAR-FIELD INTERFERENCE

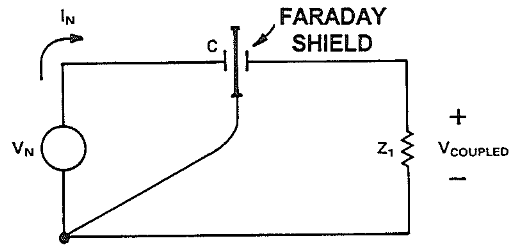
Crosstalk is the second commonest form of interference. In the vicinity of the noise source, interference is not transmitted as an electromagnetic wave, and the term "crosstalk" may apply to inductively or capacitively coupled signals.

Reducing Capacitively-Coupled Noise

Capacitively-coupled noise may be reduced by reducing the capacity (by increased separation of conductors), but is most easily cured by shielding. A grounded conducting shield (known as a "Faraday Shield") between the signal source and the affected node will eliminate capacitively-coupled noise by routing the displacement current directly to ground (Figure 1.30). It is *essential* that a Faraday Shield is grounded: a floating or open-circuit shield almost invariably increases capacitively-coupled noise.

FARADAY SHIELDING

FARADAY SHIELD INTERRUPTS THE COUPLING ELECTRIC FIELD



EQUIVALENT CIRCUIT ILLUSTRATES HOW A FARADAY SHIELD CAUSES THE NOISE CURRENTS TO RETURN TO THEIR SOURCE WITHOUT FLOWING THROUGH Z_1

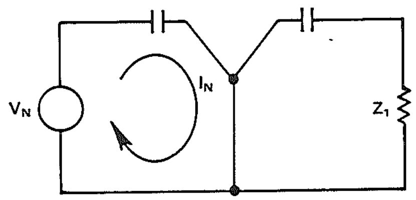


Figure 1.30

For a brief review of shielding consult the appendix, for more details consult Ott (Reference 2) and Morrison (Reference 3).

ELIMINATING CAPACITIVELY COUPLED NOISE

- Reduce Noise Sources of High dV/dt
- Proper Grounding Schemes for Cable Shields
- Reduce Stray Capacitance
  - Equalize Input Lead Lengths
  - Keep Traces Short
  - Use Signal-Ground-Signal Routing Schemes
- Use Grounded Conductive Faraday Shields to Protect Against Electric Fields

Figure 1.31

Reducing Magnetically-Coupled Noise

Methods to eliminate interference caused by magnetic fields are summarized in Figure 1.32.

ELIMINATING MAGNETICALLY COUPLED NOISE

- Careful Routing of Wiring
- Use Conductive Screens against HF Magnetic Fields
- Use High Permeability Shields (mu-Metal) for LF Magnetic Fields
- Reduce Loop Area of Receiver
  - Twisted Pairs
  - Physical Wire Placement
  - Orientation of Circuit to Interference
- Reduce Noise Source
  - Twisted Pairs
  - Driven Shields

Figure 1.32

To illustrate the effect of magnetically-coupled noise, consider a circuit with a closed-loop area A cm^2 operating in a magnetic field with an rms flux density value of B gauss. The noise voltage induced in this circuit can be expressed by the following equation:

$$V_n = 2\,\pi\, f\, B\, A \cos\theta \times 10^{-8}\ V.$$

In this equation, f represents the frequency of the magnetic field and θ represents the angle of the magnetic field B to the circuit with loop area A. The magnetic field coupling can be reduced by reducing the circuit loop area, the magnetic field intensity, or the angle of incidence. To reduce the circuit loop area requires arranging the circuit conductors closer together.

Twisting the conductors together reduces the net area of the loop, which has the effect of canceling any magnetic field pickup because the sum of positive and negative incremental loop areas is ideally equal to zero. Reducing the magnetic field directly may be difficult. However, since magnetic field intensity is inversely proportional to the cube of the distance from the source, physically moving the affected circuit away from the magnetic field has a very great effect in reducing the induced noise voltage. Finally, if the circuit is placed perpendicular to the magnetic field, pickup is minimized. If the circuit's conductors are in parallel to the magnetic field the induced noise is maximized because the angle of incidence is zero.

There are also techniques that can be used to reduce the amount of magnetic-field interference at its source. In the previous paragraph, the conductors of the receiver circuit were twisted together to cancel the induced magnetic field along the wires. If the source of the magnetic field is large currents flowing through nearby conductors, these wires can be twisted together to reduce the net magnetic field. Another technique is to use the shield of a cable as a return. If the shield current is equal to the current in the center conductor the net external magnetic field is zero.

Shields and cans are not nearly as effective against magnetic fields as against electric fields, but can be useful on occasion. At low frequencies magnetic shields using high-permeability material such a Mu-metal can provide modest attenuation of magnetic fields. At high frequencies simple conductive shields are quite effective provided that the thickness of the shield is greater than the skin depth (at the frequency involved) of the conductor used. (For copper the skin depth is $6.6/\sqrt{f}$ cm, where f is in Hz).

Passive Components: Your Arsenal Against EMI

Passive components, such as resistors, capacitors, and inductors, are powerful tools for reducing externally induced interference when used properly. Figure 1.33 summarizes the more popular low-pass filters for minimizing EMI.

USING PASSIVE COMPONENTS TO ELIMINATE EMI

LP FILTER TYPE	ADVANTAGES	DISADVANTAGE
R-C	Noise Voltage → Heat Inexpensive	Thermal Noise $I_B R$ Drop
L-C Bifilar	No Thermal Noise No IR Drop Inexpensive	e_n across L Nonlinear Core Effects
Pi (C - L - C)	Packaged Filters Low Resistance Feedthru Capacitors Low Insertion Loss High Attenuation	Expensive Nonlinear Core Effects

Figure 1.33

Simple R-C networks make efficient and inexpensive low-pass filters. The noise is converted to heat which is dissipated in the resistor. The fixed resistor does, however, produce thermal noise and, if used in the input leads of an instrumentation amplifier, can generate input-bias-current induced offset voltage. These issues affect the design of high-precision, low-noise stages.

In applications where signal and return conductors are not well-coupled magnetically, a common-mode choke can be used to increase their mutual inductance. A common-mode choke can be constructed by winding several turns of both conductors together through a high-permeability (> 2000) ferrite bead. The magnetic properties of the ferrite allow differential-mode currents to pass unimpeded while suppressing common-mode currents. Capacitors can be used before and after the choke to provide additional common-mode and differential-mode filtering, respectively. The common-mode choke is cheap and produces no thermal noise and no bias current-induced offsets. However, there is a field around the core. A metallic shield surrounding the core may be necessary to prevent coupling with other circuits. Also high-current levels should be avoided in the core as they may saturate the ferrite.

The third method for passive filtering takes the form of packaged π-networks (C-L-C). These packaged filters are completely self-contained and include feedthrough capacitors at the input and the output as well as a shield to prevent the inductor's magnetic field from radiating noise. These expensive networks offer high levels of attenuation and wide operating frequency ranges, but the filters must be selected so that for the operating current levels involved the ferrite does not saturate.

An example of shielding and filtering techniques against EMI is illustrated in Figure 1.34. In this circuit, an instrumentation amplifier is used to amplify low-level signals in the presence of high EMI. The entire circuit is enclosed in a rigid metallic shield made of copper. The layout emphasizes symmetry in the input circuit to maintain high CMR. Packaged EMI data-line and power-line filters are used to prevent EMI being carried by the conductors.

EXAMPLE DEMONSTRATING SHIELDING AND FILTERING TECHNIQUES

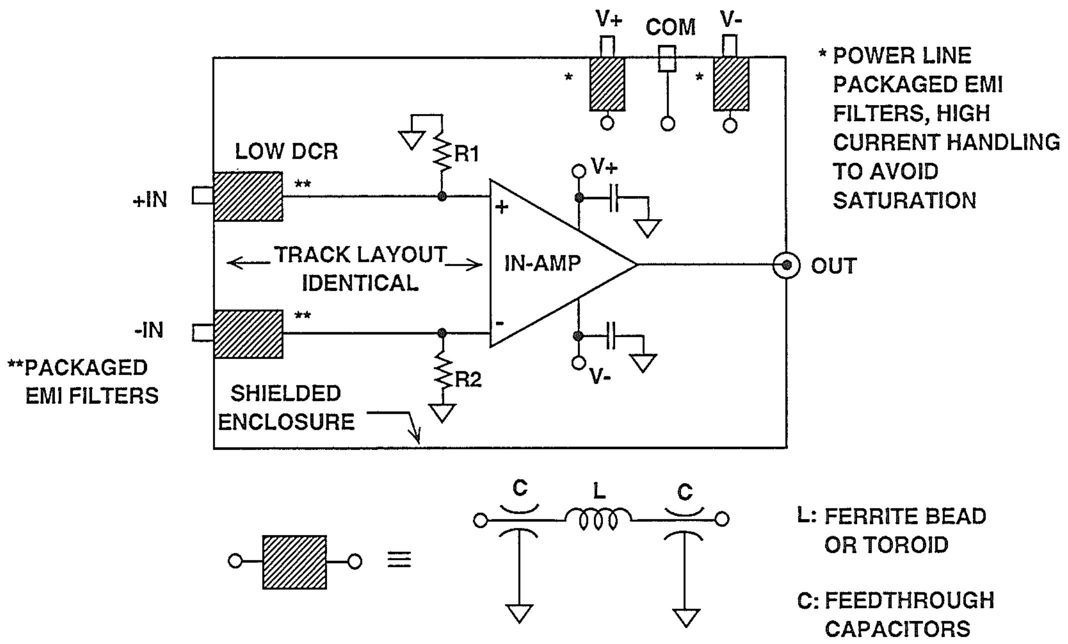


Figure 1.34

Application Circuits: EMI Reduction at Work

The circuits illustrated in Figures 1.35 through 1.39 demonstrate some of the principles outlined in this section. Each circuit use a different sensor to give the user some concrete ideas instead of theoretical remedies on how to handle interference.

The circuit illustrated in Figure 1.35 is designed to provide a 10 mV/°C output for a remotely-located Type T thermocouple. The accuracy of the circuit is ±0.4°C over a measurement temperature range of 0°C to 100°C. The circuit uses a grounded thermocouple to minimize noise pickup on long leads.

In the middle of the circuit is a low-pass filter, comprising R_N and C_N, whose function is to filter any noise along the thermocouple wires above 1.6kHz. The filter cutoff can be set lower by increasing C_N. (Although larger values for R_N might be used at the input of the OP-177, they would cause input bias-current induced offset and drift effects.) A resistor, R_P, is used in series with exposed thermocouples as protection in the event of contact with some high voltage. Otherwise, the thermocouple would be short circuited to ground and would certainly be destroyed. A capacitor connected across R_P serves to contain interference within the wires of the thermocouple cable.

USING A SIMPLE RC FILTER TO REDUCE NOISE

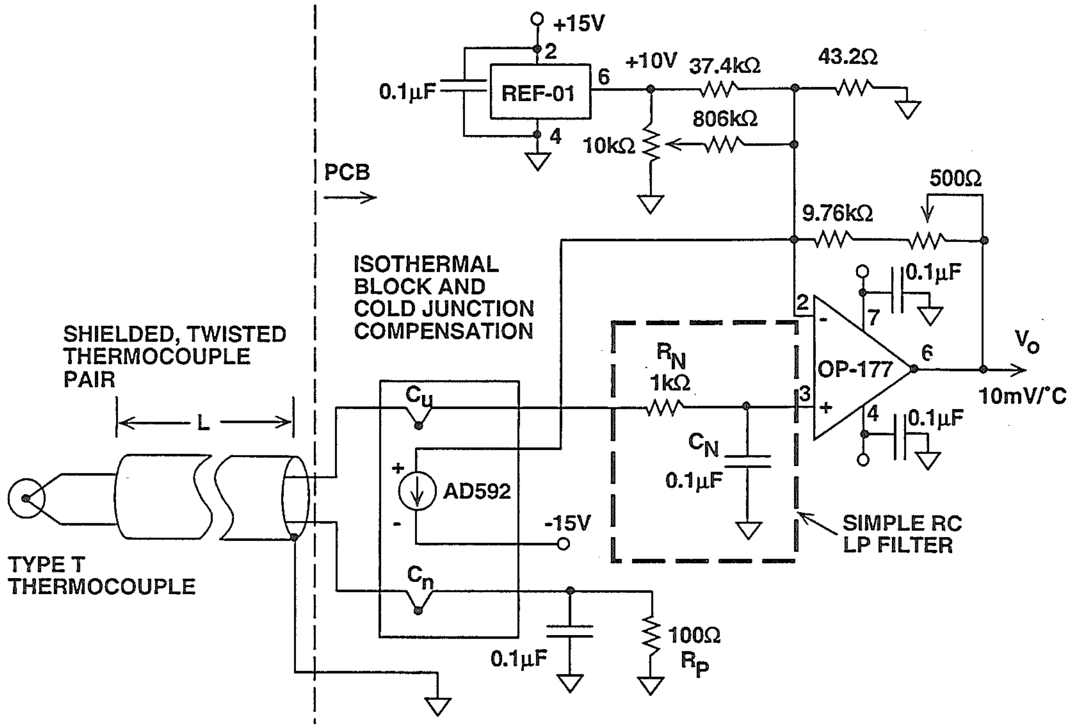


Figure 1.35

The circuit shown in Figure 1.36 (see Reference 8) illustrates a linearized thermistor in the feedback of an op amp to provide a 100mV/°C output over a temperature range of 0°C ≤ T_A ≤ 100°C. Over temperature, the linearized thermistor network ranges from 2.7kΩ at 0°C to 1 kΩ at 100°C. A 1μF capacitor is connected across the thermistor network to form a low-pass filter with cutoff frequencies between 58Hz and 150Hz. (The cutoff frequency moves as the thermistor resistance changes over temperature.) Another technique to avoid EMI is to limit the signal bandwidth. In this circuit, C2 is used across R4 and P2 and sets a low-pass cutoff at 8.5 Hz.

Figure 1.37 shows a linearized RTD amplifier that is powered by a single +5V supply. The RTD is excited by a 100μA constant current which is regulated by A1. The 0.202V reference used to generate the constant current is provided by the REF-43 and resistors R4 and R5. The AMP-04 is scaled to give 10mV/°C output. When properly calibrated, the circuit achieves better than ±0.5°C accuracy within a temperature measurement range of 0°C to 400°C. Capacitor C3 performs two functions: it stabilizes the loop around A1 and provides a low-pass cutoff frequency of 800kHz. C1 works directly with the RTD and R3 to form a low-pass filter at 8kHz to prevent any noise from appearing at the inputs of the AMP-04.

USING CAPACITORS TO FILTER NOISE AND LIMIT BANDWIDTH

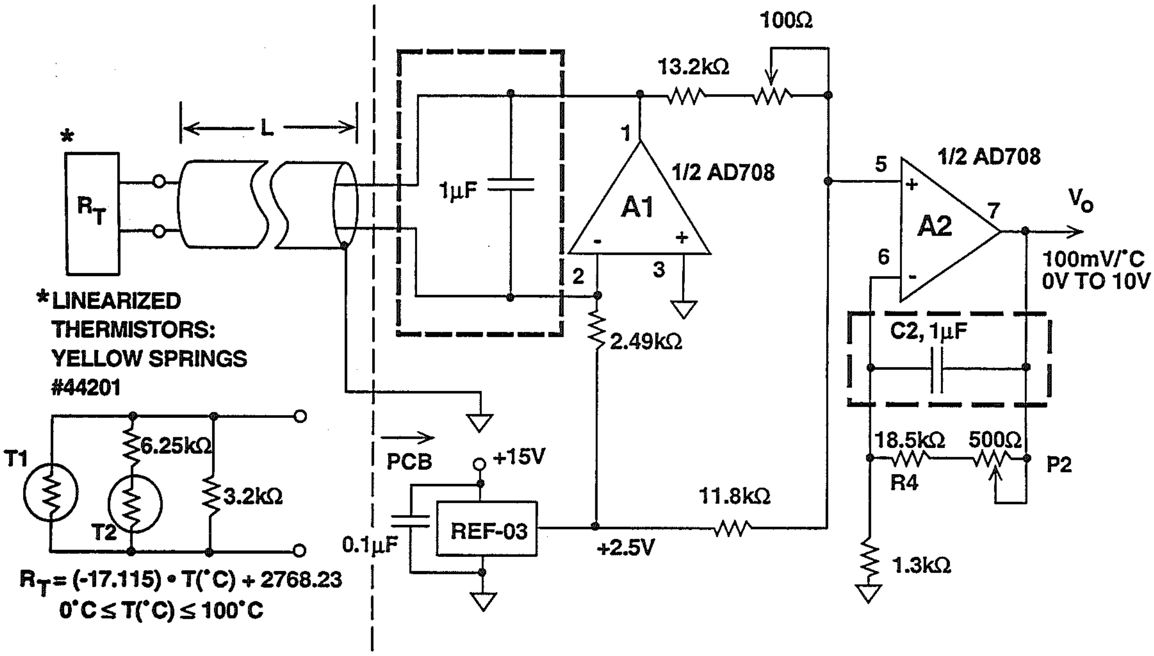


Figure 1.36

USING A CAPACITOR WITH AN RTD TO FILTER THE INSTRUMENTATION AMPLIFIER OUTPUT

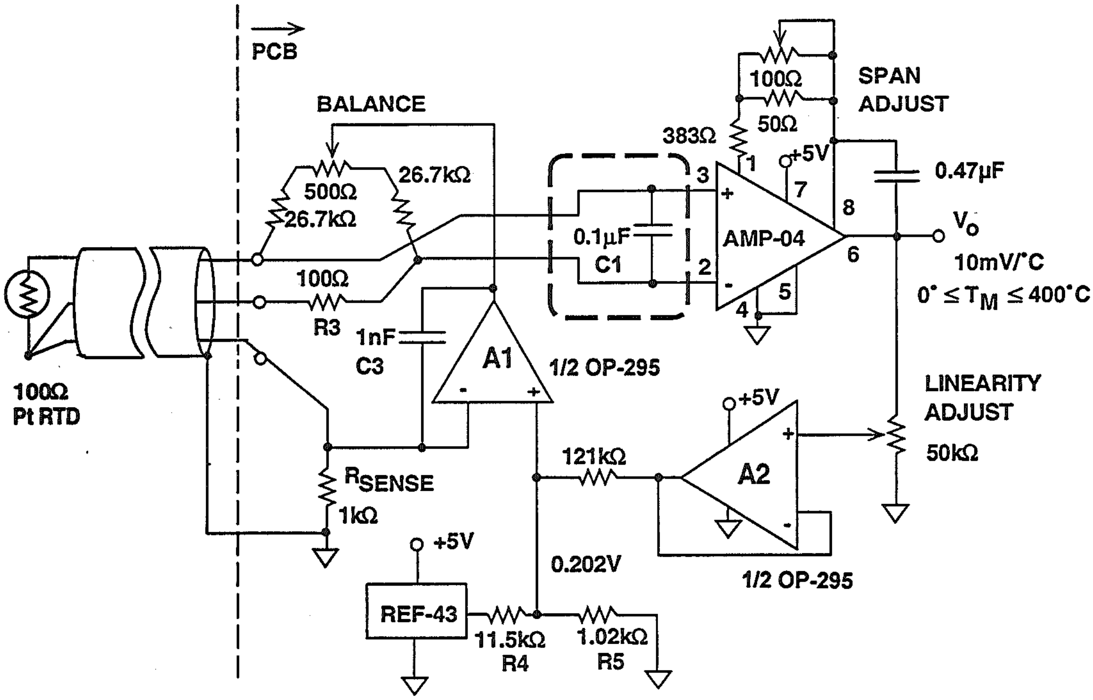


Figure 1.37

In the example illustrated in Figure 1.38 (Reference 9), a two-terminal, monolithic temperature sensor, the AD592, is used with an AD624 instrumentation amplifier to measure temperatures from –25°C to 100°C. The output voltage of the circuit is linearly scaled to register 100mV/°C. With any remotely-located sensor, noise pickup along the cables and rectification in the sensor and the amplifier can be a problem. In this circuit, resistors and capacitors are used to confine any conducted or radiated interference to the boundaries of the shielded cable. C1, which is connected across the AD590, prevents any interference from being rectified by the AD592 into a DC offset that would affect accuracy. This capacitor forms a low-pass corner at 38kHz with the resistors R1 through R5. C2 forms a low-pass cutoff frequency at 106kHz to divert any interference along the cable that gets by C3. Finally, C3 set a low-pass cutoff at 480Hz with R5 to prevent interference from reaching the inputs of the AD624.

USING RESISTORS AND CAPACITORS TO CONFINE INTERFERENCE INSIDE A LONG CABLE

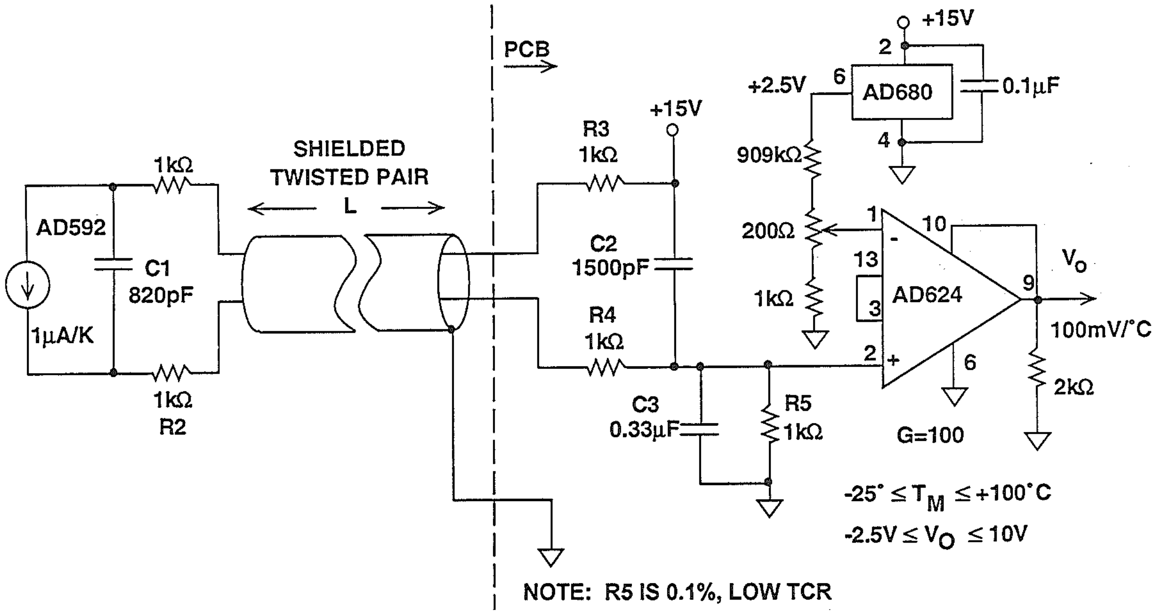


Figure 1.38

Strain or flexure can be remotely sensed with the circuit in Figure 1.39. A 1kΩ strain gauge and an AD620 are used to provide a linear 1V/1000με output. For remote sensing, current excitation is used where the OP-177 drives the bridge with 10mA derived from a reference voltage of 1.235V. The strain gauge has an output of 10.25mV per 1000με. Full-scale strain voltage may be set by adjusting the gain potentiometer so that for a strain of –3500με, the output reads –3.500V; and for a stain of 5000με, the output registers a +5.0000V output. To prevent any interference from reaching the inputs of the AD620, a capacitor is placed directly across them. This capacitor sets a low-pass corner at 1.6kHz in conjunction with the strain gauge's resistance.

USING A CAPACITOR AND A STRAIN GAUGE'S RESISTANCE TO FILTER NOISE BEFORE AMPLIFICATION

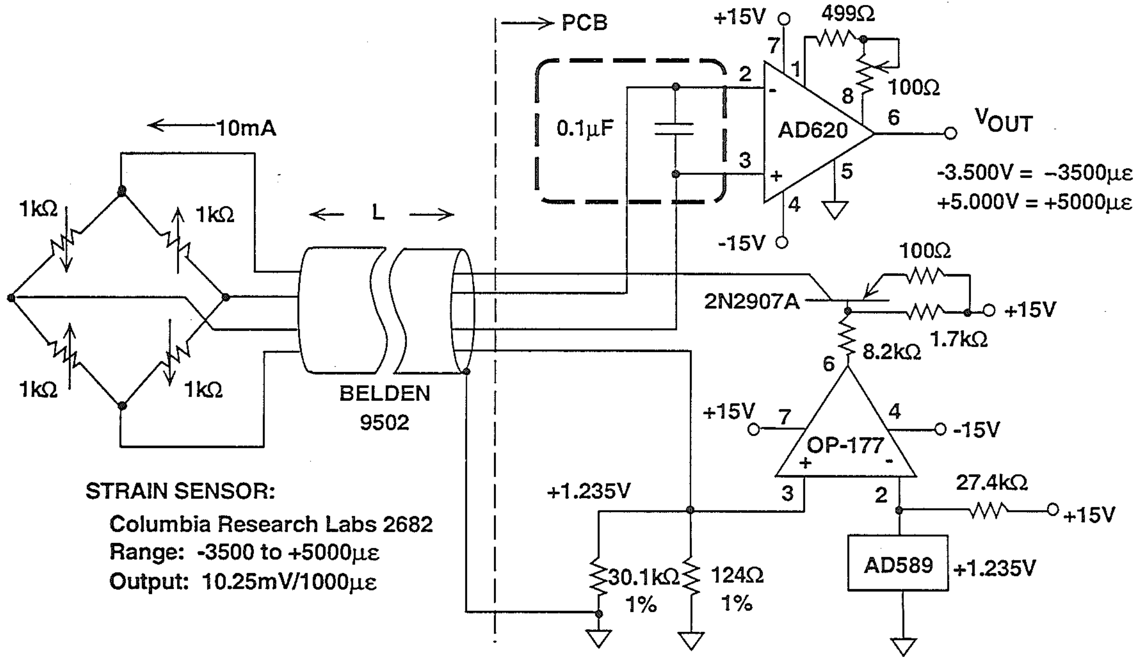


Figure 1.39

Reducing Susceptibility to EMI

The examples shown above and the techniques illustrated earlier in this section outline the procedures that can be used to reduce or eliminate electromagnetic interference. A summary of possible measures is given in Figures 1.40 and 1.41.

REDUCING SYSTEM SUSCEPTIBILITY TO EMI

- Always assume that interference exists!
- Use conducting enclosures against electric and HF magnetic fields
- Use mu-Metal enclosures against LF magnetic fields
- Implement cable shields effectively
- Use feedthru capacitors and packaged pi-filters

Figure 1.40

REDUCING CIRCUIT SUSCEPTIBILITY TO EMI

- Reduce or eliminate common impedances
- Use HF and LF power supply decoupling
- Use ferrite beads, resistors, capacitors
- Balance the layout for high AC CMR
- Limit system bandwidth to minimize noise

Figure 1.41

Appendix: A Review of Shielding

A lump of metal between an interference source and a sensitive circuit will reduce interference. But why? More detailed (and more mathematical) discussions will be found in References 1 through 11, but this appendix is intended to highlight some basic principles for the engineer who is troubled by interference and thinks a screen might help.

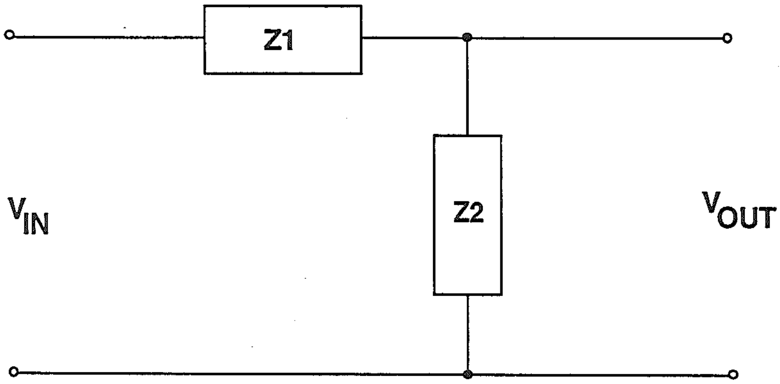
We have seen in the main text that interference can be conducted into a sensitive circuit and that this effect can be minimized by appropriate filters. We have also seen that interference can be coupled by electric, magnetic and electro-magnetic fields. Shields attenuate these fields and may be considered as "field attenuators".

When the interference source is close to its victim ("near field" conditions) the field will be electric or magnetic, depending on circumstances, and when it is far away ("far field" conditions) it will be electro-magnetic ("far away" in these circumstances is usually defined as $\geq 0.16\lambda$ where λ is the wavelength of the interference, or $\geq c/2\pi f$, where f is the frequency of the interference and c is the velocity of light).

The characteristic impedance of free space (the far field case) is 377Ω and the field strength is proportional to the inverse square of the distance from its source. In the near field the interference source may produce a predominantly magnetic field, which is generally "low impedance" (this is not a rigorous definition but is sufficient to enable one to understand the action of conducting screens), or may produce an electric field, which is "high impedance". Both of these fields are "dipole fields" which have the property of being proportional to the inverse *cube* of the distance from their source, so it is evident that simply increasing the separation between an interference source and its victim may avoid the necessity for a screen. This is a powerful argument against the trend to minimize the size of all electronic assemblies - a small increase in the size of a noise-critical PCB can often have a major beneficial effect on its overall noise performance.

The simplest attenuator consists of two impedances, Z1 and Z2, arranged as in Figure 1.42. The larger the ratio Z1:Z2, the greater the attenuation. A shield is an attenuator of this type: Z1 is the "source impedance" of the field and Z2 the impedance of the shield.

A SHIELD IS AN ATTENUATOR



$$\frac{V_{OUT}}{V_{IN}} = \frac{Z2}{Z1 + Z2}$$

Figure 1.42

Where electric fields are involved the impedance of a shield consisting of a piece of grounded conductor (a "Faraday Shield" or a "Faraday Cage") is low, and the source impedance of the electric field is high. It follows that the attenuation, and hence the efficiency of shielding, is high in this case. Faraday shields are almost invariably very efficient at eliminating the interference effects due to electric fields - but they must be connected to ground or some other low impedance, or they will increase, rather than attenuate, the noise coupling. Failure to ground Faraday shields is one of the commonest engineering oversights leading to poor interference protection.

Magnetic fields have low "source impedance" and shields of high-permitivity magnetic material, such a mu-metal, have relatively large impedance to magnetic fields. Magnetic shields are therefore much less efficient at suppressing low frequency magnetic field interference than Faraday shields are at limiting electric fields (most magnetic shields are conductors, however, and, if grounded, will act as very satisfactory Faraday shields, whatever their deficiencies as magnetic shields).

At high frequencies Faraday shields are quite effective at attenuating magnetic fields as well as electric fields. This is due to the "skin effect": a high frequency magnetic field induces eddy currents near the surface of a conductor which prevent it from penetrating to any great depth. HF currents in a conductor, accordingly, flow only in the thin "skin layer" near its surface. Therefore, if a grounded conductor is much thicker than its skin depth at a particular frequency it will act as an efficient attenuator to magnetic, electric and electro-magnetic fields at that frequency. Skin depth is a function of bulk resistivity, magnetic permeability and frequency: for copper it is $6.6/\sqrt{f}$ cm.

It is rarely necessary to consider far field effects for low frequency interference since we will rarely be concerned with sources more than a hundred meters away and at any frequency below 500 kHz a hundred meters is "near field"! The skin depth in copper for 500 kHz signals is about 0.1 mm (0.004") so a copper can with a wall more than 0.25 mm thick will be quite an efficient shield at this frequency.

Electromagnetic-radiation can enter a shield through any hole larger than about 0.1λ, so at very high frequencies hole sizes must be considered. At any frequency a conductor entering a shielded volume can admit interference if it is not grounded to the shield, so filters must be used where there is any danger of this.

To summarize: Faraday (conductive) shields are very effective against electric fields and high frequency fields of all types, provided they are grounded and not left floating, but high permeability magnetic shields are less efficient barriers to low frequency magnetic fields, although they can be useful in some circumstances. Physical separation always improves noise isolation.

INPUT-STAGE RFI RECTIFICATION SENSITIVITY

A well-known, but poorly understood phenomenon in analog integrated circuits is RF rectification, specifically in instrumentation amplifiers and operational amplifiers. While amplifying very small signals these devices can rectify unwanted high frequency signals. The result is dc errors which appear at the output in addition to the wanted sensor signal. Unwanted out-of-band signals enter sensitive circuits through the circuit's conductors. These conductors, which lead into and out of the circuit, provide a direct path for interference to couple into a circuit. These conductors pick up noise through capacitive, inductive, or radiation coupling as discussed elsewhere. Regardless of the type of interference, the unwanted signal is a voltage which appears in series with the inputs. How devices rectify signals and methods for preventing RFI rectification will be discussed in this section.

All instrumentation and operational amplifier input stages comprise emitter-coupled (BJT) or source-coupled (FET) differential pairs with resistive or current-source loading. Depending on the quiescent current level in the devices and the frequency of the interference, these differential pairs can behave as high-frequency detectors. As will be shown, this detection process produces spectral components at the harmonics of the interference as well at dc! It is the dc component that shifts the bias levels of to produce errors, which can lead to system inaccuracies.

WHAT IS RFI RECTIFICATION?

- **Source:** Out-of-Band RF Interference, Conducted or Radiated
- **Mechanisms:** Input Differential Pairs - HF Detectors

  Input and Gain Stages of Instrumentation and Operational Amplifiers
- **Result:** DC Offset Errors → System Inaccuracies

Figure 1.43

The effect of rectification on instrumentation and operational amplifiers may be evaluated with the circuits in Figure 1.44. Amplifiers are configured for a gain of 100, and the outputs connected to a 100-Hz low-pass filter to prevent other unwanted signals from interfering with the measurement. A $20mV_{p-p}$ signal at 100MHz is the stimulus and was chosen to be well above the frequency limits of the circuits under test. Outputs are monitored with a 300MHz oscilloscope and measurements made with a digital voltmeter. Five instrumentation amplifiers, and 17 single and 12 dual operational amplifiers have been tested. Devices of both bipolar and FET technologies were evaluated to determine any patterns in RFI rectification.

Based on the data which is summarized in Figures 1.46, 1.47, and 1.48, two patterns appeared. First, device susceptibility to rectification appears to be inversely proportional to supply current; that is, devices biased at low quiescent supply currents exhibited the largest amount of output voltage shift. Second, ICs with JFET-input differential pairs appeared to be less susceptible to rectification than BJT input stages.

RFI RECTIFICATION TEST CONFIGURATION

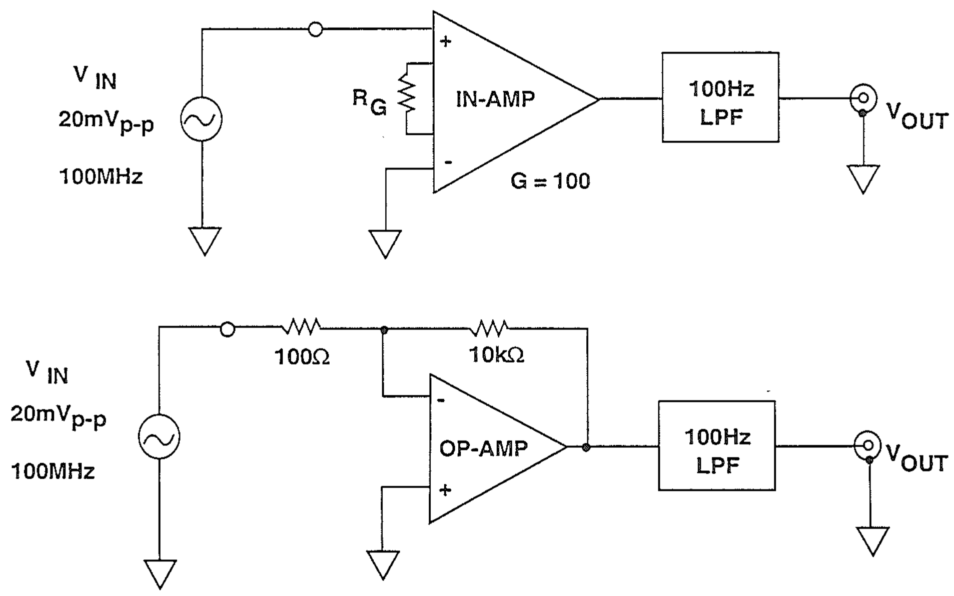


Figure 1.44

OBSERVATIONS

- Correlation: Rectification sensitivity to supply current
- JFET-Input Devices: Lower sensitivity to rectification

Figure 1.45

INSTRUMENTATION AMPLIFIERS

DEVICE	I_{SY}	ΔV_O
AMP-04	0.9mA	22mV
AD620	1.3mA	6mV
AMP-02	5mA	0.65mV
AD625	5mA	21.5mV
AMP-01	5mA	0.5mV

Figure 1.46

In Figure 1.46, all five instrumentation amplifiers use BJTs in their input stages. The AMP-04, biased at 900µA quiescent supply current, exhibited the largest output voltage shift of the group. The AMP-01, biased at a quiescent supply current of 5 mA, exhibited the least amount of error. The AD625, although biased at a supply current of 5mA, exhibited a larger output shift when compared to the AMP-01. One possible explanation for this large discrepancy could be that super-β transistors (used in the input stage of the AMP-01) exhibit a lower susceptibility to rectification than standard-process BJTs.

The results for single operational amplifiers are summarized in Figure 1.47. A clear pattern in the data is that FET-input operational amplifiers consistently exhibit a lower susceptibility than BJT-input operational amplifiers. Some FET-input devices, such as the OP-80, the OP-42, and the AD845, exhibited no observable shift in their output voltages. Of the bipolar-input operational amplifiers, the amount of output shift decreased with increasing supply current. Of this group, only the AD797 showed no observable output voltage shift.

Results of the dual operational amplifiers are illustrated in Figure 1.48. Again, the pattern is repeated in that the FET-input devices exhibited little or no observable shift in their output voltages, regardless of quiescent supply current. The OP-200 and OP-297 both exhibited very little output voltage shift. This may be due to their use of in super-β transistors *and* input bias current cancellation.

OPERATIONAL AMPLIFIERS -- SINGLES

DEVICE	I_{SY}	ΔV_O
OP-90	0.03mA	4mV
OP-20	0.095mA	58mV
OP-80 (CMOS)	0.33mA	N/C
AD705 (Super-β)	0.6mA	10mV
OP-97 (Super-β)	0.6mA	13mV
AD795	1.5mA	2mV
OP-177	2mA	2.5mV
AD707	3mA	2mV

DEVICE	I_{SY}	ΔV_O
AD711 (FET)	3.4mA	0.6mV
AD645 (FET)	3.5mA	0.3mV
AD744 (FET)	5mA	0.5mV
OP-27	6mA	2mV
OP-42 (FET)	6.5mA	N/C
AD829	7mA	3mV
AD797	9.5mA	N/C
AD743 (FET)	10mA	0.1mV
AD845 (FET)	12mA	N/C

N/C = No Change

Figure 1.47

OPERATIONAL AMPLIFIERS -- DUALS

DEVICE	I_{SY}	ΔV_O
OP-290	0.02mA	4mV
OP-295	0.175mA	3mV
OP-282 (FET)	0.25mA	1mV
OP-297 (Super-β)	0.63mA	0.14mV
OP-200 (Super-β)	0.73mA	0.19mV
OP-213	2mA	6mV
OP-275	2.5mA	6mV
AD708	2.8mA	2mV
AD712 (FET)	3.4mA	0.3mV
OP-249 (FET)	3.5mA	N/C
AD746 (FET)	5mA	0.3mV
AD827	7mA	N/C

N/C = No Change

Figure 1.48

Based on these data and from the fundamental differences between BJTs and FETs, we can summarize what we know. Bipolar transistor action is controlled by a forward-biased p-n junction (the base-emitter junction) whose I-V characteristic is exponential and quite nonlinear. FET behavior, on the other hand, is controlled by voltages applied to a reverse-biased p-n junction diode (the gate-source junction). The I-V characteristic of FETs is a square-law, and, thus more linear than that of BJTs.

In the case of low current devices, transistors in the circuit are biased below their peak f_T collector currents. Although the devices are constructed on processes whose device f_Ts can operate at hundreds of MHz, charge transit times through the devices increase at low quiescent collector currents. RF rectification in these devices is also exacerbated by the impedance levels used. In low-power operational amplifiers, impedance levels are on the order of hundreds to thousands of kΩs whereas in moderate supply-current designs the impedance levels might be no more than a few kΩ. These factors combine to affect a device's sensitivity to RF rectification.

OBSERVATIONS

- Devices with bipolar input transistors do rectify
  - Forward-Biased Base-Emitter Junctions
  - Exponential I-V Transfer Characteristic
- Devices with FET input transistors are less sensitive
  - Reverse-Biased p-n Junctions
  - Square Law I-V Transfer Characteristic
- Low I_{sy} Devices versus High I_{sy} Devices
  - Low I_{sy} → Higher Sensitivity
  - Low I_C (Q1, Q2) → Longer Transit Times
  - Low I_{sy} → Higher Internal Z Levels

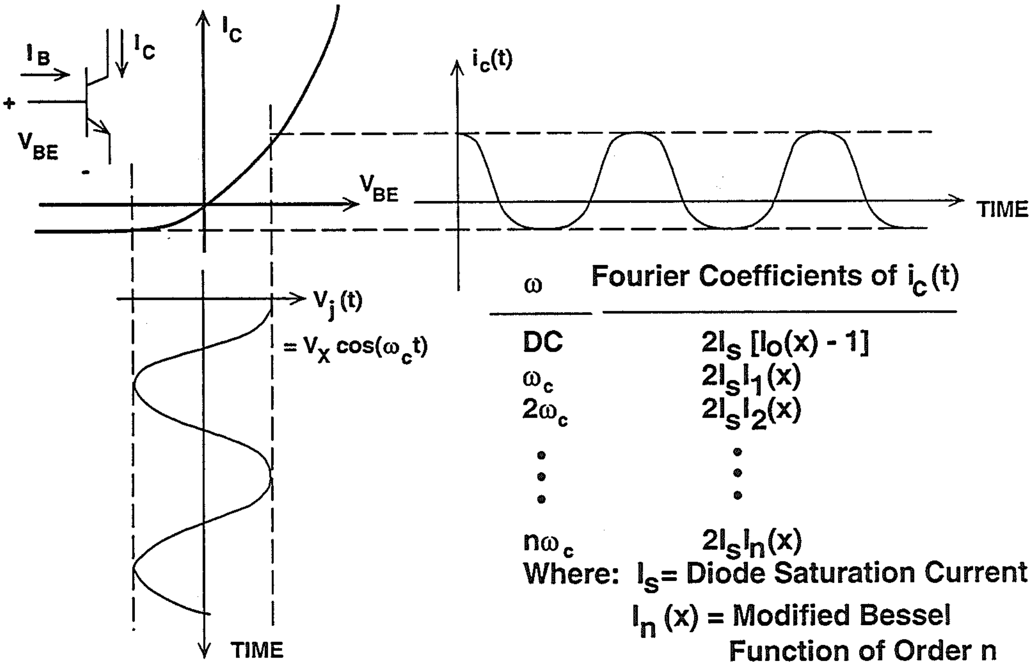
Figure 1.49

RFI Rectification: An Analytical Approach

These experiments have demonstrated that BJT-input devices exhibit a greater amount of output voltage shift on exposure to RFI than devices with FET inputs. In this section, an analytical approach will be taken to explain the phenomenon.

RF circuit designers have long known that p-n junction diodes are extremely efficient rectifiers because of their nonlinear I-V characteristics. Figure 1.50 illustrates the base-emitter junction diode I-V characteristic of an NPN bipolar transistor whose collector current, $i_C(t)$, as a function of the applied voltage, $V_j(t)$, is superimposed. The spectral components of the Fourier analysis of the current output for an HF sinewave input reveals that, as the device is biased closer to its "knee," device nonlinearity increases. This, in turn, makes the detection process more efficient. This is especially true in low-power operational amplifiers where the transistor is biased at very low collector currents.

HF MODULATION OF THE BASE-EMITTER JUNCTION CAUSES RECTIFICATION



ω	Fourier Coefficients of $i_C(t)$
DC	$2I_S\,[I_0(x) - 1]$
ω_c	$2I_S I_1(x)$
$2\omega_c$	$2I_S I_2(x)$
⋮	⋮
$n\omega_c$	$2I_S I_n(x)$

Where: I_S = Diode Saturation Current
$I_n(x)$ = Modified Bessel Function of Order n
$x = V_X/V_T$

Figure 1.50

Rectification analysis of an NPN transistor's collector current is shown in Figure 1.51. The analysis begins with the transistor's I_C–V_{BE} characteristic equation. If a small voltage, given by $\Delta V = V_X \cos(\omega_c t)$, is applied to the device, then a Taylor series expansion of the collector current around the operating point (I_C @ V_{BE}) produces three dominant spectral components (terms above the second order are neglected). These spectral components are the quiescent collector current, I_C, a linear term at the input frequency, $\cos(\omega_c t)$, and a quadratic term at the input frequency, $\cos^2(\omega_c t)$. The linear spectral component is either filtered (hopefully) or rectified by other gain stages within the device. It is the quadratic term that contains the primary rectified information.

RECTIFICATION ANALYSIS CONTINUED

- Start $\rightarrow$ $i_c = I_s\left[e^{qV_j/kT} - 1\right]$
- For a small applied voltage, ΔV, a Taylor Series Expansion

$$i_c = i_c(V_{BE} + \Delta V)$$

$$= I_C(V_{BE}) + \Delta V \cdot \left.\frac{di}{dv}\right|_{I_C} + \frac{\Delta V^2}{2!}\cdot\left.\frac{d^2i}{dv^2}\right|_{I_C} + \cdots + \frac{\Delta V^n}{n!}\cdot\left.\frac{d^ni}{dv^n}\right|_{I_C}$$

Where V_{BE} = dc bias voltage
I_C = dc or average bias current
ΔV = ac voltage across base-emitter junction

- If $\Delta V = V_X\cos(\omega_C t)$, then

$$i_c = I_c(V_{BE}) + V_X\cos(\omega_c t)\left.\frac{di}{dv}\right|_{I_C} + \frac{V_X^{\ 2}}{2}\cos^2(\omega_C t)\left.\frac{d^2i}{dv^2}\right|_{I_C}$$

- Thus, a second-order rectified term appears:

$$\Delta i_c = \frac{V_X^{\ 2}}{4}\left[1 + \cos(2\omega_c t)\right]\left.\frac{d^2i}{dv^2}\right|_{I_C}$$

Figure 1.51

The second-order term can be simplified by using the following trigonometric identity:

$$\cos^2(\omega_c t) = \frac{1}{2}\left[1 + \cos(2\omega_c t)\right] \qquad \text{Eq. 1.1}$$

Substituting the right-hand side of Eq. 1.1 into the expression for the second-order term yields:

$$\Delta i_c = \frac{V_X^{\ 2}}{4}\left[1 + \cos(2\omega_c t)\right]\left.\frac{d^2i}{dv^2}\right|_{I_C} \qquad \text{Eq. 1.2}$$

Eq. 1.2 reveals that the original quadratic second-order term can be simplified into a frequency-dependent term, $\Delta i_C(AC)$, at twice the input frequency and a dc term, $\Delta i_C(DC)$, as shown in Figure 1.52. The dc term can be further simplified by evaluating the second derivative of the collector current with respect to the base-emitter voltage at the quiescent collector current. The result of this operation yields the final form for the rectified DC term:

$$\Delta i_C(DC) = \left(\frac{V_X}{V_T}\right)^2 \bullet \frac{I_C}{4} \qquad \text{Eq. 1.3}$$

THE CHANGE IN A BJT's COLLECTOR CURRENT HAS TWO COMPONENTS

- AC Component @ $2\omega_C$:

$$\Delta I_C(AC) = \frac{V_x^2}{4}\cos(2\omega_C t)\left.\frac{d^2 i}{dv^2}\right|_{I_C}$$

- DC Component:

$$\Delta I_C(DC) = \frac{V_x^2}{4}\cdot\left.\frac{d^2 i}{dv^2}\right|_{I_C}$$

$$= \frac{V_X^2}{4}\cdot\frac{I_S \exp[V_{BE}/V_T]}{V_T^2}$$

$$\Delta i_C(DC) = \left(\frac{V_X}{V_T}\right)^2 \cdot \frac{I_C}{4}$$

- DC Component is Operating Point Dependent!

Figure 1.52

The result in Eq. 1.3 shows that the dc component of the second-order term is directly proportional to the square of the noise amplitude of the HF noise and, more importantly, to the quiescent collector current of the transistor. To illustrate the effect of transistor operating point on rectification, Figure 1.53 shows the change in the dc collector current of an NPN transistor when a $20mV_{p\text{-}p}$, high-frequency signal impinges upon it.

RFI RECTIFICATION VERSUS BJT OPERATING POINT T_A = 25°C

Quiescent Collector Current, I_C	Rectified Collector Current, Δi_C
1 µA	38 nA
10 µA	380 nA
100 µA	3.8 µA
1 mA	38 µA

Figure 1.53

To reduce the amount of rectified collector current is therefore a matter of reducing the quiescent current or the magnitude of the interference. Since the input stages of instrumentation and operational amplifiers do not often provide access for adjusting quiescent collector currents, reducing the level of noise is the best policy. For example, reducing the amplitude of the interference by 50% produces a 75% reduction in the rectified collector current.

A similar approach can be used for the rectification analysis of a FET's drain current as a function of a small HF voltage applied to its gate. The results of evaluating the second-order rectified term for the FET's drain current are summarized in Figure 1.54. Like the BJT, the FET's second-order term has an ac and a dc component. The simplified expression for the dc term of the rectified drain current is given in Eq. 1.4, where the rectified dc drain current is directly proportional to the square of the amplitude of the offending signal. However, Eq. 1.4 reveals an important difference between the amount of rectification produced by BJTs and FETs. Whereas in a BJT the change in collector current has a direct relationship to its quiescent collector current level, the change in a FET's drain current is proportional to its drain current at zero gate-source voltage, I_{DSS}, and inversely proportional to the square of its channel pinch-off voltage, V_P — parameters that are geometry- and process-dependent. Typically, FETs used in the input stages of instrumentation and operational amplifiers are biased such that their quiescent drain current is approximately $0.5 \bullet I_{DSS}$. Therefore, the change in a FET's drain current is independent of its quiescent drain current; hence, independent of the operating point.

$$\Delta i_D(DC) = \left(\frac{Vx}{Vp}\right)^2 \bullet \frac{I_{DSS}}{2} \qquad \text{Eq. 1.4}$$

A quantitative comparison of second-order rectified dc terms between BJTs and FETs is illustrated in Figure 1.55. In this example, a bipolar transistor with a unit emitter area of $576\mu m^2$ is compared to a unit-area JFET designed for an I_{DSS} of $20\mu A$ and a channel pinch-off voltage of 2V. Each device is biased at $10\mu A$ and operated at T_A = 25°C. Using these parameters, Eq. 1.3 and Eq. 1.4 yield the result that, under identical quiescent current levels, the change in collector current in bipolar transistors is ≈1500 times greater than the change in a JFET's drain current.

Although the rectified dc term for BJTs and FETs is directly proportional to either the BJT's quiescent collector current or the FET's I_{DSS}, no access to the internal circuitry is provided to adjust the operating point of the differential pair at the input of IC op amps and instrumentation amplifiers. However, the analysis showed that, regardless of the amplifier type used, RFI rectification is directly proportional to the *square of the magnitude* of the applied interference. Therefore, to minimize rectification the interference should be reduced or eliminated. The most direct way to reduce or eliminate unwanted noise is through filtering.

ANALYSIS FOR JFETs YIELDS SIMILAR RESULTS

- Start $\rightarrow i_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_p}\right)^2$
- $\Delta i_D(AC) = \frac{V_X^2}{4}\cos(2\omega_C t)\left.\frac{d^2 i_D}{dV_{GS}^2}\right|_{I_D}$
- $\Delta i_D(DC) = \frac{V_X^2}{4} \cdot \left.\frac{d^2 i_D}{dV_{GS}^2}\right|_{I_D} = \frac{V_X^2}{4} \cdot \frac{2I_{DSS}}{V_P^2}$
- $\Delta i_D(DC) = \left(\frac{V_X}{V_P}\right)^2 \cdot \frac{I_{DSS}}{2}$
- Rectified Term Dependent upon I_{DSS} and V_P, not Operating Point

Figure 1.54

COMPARISON: Δi_C (BJT) VERSUS Δi_D (FET)

- Bipolar Transistor:

  Emitter Area = 576 µm^2

  I_C = 10 µA

  V_T = 25.68mV @ 25°C

  $$\Delta i_C = \left(\frac{V_X}{V_T}\right)^2 \cdot \frac{I_C}{4} = \frac{V_X^2}{264}$$

- JFET

  I_{DSS} = 20 µA (Z/ L = 1)

  V_P = 2V

  I_D = 10 µA

  $$\Delta i_D = \left(\frac{V_X}{V_P}\right)^2 \cdot \frac{I_{DSS}}{2} = \frac{V_X^2}{400 \times 10^3}$$

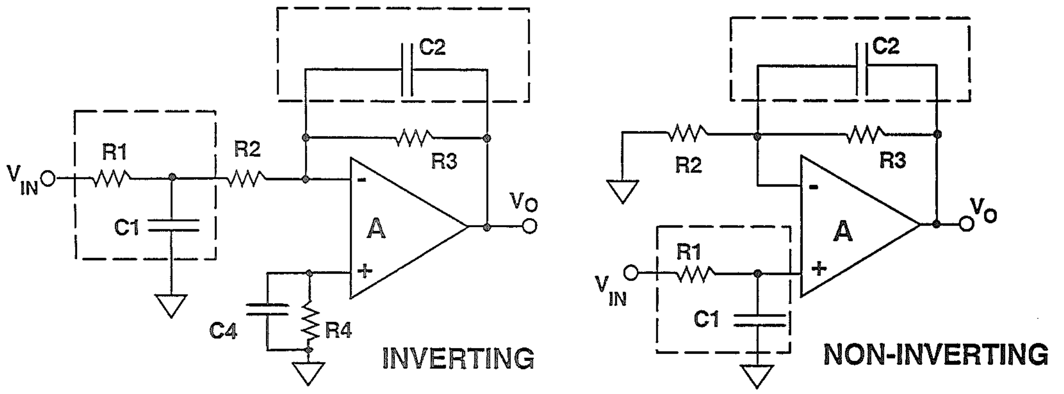
- BJTs are about 1500 times more sensitive than JFETs

Figure 1.55

Reducing RFI Rectification in Operational Amplifiers

Figure 1.56 shows low-pass filters used to reduce or eliminate unwanted interference. For the inverting circuit configuration on the left, the input resistor was split into two equal halves with the common point bypassed to ground by C1. The input resistor should be split into equal halves to prevent large values of C1 from appearing directly across the operational amplifier's input terminals. High capacitance at the input of some operational amplifiers can lead to loop instability. This is especially true for high-speed amplifiers. In general, the input filter's cutoff frequency should be no greater than 100 times the signal bandwidth to ensure that the circuit responds quickly enough to a step change in the input signal. For non-inverting amplifier configurations, the R1-C1 combination form the input filter that prevents noise from coupling directly into the amplifier's noninverting input. C2 limits the bandwidth (and hence wideband noise) of an amplifier but does not greatly assist in preventing rectification since it is unlikely to affect RF levels in the input stage.

USE LOWPASS FILTERS TO PREVENT RFI RECTIFICATION



SIGNAL GAIN $\approx \dfrac{-R3}{R1 + R2}$	SIGNAL GAIN $\approx 1 + \dfrac{R3}{R2}$
SIGNAL BW $\approx \dfrac{1}{2\pi R3C2}$	SIGNAL BW $\approx \dfrac{1}{2\pi R3C2}$
FILTER BW $\approx \dfrac{1}{2\pi R1C1} > 100 \bullet$ SIGNAL BW	FILTER BW $\approx \dfrac{1}{2\pi R1C1} > 100 \bullet$ SIGNAL BW

- C1, C2: Low-Inductance Capacitors
- Use Metal-Film Resistors for Low Noise and Parasitic Reactance

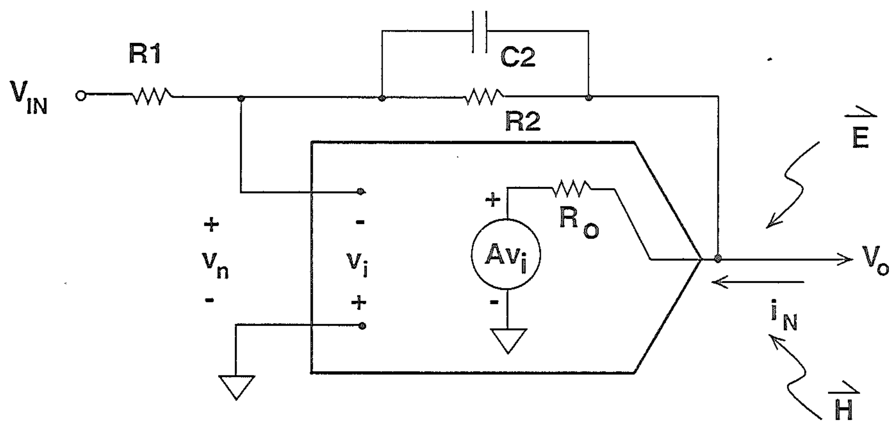
Figure 1.56

Equations illustrating the relationship between circuit element values and filter cutoff frequencies are given in Figure 1.56. Proper component selection is as important as the technique itself for successful noise filtering. Metal-film resistors should be chosen over composition or wire-wound ones for their low excess noise and low parasitic reactances. For best high-frequency noise filtering, low inductance NPO, COG, or film capacitors should be used.

Filtering operational amplifier inputs and limiting circuit bandwidths are both good techniques for eliminating noise. However, noise can also couple into an amplifier's input stages via its output (Figure 1.57). This is especially important in low-power circuits where operational amplifier output resistance and external circuit resistance levels are large.

The output resistance of operational amplifiers increases with frequency; thus it exhibits inductive behavior. If an amplifier has a low output impedance, external noise which finds its way to the output terminal is effectively grounded harmlessly. But if the output impedance rises with frequency then HF noise at the output is not attenuated so much and may couple to the amplifier input by passing backwards through the feedback network (Figure 1.58).

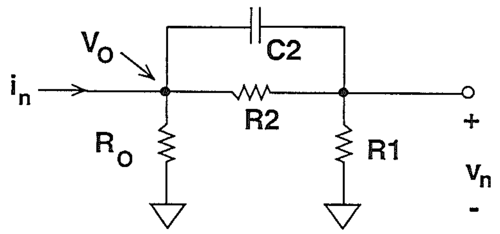
OP AMP OUTPUT IMPEDANCE ALSO AFFECTS RECTIFICATION SENSITIVITY



- Op Amp Output Impedance Increases With Frequency
- Larger R_O Produces More v_n at Op Amp Input

Figure 1.57

EQUIVALENT CIRCUIT OF OP AMP FEEDBACK NETWORK



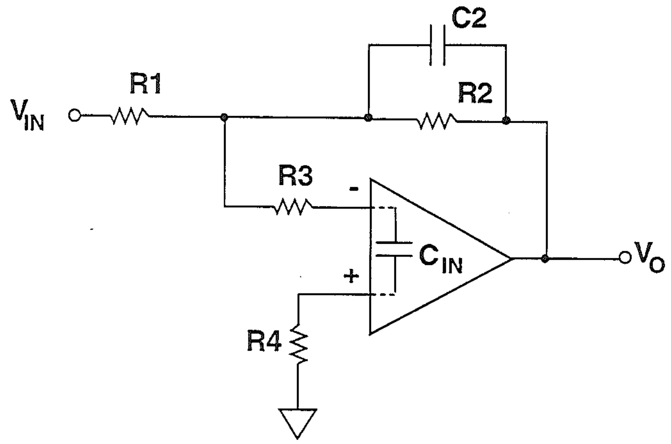
OP AMP R_O (% OF R1 + R2)	NOISE CURRENT IN R1 (% OF i_n)
0	0
0.1	0.09
1	0.9
10	9

- C2 INCREASES NOISE BY SHUNTING R2 @ HF

Figure 1.58

Here the feedback network is viewed from the source of the interference at the output looking toward the input. Noise currents produced are normally shunted to ground through the amplifier's low output resistance. As the frequency increases, the output resistance of the device also increases. This causes more noise current to appear as a noise voltage across R1. For example, when R_O is expressed a percent of R1+R2, the amount of noise appearing at the inputs of the device is directly proportional to the ratio of R_O to R1+R2. Although the capacitor connected across the feedback resistor is used to limit signal bandwidth, it unfortunately provides a direct path for noise at high frequencies by shunting R2. To prevent output-coupled noise from appearing at the inputs of a device would require an operational amplifier with zero output resistance over frequency — clearly unrealistic. Therefore, to prevent noise from coupling into the inputs via the feedback network, a resistor (R3) is connected between the sum node and the inverting terminal of the operational amplifier and forms a low-pass filter with the amplifier's input capacitance. Details for selecting the value of the resistor are outlined in Figure 1.59.

CURING OUTPUT CIRCUIT INDUCED RECTIFICATION



- R3 Forms LPF with C_{IN} ($3pF \leq C_{IN} \leq 5pF$)
- Choose $1k\Omega \leq R3 \leq 10k\Omega$
- Set $R4 = R3 + R1 \| R2$
- Watch for e_n (R3)

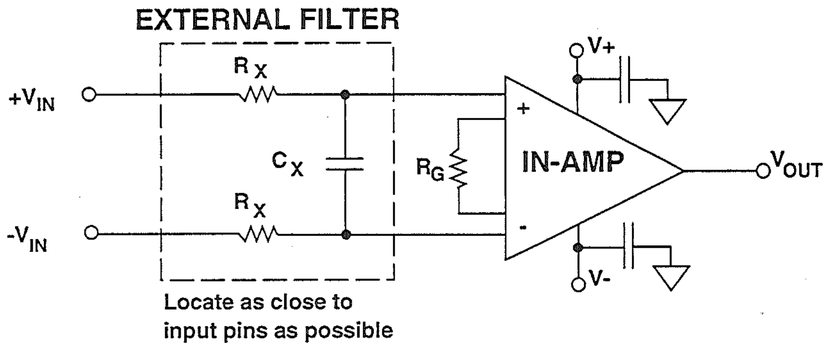
Figure 1.59

Reducing RFI Rectification in Instrumentation Amplifier Circuits

Filtering techniques used for operational amplifier circuits can also be applied to instrumentation amplifier circuits. As shown in Figure 1.60, low-pass filters are used in series with the differential inputs to prevent unwanted noise from reaching the inputs. Here, a capacitor, C_X, is connected across the inputs of the instrumentation amplifier and forms a differential low-pass filter with the two resistors R_X. An additional benefit of using a differentially-connected capacitor is that it reduces common-mode capacitance imbalance which helps to preserve high-frequency AC common-mode rejection. Since series resistors are required to form the low-pass filter, errors due to poor layout (CMR imbalance), component tolerance of R_X (input bias current-induced V_{OS}) and resistor thermal noise must be considered in the design process. In applications where the sensor is an RTD or a resistive strain gauge, R_X can be omitted, provided the sensor is close to the amplifier.

Some instrumentation amplifiers provide access to the base and emitter junctions of the input transistors (Figure 1.61). An external capacitor (C_X) can be connected across these terminals to reduce RFI rectification. This base-emitter bypass capacitor shunts high-frequency interference away from the base-emitter junction by forming a low-pass filter with an external resistance, R_X; however, a drawback to this technique is that the loop response of the amplifier input circuit is affected by the filter capacitor. As in the previous example, errors arising from the addition of a series resistance must be evaluated.

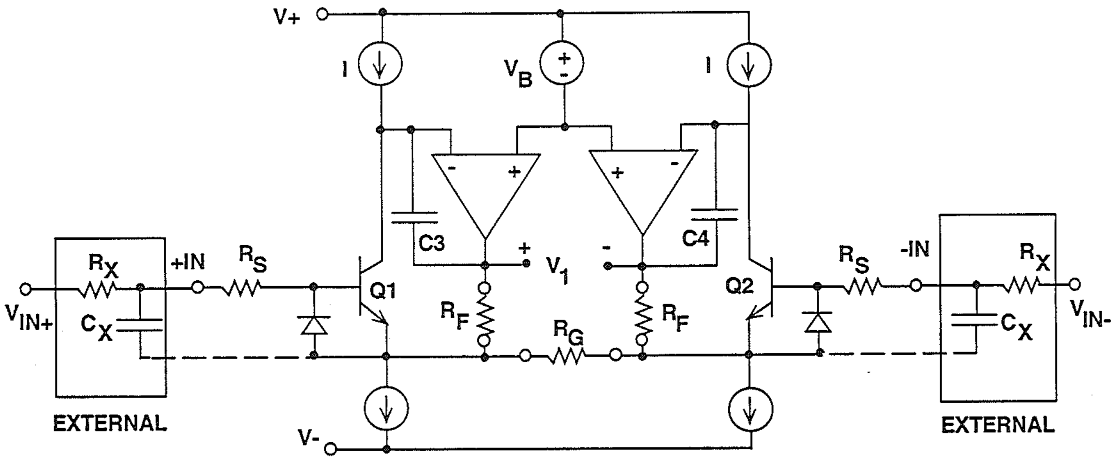
USE THE SAME TECHNIQUE FOR INSTRUMENTATION AMPS



- C_X forms differential RC filter with R_X; $\tau(LPF) = 2R_XC_X$
- C_X reduces common-mode capacitance imbalance → AC CMRR
- Evaluate R_X errors -- CMR, $I_b \Delta R_X$, $e_n(R_X)$
- R_X may be omitted if transducer is resistive and close to the amplifier

Figure 1.60

METHOD 2: BYPASS Q1 AND Q2 BASE-EMITTER JUNCTIONS WITH A CAPACITOR



- Must have access to the amplifier's base-emitter junctions
- C_X shunts interference away from base-emitter junctions
- C_X affects the amplifier's transient response
- Evaluate R_X errors

Figure 1.61

Since these capacitors can affect the loop stability of the input stage, they cannot be arbitrarily large. In fact, each instrumentation amplifier requires a specific value of capacitance depending on the gain setting. Exceeding the recommended capacitance value will damage the loop response. Figure 1.62 is a guide to selecting the recommended value of capacitance for each ADI instrumentation amplifier that provides direct access to the base and emitter terminals. Some, such as the AD625 and the AMP-01, require different values of capacitance for different gain settings. On the other hand, the AD626 does not require external R-C networks for noise filtering. At each input of the AD626, an internal 200kΩ resistor forms a 160kHz low-pass filter with an internal 5pF capacitor. The AD626 also provides access to an internal node for connecting a external capacitor for additional filtering.

DIFFERENT INSTRUMENTATION AMPLIFIERS REQUIRE DIFFERENT BASE-EMITTER BYPASSING

DEVICE	C_X	CONNECT C_X BETWEEN
AD524 AD624	<220pF	+IN and RG2 –IN and RG1
AD625	220pF (G<10) 680pF (G>10)	+IN and +GS –IN and –GS
AMP-01	220pF (G<10) 680pF (G>10)	+IN and pin 1 –IN and pin 2
AD620	<100pF (G>10)	+IN and pin 8 –IN and pin 1
AD626	RFI Filter Built-In	

Figure 1.62

Unfortunately not all instrumentation amplifiers are created equal. In Figure 1.63 are devices that do not provide direct access to the input transistors. Therefore, the base-emitter bypass technique cannot be used, and external networks for noise filtering must be considered. Before you use the base-emitter capacitance bypass technique do understand the topology of the amplifier.

NOT ALL INSTRUMENTATION AMPLIFIERS ARE EQUAL

- Do not use base-emitter bypass on the following devices:
  - AMP-02
  - AD365
  - AMP-03
  - AD521
  - AD522
  - AMP-04
  - AMP-05

Figure 1.63

Preventing RFI rectification in high-accuracy circuits requires an understanding of the electrical environment in which the circuit operates and using passive components (resistors, capacitors, ferrite beads) wisely. Passive components that exhibit low parasitic reactances should always be used. Always consider the impact of external filtering networks (input or output) on the overall performance. Finally, whatever methods are used to reduce RFI rectification, preserve high AC common-mode rejection by using balanced layout to maintain input symmetry.

PREVENTING RFI RECTIFICATION IN HIGH ACCURACY CIRCUITS

- Know your environment
- Use your arsenal in filtering:
  - Metal film resistors
  - Ferrite beads
  - Low-inductance capacitors
- Consider the effects of external filtering
- Balance input layout for high AC CMR

Figure 1.64

INPUT OVERVOLTAGE PROTECTION

Op amps and instrumentation amplifiers must often have the interface to the outside world, which may entail handling voltages that exceed their absolute maximum ratings. Sensors are often placed in an environment where a fault may connect the sensor to high voltages: if the sensor is connected to an amplifier, the amplifier inputs may see voltages exceeding its power supplies. Whenever its input voltage goes outside its supply range, an op amp may be damaged, even when they are turned off. Almost all op amps' input absolute maximum ratings limit the maximum allowable input voltage to the positive and negative supplies or possibly 0.3V outside these supplies. A few exceptions to this rule do exist, which can be identified from individual data sheets, but the vast majority of amplifiers require input protection if over-voltage can possibly occur.

INPUT OVERVOLTAGE PROTECTION

- Input Overvoltage Occurs When Either Input Exceeds the Supply Voltage
- Overvoltage May Occur When Supplies Are Turned Off
- Most Amplifiers can be Damaged by Overvoltage
- External Protection may be Required

Figure 1.65

Any op amp input will break down to the positive rail or the negative rail if it encounters sufficient over-voltages. The breakdown voltage is entirely dependent on the structure of the input stage. It may be equivalent to a diode drop of 0.7V or to a process breakdown voltage of 50V or more. The danger of an overvoltage is that when conduction occurs large currents may flow, which can destroy the device. In many cases, overvoltage results in current well in over 100mA, which can destroy a part almost instantly.

To avoid damage, current should be limited to less than 5mA. This value is a conservative rule of thumb based on metal trace widths in a typical op amp input stage. Higher levels of current can cause metal migration, which will eventually lead to an open trace. Migration is a cumulative effect that may not result in a failure for a long period of time. Failure may occur due to multiple over-voltages, which is a difficult failure mode to identify. Thus, even though an amplifier may appear to withstand over-voltage currents well above 5mA for a short period of time, it is important to limit the current to guarantee long term reliability.

Two types of conduction occur in over-voltage conditions, forward biasing of PN junctions inherent in the structure of the input stage or, given enough voltage, reverse junction breakdown.

The danger of forward biasing a PN-junction is that excessive current will flow and damage the part. As long as the current is limited no damage should occur. However, when the conduction is due to the reverse breakdown of a PN junction, the problem can be more serious. In the case of a base-emitter junction break down, even small amounts of current can cause degradation in the beta of the transistor (Reference 15). After a breakdown occurs, input parameters such as offset and bias current may be well out of specification. Diode protection is needed to prevent base-emitter junction breakdown. Other junctions, such as base-collector junctions and JFET gate source junctions do not exhibit the same degradation in performance on break down, and for these the input current should be limited to 5mA.

OVERVOLTAGE CHARACTERISTICS

- Op Amps Will Conduct to the Positive and Negative Rail Given Enough Input Voltage
- The Voltage at Which Conduction Occurs Depends on the Input Stage Construction
- This Voltage may be as Low as a Diode Drop or as High as 30V or More
- Conduction Current Needs to be Limited (Rule of Thumb: 5mA)
- Avoid Reverse Bias Junction Breakdown in Base-Emitter Junctions
- No Two Amplifiers Are the Same

Figure 1.66

Various factors affect the over-voltage characteristics of an op amp. Figure 1.67 lists some of the structures commonly found in input stages. Such circuit elements as input protection diodes usually form a diode from a supply. When the input voltage goes more than 0.6V outside that supply, then a diode from the substrate through the protection diode conducts current and limits the input voltage. There are many different ways of building an input stage, and, with them, many different types of over-voltage characteristics. This list gives examples of various structures and their possible characteristics, but it certainly not exhaustive. Sometimes the characteristic of a particular op amp can be determined from a simplified schematic in the data sheet by looking for PN-junctions. Often, though, the data sheet schematic does not contain enough information to determine over-voltage characteristics. In these where no obvious breakdown path exists or the simplified schematic is not shown, the best strategy is to measure the performance.

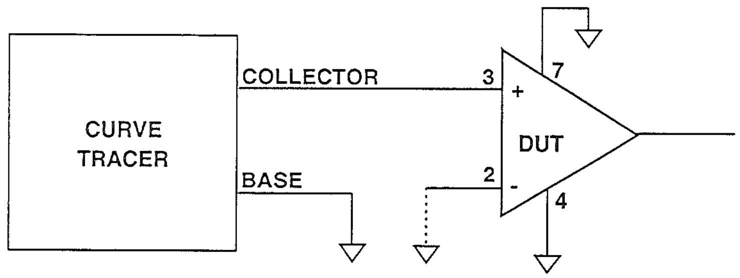
FACTORS AFFECTING OVERVOLTAGE CHARACTERISTICS

- INPUT PROTECTION DIODES: Usually Form a Diode from V–
- INPUT SERIES RESISTORS:
  - POLY: Form a Diode to V+
  - THIN FILM: Provides Current Limiting
- SUBSTRATE CONNECTION: Positive or Negative Supply
- INPUT TECHNOLOGY:
  - BIPOLAR: No Inherent Paths to Either Supply
  - JFET: Diode Formed from V– for P-Channel JFETs
- INPUT STAGE STRUCTURE: Is There a Conduction Path to Either Supply? Look for PN Junctions

Figure 1.67

A curve tracer can be configured to measure the over-voltage characteristics of any amplifier by connecting the part as shown in Figure 1.68. The curve tracer ramps a DC voltage on the input and measures the current flowing in the input stage. The supply voltages can be configured as desired; however for simplicity all measurements in this section were made with both supplies connected to ground. For most amplifiers this measurement yields the same results on both inputs. An exception to this is found with current feedback amplifiers, which have different characteristics at their inverting and noninverting inputs. If a curve tracer is not available, the measurements can be done using a DC voltage source and a multi-meter. Connect the source to the input through a 10kΩ resistor, and measure the current as a function of the input voltage. Both methods yield the same results.

OVER-VOLTAGE CURVE TRACER TEST SETUP



- Test Both Inputs of Op Amp -- Results are Identical for Voltage Feedback Types, but Not Current Feedback
- Force a Voltage Using the Collector Output
- Display Collector Current Versus Voltage
- Results May Differ Depending on Whether the Other Input is Open or Grounded

Figure 1.68

Figures 1.69 through 1.78 show different examples of over-voltage characteristics of various op amps. A varied sample of devices were tested but this is certainly not an exhaustive selection of amplifiers. As mentioned above, two basic types of over-voltage characteristics are exhibited in amplifiers. Forward biased junctions are evident because, as the input voltage rises above the supplies, distinct break points are encountered when internal diodes turn on. In most cases, this characteristic occurs within one or two diode drops of either supply. On the other hand, reverse junction breakdown does not occur until the input goes 4V or more outside either supply. Breakdown voltages may be as large as 50V or even more. Such measurements are important because it is much easier to protect an amplifier against damage due to over-stress if its behavior when over-stressed is well understood.

Figure 1.69 shows the over-voltage characteristic of an OP-177 precision amplifier. As the curve tracer shows, an internal diode turns on and starts conducting when the input reaches 0.6V above the positive supply. In the negative direction, we see that a diode also turns on, but the input current is limited by a series resistance. The OP-177 input stage's schematic partially illustrates where the parasitic diodes develop. Figure 1.70 shows the cross section for the process to illustrate more clearly where the diodes originate. When the input goes 0.6V below the negative supply the cathode of the protection diode forms the cathode of a parasitic diode from the p-type substrate. When this substrate diode turns on, current flows form the negative supply through the diode and the 500Ω input resistor, which accounts for the slope seen in the curve tracer photo.

The actual resistor is fabricated with diffused p-type material in an n-well as shown in Figure 1.70. The n-well is biased to the positive supply potential. Thus, when the input voltage rises 0.6V above the positive supply, the diode formed by the p-type resistor and the n-well is turned on. In this case the resistor is not in series with the input, so the current is not limited. Because PN-junctions are forward biased when the input goes outside either supply of the OP-177, the input current need only to limited to 5mA to protect the device.

EFFECTS OF INPUT PROTECTION DIODES AND DIFFUSED RESISTORS FOR THE OP-177

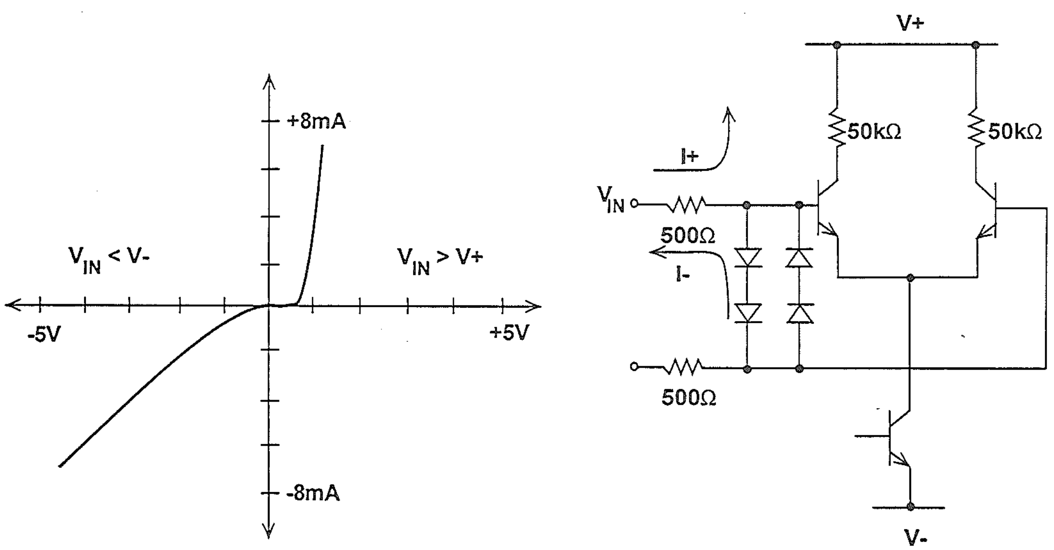


Figure 1.69

OP-177 CROSS-SECTION SHOWS CURRENT PATHS

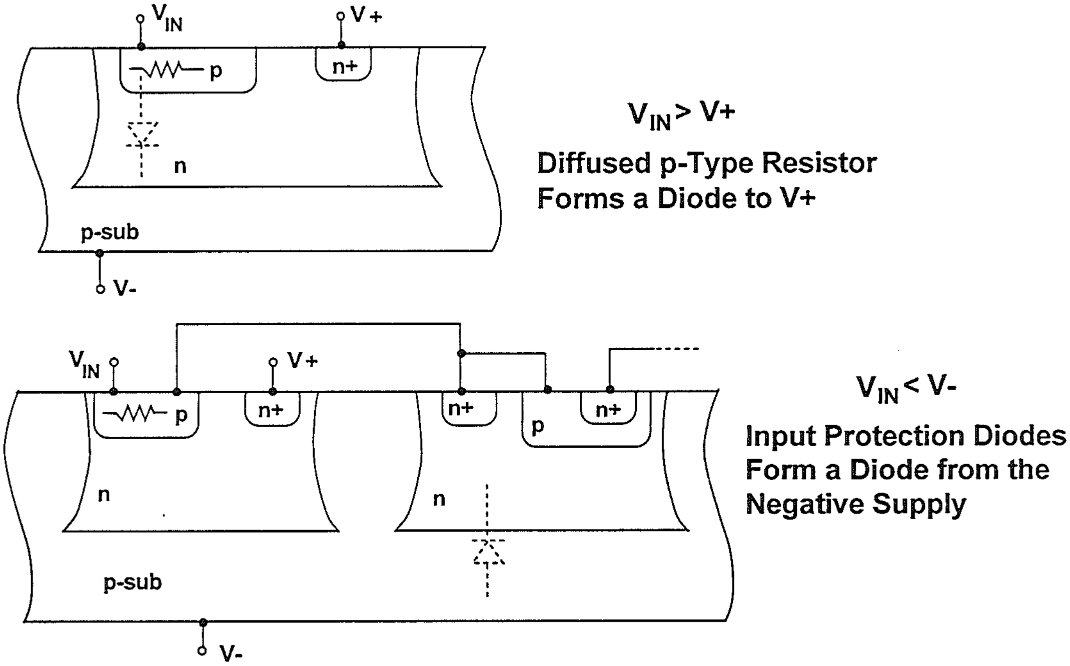


Figure 1.70

The OP-27 is an example of an input stage without input protection resistors (Figure 1.71). It still has input protection diodes, and like the OP-177, these form a diode from the substrate (Figure 1.72). As is evident in the curve tracer photo, the current is no longer limited in the negative direction because there are no input resistors. In the positive direction, the base collector junction of the NPN input transistor is forward biased at 0.6V above V+, and current starts to flow through the collector resistor. This collector resistor is also a p-type diffused resistor in an n-well. Thus, when the voltage across the resistor exceeds 0.6V a second diode turns on. The curve tracer actually shows the two diodes turning on. The first breakpoint occurs at 0.6V, which is followed by a sloped section due to the 20kΩ collector resistor, then a second breakpoint at 1.2V. Above this point, current flows with no limit. Since current is not limited in either direction, external protection is required to limit the input current to less than 5mA.

NO INPUT RESISTORS TO LIMIT OVERVOLTAGE CURRENT OF THE OP-27

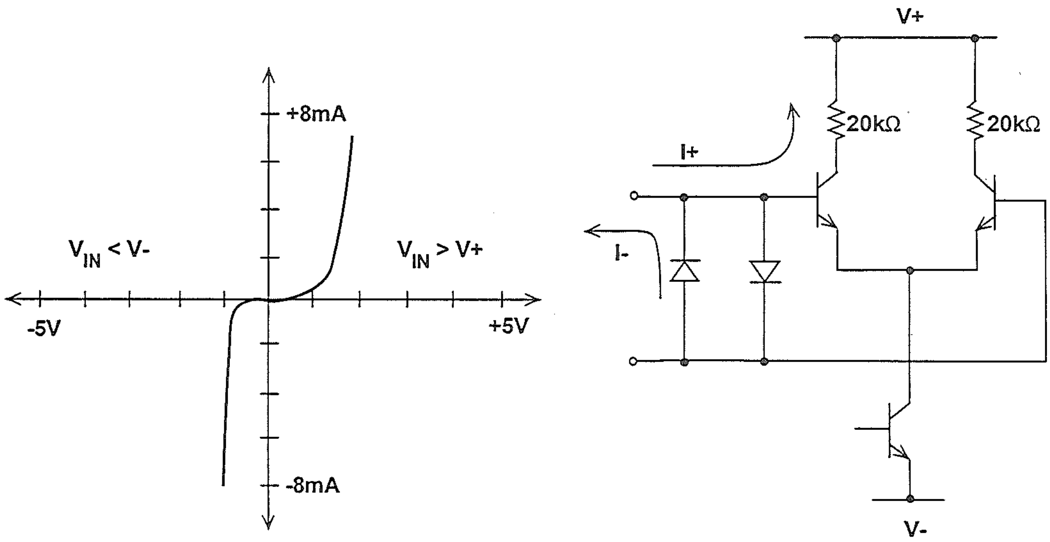
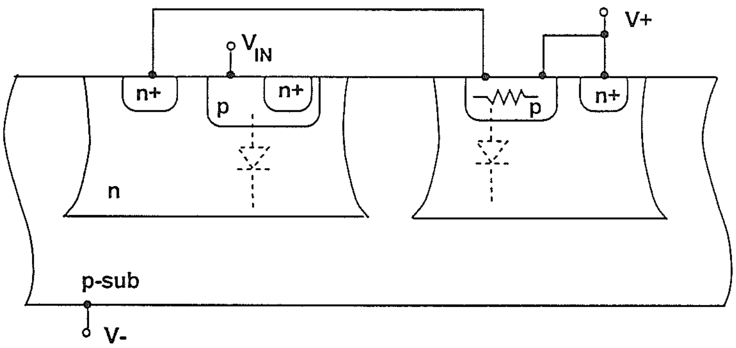


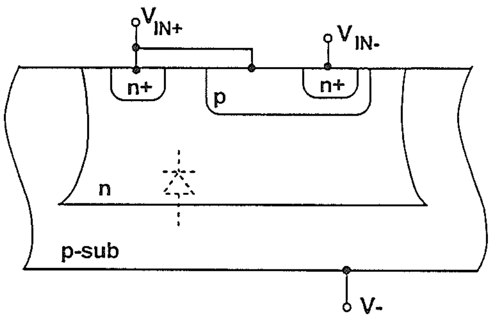
Figure 1.71

OP-27 CROSS SECTION



V_{IN} > V+

Forward Biases Base-Collector Junction First. Current is Limited By Collector Resistance. At 1.4V, Diffused Resistor Also Forward Biases.



V_{IN} < V-

Forward Biased Substrate Junction of Input Protection Diodes

Figure 1.72

Another common input stage device is a p-channel JFET differential pair. The part used as an example is the OP-42, but other JFET input devices such as the AD711, AD712, and OP-282 exhibit similar characteristics. As the curve tracer photo shows (Figure 1.73) the over-voltage characteristic is very different from those of the previous amplifiers. There is no over-voltage conduction path in the positive direction because there are no PN-junctions from the gate (Figure 1.74). When a positive over-voltage is applied to the inputs, the n-well and n-gate are pulled above V+. Thus, the PN-junctions (formed by the gate to drain and n-well to substrate) are reverse biased and no conduction occurs. If the input voltage is increased the input junctions eventually break down. For the OP-42 this breakdown occurs at approximately 44V. This breakdown, unlike base-emitter breakdown, is not damaging to the JFETs as long as the input current is limited to less than 5mA. However, to ensure that the input is not damaged, external clamping, using a diode, is recommended if the input voltage approaches the breakdown region.

P-CHANNEL JFET INPUT OF THE OP-42

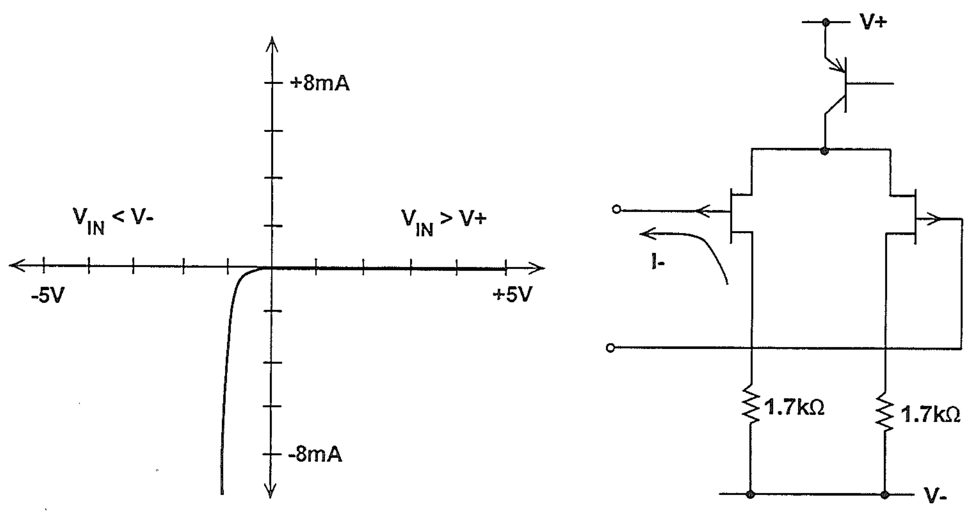


Figure 1.73

OP-42 CROSS SECTION

- $V_{IN} > V+$ No Current Path
  Breakdown Voltage > 30V
- $V_{IN} < V-$ P-Channel JFET Forms a Diode
  From V-

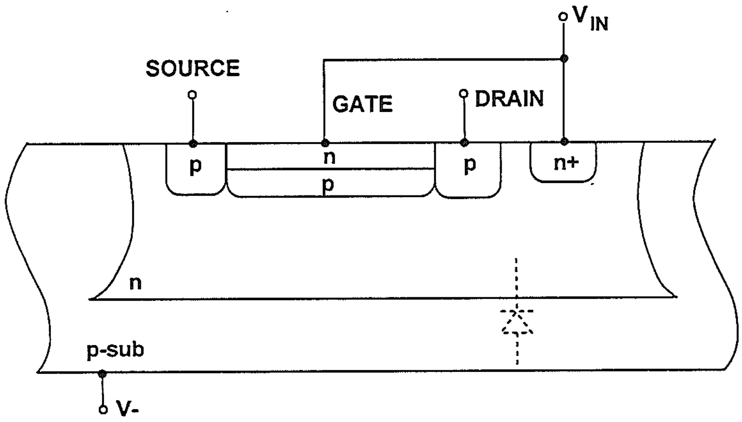
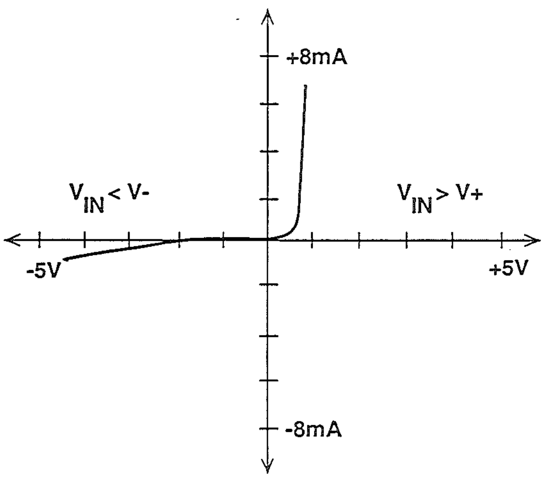


Figure 1.74

As the cross-section shows, the conduction path when the input goes below the negative supply is through the n-well, which is connected to the input pins. This is a simple PN-junction from the substrate. As with the previous two amplifiers, as long as input current is limited to 5mA, no damage will occur.

All the parts discussed so far are fabricated on a p-type substrate, which is biased to the negative supply, but amplifiers may also be made on n-type substrates which are biased positive. The OP-295 is an example. Figure 1.75 shows that the OP-295 has almost identical characteristics to the OP-177. For a negative over-voltage a diode from V- turns on, and the diffused 5kΩ input resistors limit the over-voltage current. In the positive direction, a diode from the p-type resistor turns on to V+. The process cross-sections are shown in Figure 1.76.

OP-295 ILLUSTRATES N-TYPE SUBSTRATE



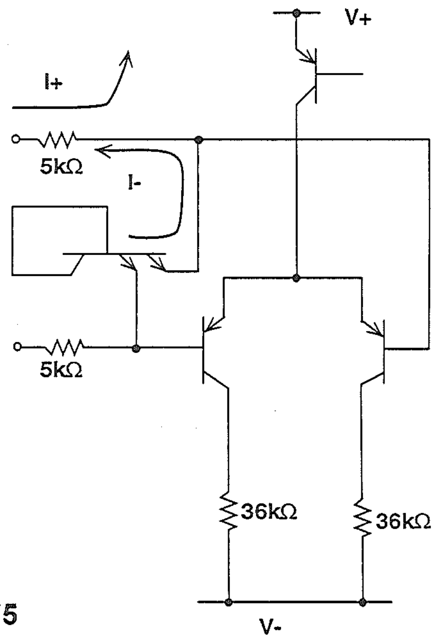
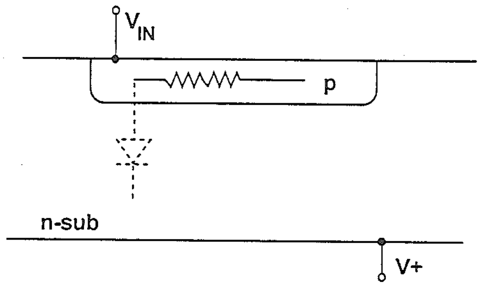


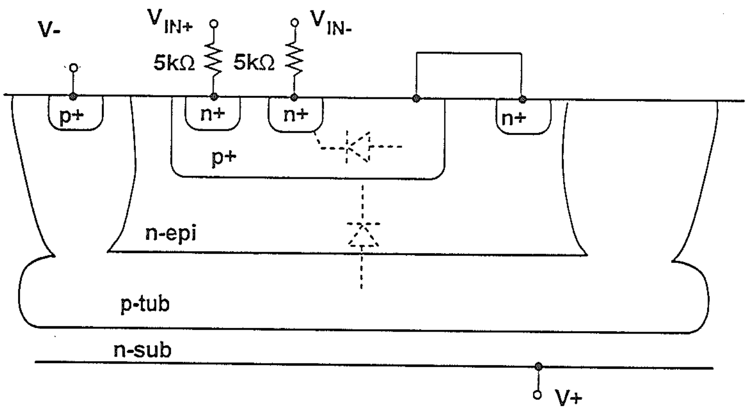
Figure 1.75

OP-295 CROSS SECTION



V_{IN} > V+

Diffused Resistor Forms Diode to Positive Supply



V_{IN} < V-

Protection Diode Provides Path from V-

Figure 1.76

The substrate is n-type and biased to the positive supply. NPN devices are built in p-wells biased to the negative supply. The input protection device on the OP-295 is an NPN transistor with two emitters which serves as a zener in series with a diode to clamp the differential input voltage to approximately ±10V. As the cross-section shows, this combination forms two diodes when the input goes below V-. Once these diodes conduct the input resistor limits the current. In the positive direction, the diffused input resistor forms a diode to the positive supply, which turns on when the input goes 0.6V above V+. In this case the resistor is bypassed, no current limiting occurs, and an external series resistor is needed to limit current to less than 5mA.

Zener diodes are sometimes used to protect PNP input stages from large differential voltages. The OP-213 input stage has two pairs of PNP input transistors (Figure 1.77). The breakdown in the negative direction is very simple. The collector-base junction of the PNP transistor is turned on when the input goes below V- (Figure 1.78). In the positive direction, however, the base-emitter junction of the protection transistor is reverse biased and experiences zener breakdown. Current flows to the positive supply through diode D1. In this case, the protection transistors are designed to experience zener breakdown and no damage occurs as long as the current is limited. If there is uncertainty about the type of breakdown occurring, an external diode to the positive supply will ensure that base-emitter breakdown does not occur.

OP-213 ILLUSTRATES ZENER DIODE BREAKDOWN

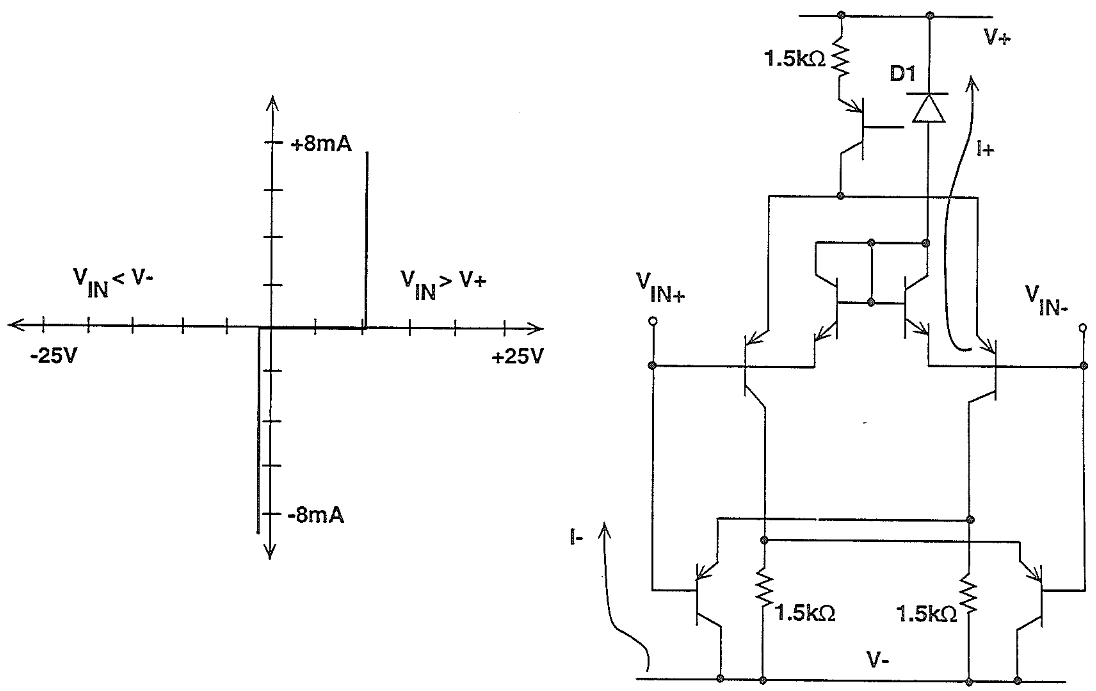


Figure 1.77

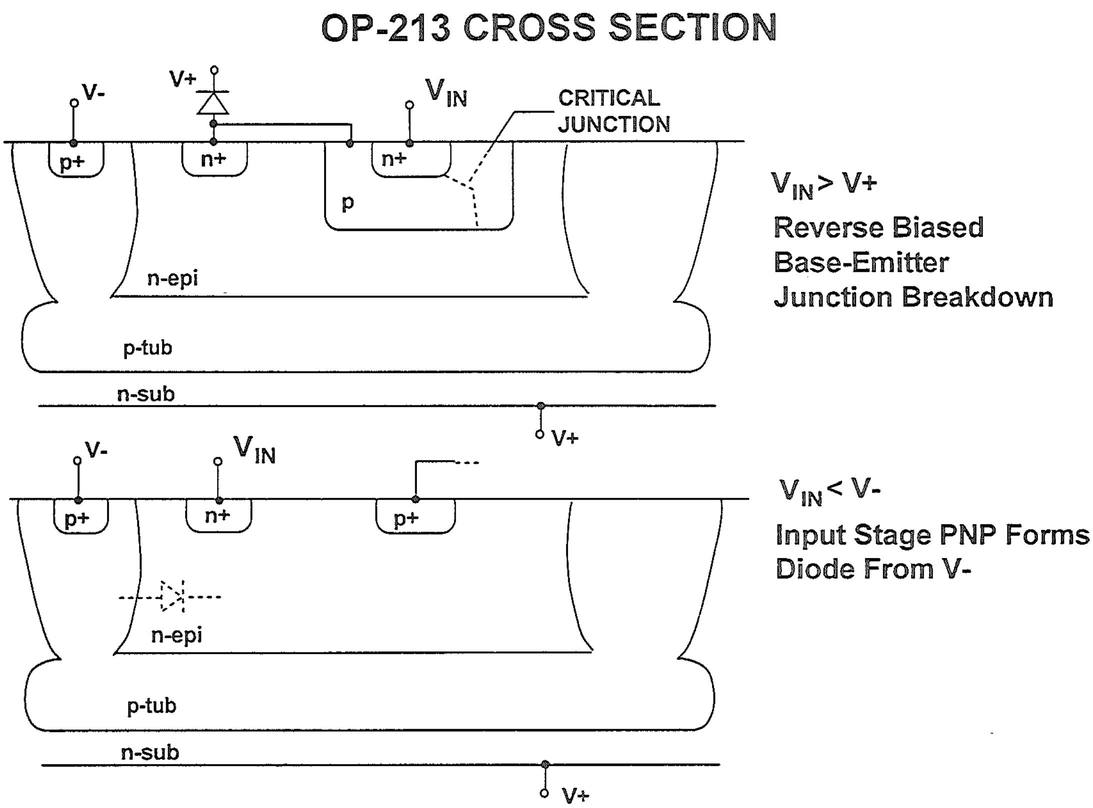


Figure 1.78

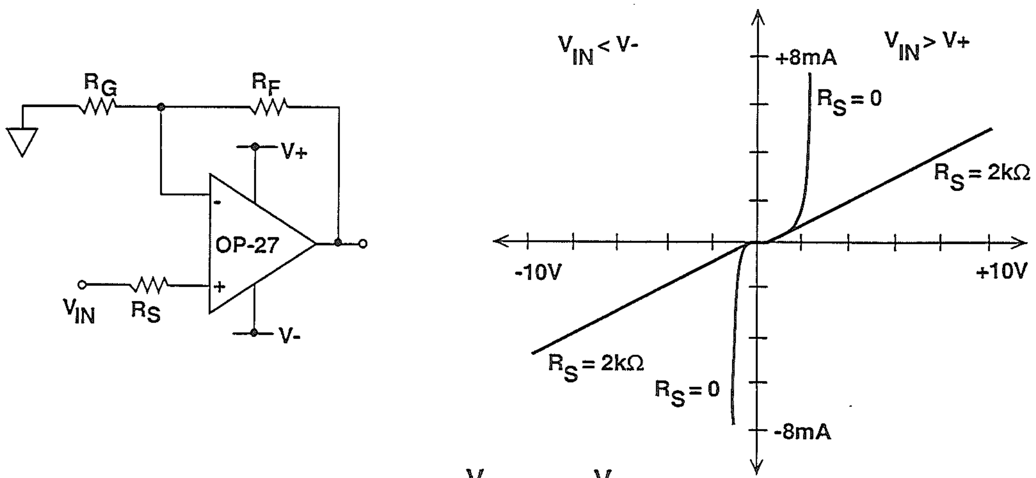
Whenever an op amp may experience over-voltage it should be protected. Except for a few amplifiers with internal protection, most amplifiers will need external resistors and, in some cases, diodes, as summarized in Figure 1.79. If an amplifier has internal PN-junctions that turn on when the input voltage goes outside the rail, then the only protection needed is a resistor to limit the current to 5mA. This current level is a safe value based on typical amplifier input stage structures. In cases where reversed biased junctions break down external diodes are needed. Usually breakdown occurs to one of the supplies only, and only one diode is needed. A series resistor should be also included to limit the current through the diode to safe levels. Whenever there is doubt about how a particular amplifier behaves, its characteristics should be checked on a curve tracer.

OVERVOLTAGE EFFECTS

- Junctions may be Forward Biased if the Current is Limited
- In General a Safe Current Limit is 5mA
- Reverse Bias Junction Breakdown is Damaging Regardless of the Current Level, Unless the Junction has been Designed to Break Down
- When in Doubt, Protect with External Diodes
- Curve Tracers Can be Used to Check the Overvoltage Characteristics of a Device

Figure 1.79

SERIES RESISTOR LIMITS OVERVOLTAGE CURRENT



- Pick R_S Such That: $R_S = \dfrac{V_{in\ max} - V_{supply}}{5mA}$
- For Example, When V_{supply} is Turned Off, and $V_{in\ max} = 10V$, $R_S = 2k\Omega$

Figure 1.80

Figure 1.80 shows the difference between a protected and an unprotected input stage. The curve tracer photo shows the OP-27 over-voltage characteristic without a series resistor. The current reaches 10mA with less than 1V on the input. With a 2kΩ resistor in series with the input, the current is limited to ±5mA for input voltages up to 10V outside the supply voltages. R_S is chosen so that at maximum over-voltage the current is limited to 5 mA. Most op-amp applications require over-voltage protection at only one input, but there are a few (differential amplifiers, for example) where both inputs can experience over-voltage and both must be protected. The need for both inputs to be protected is much commoner with instrumentation amplifiers.

It is evident that whenever resistance is added in series with an amplifier's input, its offset and noise performance is slightly degraded. In Figure 1.81, the additional noise and offset are calculated. The thermal noise of the resistor, the voltage noise due to the amplifier noise current flowing in the resistor and the input noise voltage of the amplifier are added together (using root sum of squares addition, since the noise voltages are uncorrelated) to determine the total input noise and may be compared with the input voltage noise in the absence of the protection resistor.

A protection resistor in series with an amplifier input has a voltage drop due to the amplifier bias current flowing in it. This drop appears in the system as an increase of offset voltage (and, if the bias current changes with temperature, offset drift). In amplifiers where bias currents are approximately equal a resistor in each input will tend to balance the effect and reduce the error.

If large resistors are necessary for adequate protection these noise and offset effects may become intolerable. In such a case it may be possible to use external Schottky diodes to clamp at higher current levels than the 5 mA limit on internal currents.

Figure 1.82 shows the benefit of adding an external diode to clamp the input over-voltage. The OP-213 with its 10V positive breakdown is used as an example. As the curve tracer shows, the clamp diode turns on when the input exceeds V+ by 0.6V, and the series resistor limits current in both directions. Two configurations are shown. The upper configuration offers lower leakage for small input voltages because the reverse bias on the diode is small but cannot be used with inputs >50mV as the diode turns on and the leakage goes up. In the lower configuration the cathode of the diode is connected to V+. The diode will not turn on until the input exceeds V+. This arrangement has higher leakage. A series resistor is still needed to limit the current.

SERIES RESISTOR EFFECTS

- Increases Noise:

$$E_{n\,Total} = \sqrt{e_{n\,op\,amp}^{\,2} + e_{n\,Rs}^{\,2} + (Rs \cdot i_{n\,op\,amp})^2}$$

- Voltage Offset Increases Due to Bias Currents:
  - V_{OS} Total = $V_{OS} + I_b R_S$
  - Balance Input Impedance to Minimize I_b Errors, $R_S = R_1 || R_2$
  - V_{OS} Total = $V_{OS} + I_{OS} R_S$

Figure 1.81

DIODE CLAMP PREVENTS REVERSE BIAS BREAKDOWN

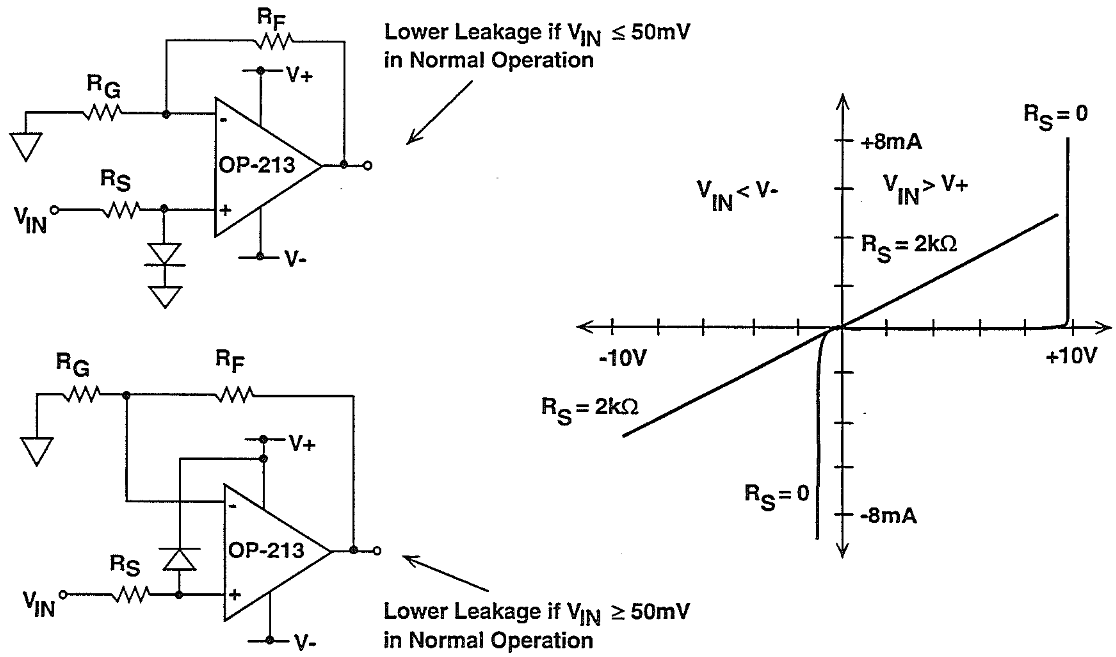


Figure 1.82

The choice of diode is important in applications where low bias current is important. Any component added to the input pin contributes some leakage. The choice of the diode depends on the bias current requirement for the application. The table in Figure 1.83 shows various protection diodes and their leakage currents. The common 1N914 or 1N4148 diode typically has 10nA of leakage current and is not a very good choice when using a JFET amplifier. A good choice for a protection diode is the base-collector junction of a 2N3906 transistor which is inexpensive and has leakage currents on the order of 10pA. For the most demanding applications, low leakage FETs may be required. Such parts as the 2N4117A and PAD1, from Siliconix, provide protection with leakage currents under 1pA. Temperature effects must also be considered, as the leakage current of diodes and JFET junctions doubles for every 10°C temperature rise.

The parts discussed do not have internal over-voltage protection. However, there is a small group of amplifiers that are designed to handle input voltages which exceed their supplies (Figure 1.84). Some of these parts have thin film resistors in series with their inputs. These resistors are isolated from the silicon and do not form a diode to a supply. They limit current as an external resistor does. The AMP-02 and the AD524 have protection JFETs in the input stage that turn on during overvoltage and allow the inputs to reach ±60V without damage.

DIODE CLAMP LEAKAGE AFFECTS BIAS CURRENT

- Diode Leakage May Be Significant Compared To I_b
- If Important: Pick A Low Leakage Diode
- Remember JFET Leakage Doubles for every 10°C Rise

Diode	Leakage @ 25°C
1N914 1N4148	10nA 10nA
2N3906 (Base-Collector Junction)	10pA
2N4117A* PAD1* * Siliconix JFETS	1pA 1pA

Figure 1.83

SOME AMPLIFIERS HAVE BUILT-IN PROTECTION

- OP-90: Thin Film Input Resistors Allow the Input Voltage to Exceed the Supplies By 20V
- AMP-02: Protection FETs Allow 60V Inputs
- AD524, : AD626 Thin-Film Input Resistors Allow 60V Inputs
- AMP-03, : SSM-2141 25kΩ Thin-Film Input Resistors
- SSM-2143: 12kΩ Thin-Film Input Resistors

Figure 1.84

Input over-voltage is a common problem. It can cause serious damage if suitable protection is not provided by the circuit designer. To provide such protection it is necessary to identify the values of over-voltage to be expected, and to understand the behavior of the amplifier used under over-voltage stress. Armed with this information the designer can provide suitable protection for his circuit.

AMPLIFIER OUTPUT PHASE REVERSAL

Some amplifiers exhibit output voltage phase reversal when their input exceeds their common mode range. Phase reversal is usually associated with JFET input amplifiers, but some bipolar devices (especially single supply amplifiers) may also be susceptible. Phase reversal does not harm the amplifier, but it may be disastrous in a servo loop! Once the amplifier inputs return within the correct operating range, output phase reversal ceases. Although a number of op amps suffer from phase reversal it is, surprisingly, rarely a problem in system design. It may also be necessary to consult the amplifier manufacturer, since phase reversal information rarely appears on data sheets.

Phase reversal in a BiFET op amp may be prevented by adding an appropriate input series resistance to limit the input current. Biplolar input devices may be protected by using a schottky diode to clamp the input to within a few hundred millivolts of the appropriate, usually negative, rail. For more information on phase reversal and how to prevent it, refer to Reference 1, Section XI, pages 1-10.

AMPLIFIER LATCH-UP

Destructive latch-up on power-up is rare in modern ICs. The typical trigger mechanism for latch-up is a signal which is present at an input of a device before power is applied. This may cause a parasitic SCR to fire, large currents flow between the supply pins and the device is damaged. This was common in older CMOS devices, but most IC op amps do not behave in this way.

If latch-up is a concern, the amplifier may be easily checked easily for susceptibility. Apply an input to the amplifier and, using current-limited supplies so that no damage is done if latch-up does occur, apply power to the device. If the device has more than one supply, it should be tested with every possible sequence of supply turn-on (i.e., V_{CC} then V_{SS} then V_{logic}; V_{SS} then V_{CC} then V_{logic}; etc., through ALL combinations). The test should be performed with all possible inputs as well. All op amps discussed in this section of the book have been checked in this way and are free from latch-up.

BEWARE OF AMPLIFIER OUTPUT PHASE REVERSAL

- Sometimes Occurs in FET and Bipolar Input (Especially Single-Supply) Op Amps when Input Exceeds Common Mode Range
- Does Not Harm Amplifier, but may be Disastrous in Servo Systems!
- Not Usually Specified on Data Sheet, so Amplifier Must be Checked
- Easily Prevented:

  BiFETs: Add Appropriate Input Series Resistance

  Bipolars: Use Schottky Diode to Clamp Input Within the Supply Rails. Note: Some CB Process Devices Exhibit Phase Reversal with *POSITIVE*, and not Negative Inputs.

Figure 1.85

OP AMP LATCH-UP RARELY OCCURS, BUT:

- If it does occur, it usually occurs when power is applied in the presence of an input signal (common occurrence in older CMOS devices)
- Large currents due to SCR action may destroy device
- Carefully observe all *Absolute Maximum* specifications
- If in doubt, consult manufacturer and test amplifier for latch-up
- "Catch" diodes (Schottky) from input to positive and negative supply rails may be added for additional protection against latch-up

Figure 1.86

TRANSMITTING PRECISION SIGNALS IN NOISY ENVIRONMENTS

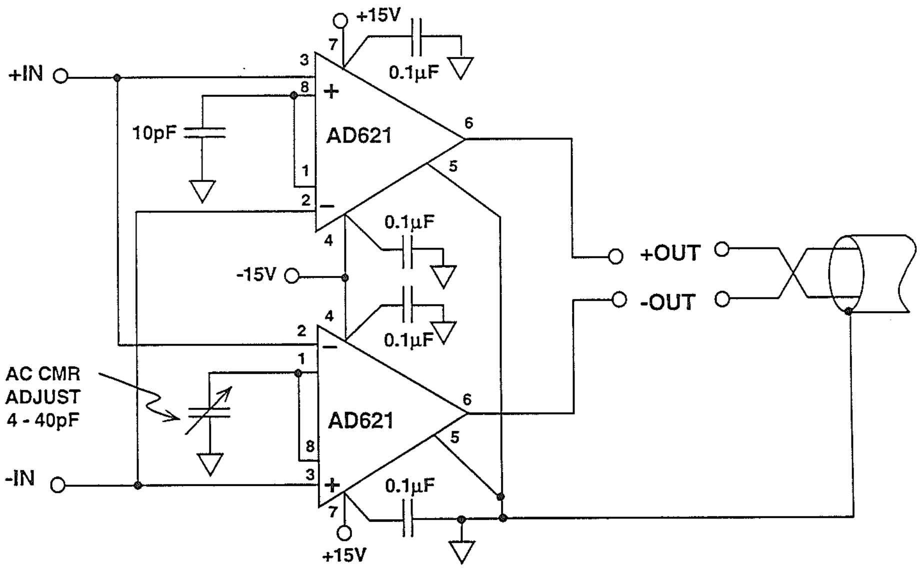
Signal integrity is often compromised in the interconnections between portions of electronic systems (Reference 16) . In industrial applications, precision signals must often be transmitted long distances to other signal conditioning or A/D conversion circuitry without loss of accuracy. Balanced transmission and reception techniques can reject both HF and LF interference and maintain accuracy. There are many ways of implementing balanced signal transmission and reception, but not all of them are suitable for the transmission of high precision DC and LF signals. Most of the viable techniques use op-amps and instrumentation amplifiers.

PARALLELING AMPLIFIERS FOR A PRECISION LINE DRIVER [REFERENCE 16]

One of the more important tasks of any signal conditioner is preserving signal-to-noise ratio. The circuit in Figure 1.87 illustrates a method of preserving signal-to-noise ratio in a system where two instrumentation amplifiers are connected together to form a precision differential line driver. The signal-to-noise ratio of the circuit is improved by $\sqrt{2}$ over a single amplifier because the output signals add differentially and are multiplied by 2, while the input noise only increases by $\sqrt{2}$.

In this circuit, the inputs of a pair of AD621s are connected in anti parallel while the outputs signal appears differentially between the outputs of the AD621s. Even though each instrumentation amplifier can be configured to gains of 10 or 100, the circuit gain is twice this because of the differential output connection. The line driver exhibits an excellent transient response even when the output is 40 V_{p-p} (Figure 1.88). Its performance is summarized in Figure 1.89.

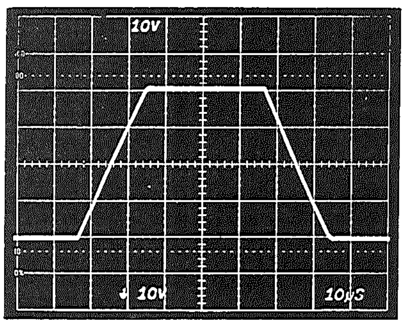
STACKING TWO INSTRUMENTATION AMPLIFIERS MAKES A PRECISION DIFFERENTIAL LINE DRIVER



G = 100

Figure 1.87

STACKED INSTRUMENTATION AMPLIFIER EXHIBITS EXCELLENT TRANSIENT RESPONSE



Horizontal Scale: 10µs/div.
Vertical Scale: 10V/div.

G = 10

V_{in} = 2V peak-to-peak

Figure 1.88

PERFORMANCE OF PRECISION STACKED LINE DRIVER

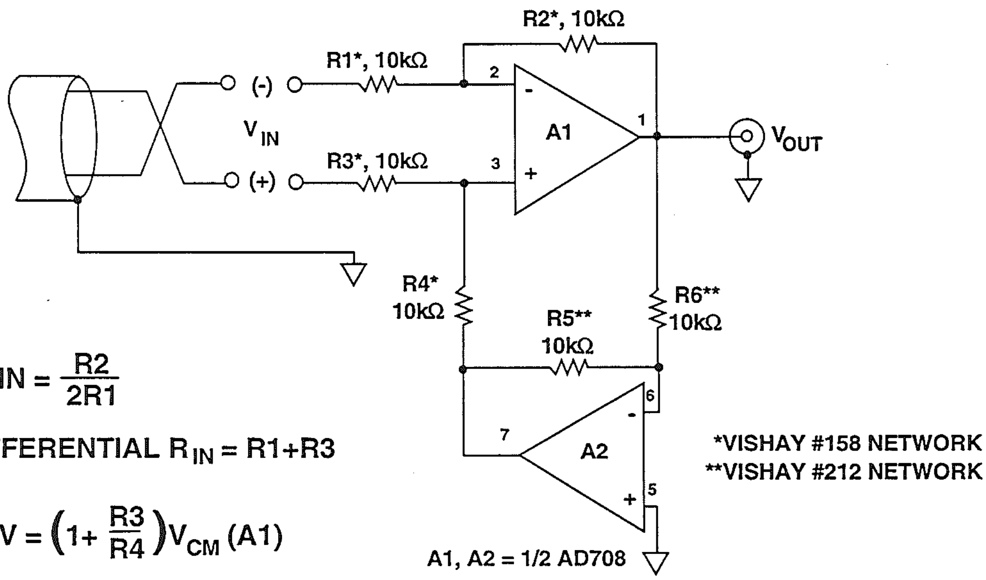
Nonlinearity:		
	Gain = 10	0.0002 %
	Gain = 100	0.004 %
Slew Rate:		2.4 V/µs
Maximum Output Level:		40 V_{p-p} (14.2 V_{rms})
–3 dB Bandwidth:		
	Gain = 10	650 kHz
	Gain = 100	170 kHz
CMRR:		
	f = 60 Hz	95 dB
	f = 1 kHz	73 dB
Input Noise Spectral Density @ 1 kHz:		
	G = 10	16.2 nV/$\sqrt{Hz}$
	G =100	10.5 nV/$\sqrt{Hz}$

Figure 1.89

A HIGH-PRECISION, BALANCED LINE RECEIVER [REFERENCE 17]

As important as the line driver is the differential line receiver. A line receiver's primary function is to reject common-mode noise. Figure 1.90 is an example of a high-precision, balanced line receiver built with a precision dual op amp, the AD708. The AD708's tight matching of V_{OS}, TCV_{OS}, I_B, and CMRR yield a design that achieves 95 dB CMR from dc to 1 kHz. The common-mode performance of the circuit is shown in Figure 1.91.

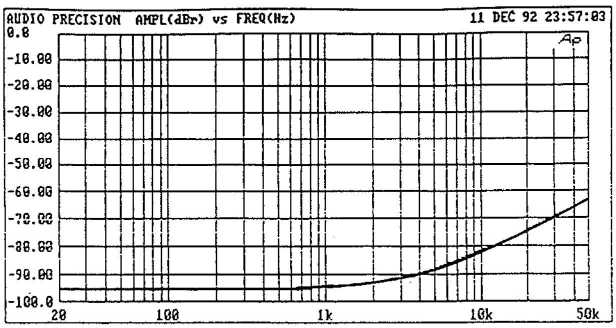
USING A HIGH PERFORMANCE DUAL OP AMP AS A PRECISION DIFFERENTIAL LINE RECEIVER



- GAIN = $\frac{R2}{2R1}$
- DIFFERENTIAL R_{IN} = R1+R3
- CMV = $\left(1+\frac{R3}{R4}\right)V_{CM}$ (A1)

Figure 1.90

USING A PRECISION DUAL OP AMP WITH MATCHED CHARACTERISTIC YIELDS HIGH CMRR

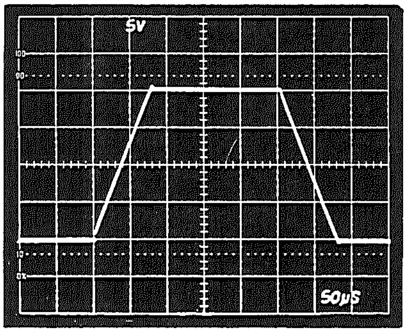


V_{in} = 2V peak-to-peak

Figure 1.91

The circuit uses a balanced topology that equalizes the signal currents in the two inputs, which is important in applications where a number of twisted pairs are bundled together. In a bundle the risk of crosstalk is high — balancing the signal currents in each pair of cables greatly reduces it by minimizing interfering magnetic fields. A2 provides a push-pull feedback arrangement to drive R4, the previously grounded reference terminal, in antiphase to V_{OUT}. This circuit provides a single-ended output with a gain of one-half, assuming that R2 = R1. The common-mode range of the circuit is equal to the common-mode range of conventional line receivers, but common-mode rejection at VOUT is about double. A2's feedback network ratio, R5/R6, affects the balance in the circuit and not its CMR but the matching of R1 and R3, and R2 and R6 is critically important to the CMR. The large-signal behavior of this balanced line receiver is shown in Figure 1.92, and its performance is summarized in Figure 1.93.

PRECISION DIFFERENTIAL LINE RECEIVER EXHIBITS EXCELLENT LARGE SIGNAL TRANSIENT RESPONSE



Vertical Scale: 5V/div.
Horizontal Scale: 50µs/div.

V_{in} = 20V peak-to-peak

Figure 1.92

PERFORMANCE OF PRECISION LINE RECEIVER

Nonlinearity	0.00045
Slew Rate	0.3 V/µs
Common-Mode Input Range	± 28 V
–3 dB Bandwidth	660 kHz
CMRR	95 dB

Figure 1.93

ISOLATING TRANSDUCERS

There are many applications where it is desirable, or even essential, for a transducer to have no direct ("galvanic") electrical connection with the system to which it is supplying data, either in order to avoid the possibility of dangerous voltages or currents from one half of the system doing damage in the other, or to break an intractable ground loop. Such a system is said to be "isolated" and the arrangement which passes a signal without galvanic connections is known as an "isolation barrier".

The protection of an isolation barrier works in both directions, and may be needed in either, or even in both. The obvious application is where a sensor may accidentally encounter high voltages and the system it is driving must be protected, but it is equally possible that a sensor may need to be isolated from accidental high voltages arising downstream, in order to protect its environment:- examples include the need to prevent the ignition of explosive gases by sparks at sensors and the protection from electric shock of patients whose ECG, EEG or EMG is being monitored. The ECG case is interesting as protection may be required in *both* directions: the patient must be protected from accidental electric shock, but if the patient's heart should stop the ECG machine must be protected from the very high voltages (>7.5 kV) applied to the patient by the defibrillator which will be used to attempt to restart it.

WHERE IS ISOLATION USED?

- Transducer is at a High Potential Relative to other Circuitry (or may become so under fault conditions)
- Transducer may not Carry Dangerous Voltages, Irrespective of Faults in other Circuitry (e.g. Patient Monitoring and Intrinsically Safe Equipment for use with Explosive Gases)
- To Break Ground Loops

Figure 1.94

Just as interference, or *unwanted* information, may be coupled by electric or magnetic fields, or by electro-magnetic radiation, these phenomena may be used for the transmission of *wanted* information in the design of isolated systems. The commonest isolation amplifiers use transformers, which exploit magnetic fields, and another common type uses small high voltage capacitors, exploiting electric fields.

Opto-isolators, which consist of an LED and a photocell, provide isolation by using light, a form of electro-magnetic radiation. Different isolators have differing performance: some are sufficiently linear to pass high accuracy analog signals across an isolation barrier, with others the signal may need to be converted to digital form before transmission if accuracy is to be maintained.

TECHNIQUES FOR ISOLATION

- Electric Field — Capacitive Signal Coupling
- Magnetic Field — Transformer Coupling
- Electromagnetic — Optical Coupling
- Sometimes a Technique will not have Adequate Linearity-- In Such Cases There are two Possibilities:
  - Voltage-Frequency Conversion / Transmission / Frequency-Voltage Conversion
  - Analog-Digital Conversion Before Transmission Across the Isolation Barrier

Figure 1.95

Transformers are capable of analog accuracy of 12-16 bits and bandwidths up to several hundred kHz, but their maximum voltage rating rarely exceeds 10 kV and is often much lower. Capacitively coupled isolation amplifiers have lower accuracy, perhaps 12-bits maximum, lower bandwidth and lower voltage ratings - but they are cheap. Optical isolators are fast and cheap, and can be made with very high voltage ratings (although 4-7 kV is one of the commoner ratings), but they have poor linearity and are not usually suitable for direct coupling of precision analog signals (although there are some exceptions to this generalization they tend to be expensive).

Linearity and isolation voltage are not the only issues to be considered in the choice of isolation systems. Power is essential. Both the input and the output circuitry must be powered and unless there is a battery on the isolated side of the isolation barrier (which is possible, but rarely convenient) some form of isolated power must be provided. Systems using transformer isolation can easily use a transformer (either the signal transformer or another one) to provide isolated power, but it is impracticable to transmit useful amounts of power by capacitive or optical means and systems using these forms of isolation must make other arrangements to obtain isolated power supplies - this is a powerful consideration in favor of choosing transformer isolated isolation amplifiers: they almost invariably include an isolated power supply. An example is the AD210 (Figure 1.96).

AD210 3-PORT ISOLATION AMPLIFIER

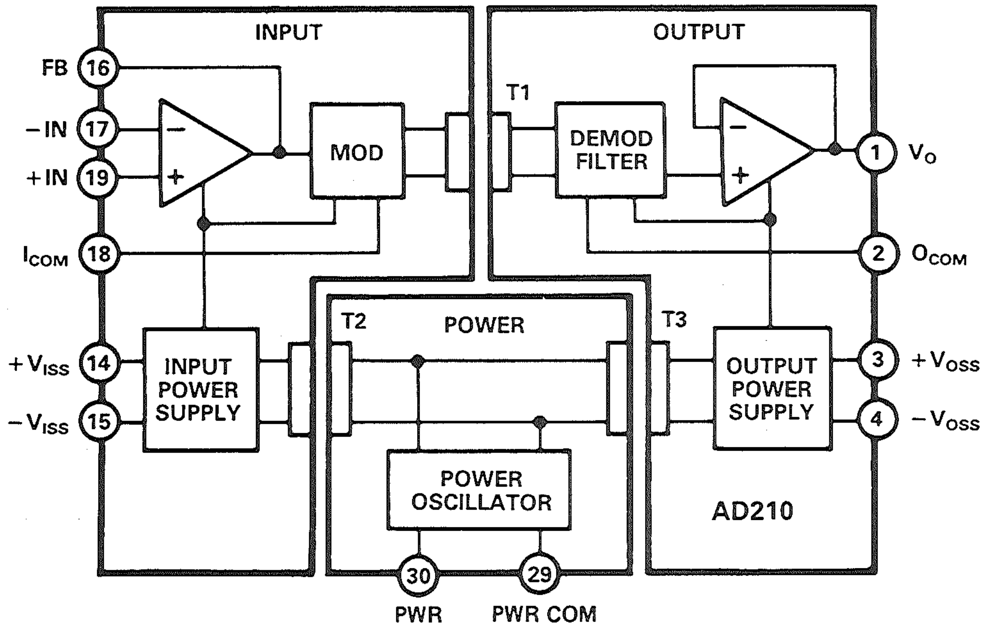


Figure 1.96

The AD210 is a 3-port isolation amplifier: the power circuitry is isolated from both the input and the output stages and may therefore be connected to either - or to neither. It uses transformer isolation to achieve 3500 V isolation with 12-bit accuracy. Key specifications for the AD210 are summarized in Figure 1.97.

AD210 ISOLATION AMPLIFIER KEY FEATURES

- Transformer Coupled
- High Common-Mode Voltage Isolation:

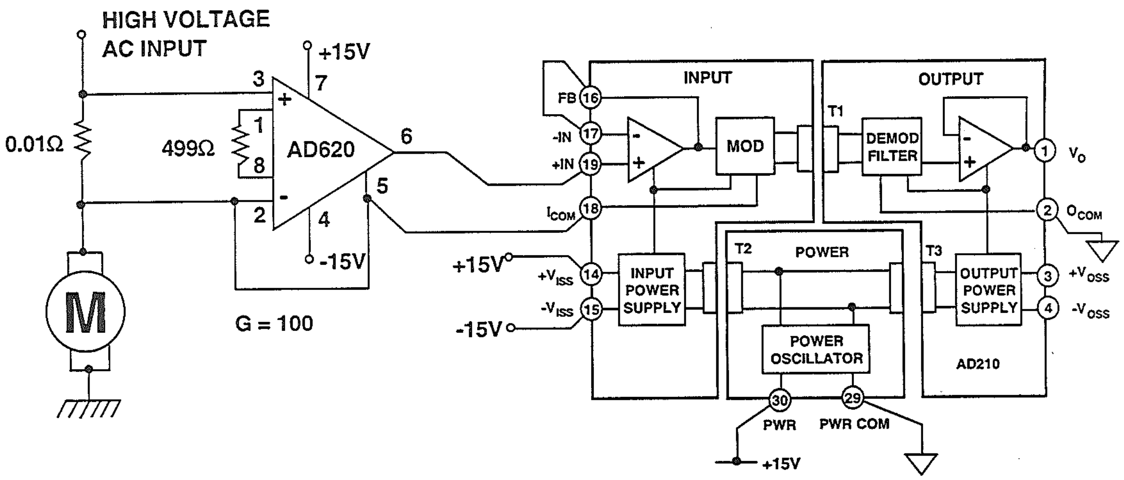
  2500V RMS Continuous
  ± 3500V Peak Continuous

- Wide Bandwidth: 20kHz (Full-Power)
- ± 0.012% Maximum Non-Linearity
- Input Amplifier: Gain 1 to 100
- Isolated Input and Output Power Supplies, ± 15V @ ± 5mA

Figure 1.97

A typical isolation amplifier application using the AD210 is shown in Figure 1.98. The AD210 is used with an AD620 instrumentation amplifier in a current-sensing system for motor control. The input of the AD210, being isolated, can be connected to a 110 or 230 V power line without any protection, and the isolated ±15 V powers the AD620, which senses the voltage drop in a small current sensing resistor. The 110 or 230 V rms common-mode voltage is ignored by the isolated system. The AD620 is used to improve system accuracy: the V_{OS} of the AD210 is 15 mV, while the AD620 has V_{OS} of 30 µV and correspondingly lower drift. If higher DC offset and drift are acceptable the AD620 may be omitted and the AD210 used directly at a closed loop gain of 100.

MOTOR CONTROL CURRENT SENSE



- High Accuracy/Low-Drift of AD620
- AD620 Powered By AD210
- Floating, Isolated, Senses Up To 2000V

Figure 1.98

ADI's range of transformer coupled isolation amplifiers is listed in Figure 1.99. There are a variety of devices with differing accuracies, bandwidths and breakdown voltages (as high as 3.5 kV), and, using transformer isolation, all of them have isolated power supplies, which simplifies system design.

ISOLATION AMPLIFIER FAMILY

Part	Peak Isolation Voltage	Gain Range V/V	Gain Nonlinearity % Maximum	Frequency Response, kHz
AD202	1000 - 2000	1 - 100	0.025 - 0.05	2
AD203	2000	1 - 100	0.025	10
AD204	1000 - 2000	1 - 100	0.025 - 0.05	5
AD208	1000 - 2000	1 - 1000	0.015 - 0.03	0.4 - 4
AD210	3500	1 - 100	0.012 - 0.025	20

Figure 1.99

Optical isolators using optical fibers may be made with isolation voltages of tens of MV. More general purpose devices consist of an LED and a photocell, electrically isolated but in a single DIL package, and having a breakdown voltage in the range 3.5 - 10 kV. The coupling between the two elements is not linear, so they cannot be used in simple analog isolation amplifiers, although they will carry digital signals and hence the results of A-D or V-F conversion very efficiently.

FOR HIGHER VOLTAGE BARRIERS USE OPTO-ISOLATORS

- Uses Light for Transmission Over a High Voltage Barrier
- An LED is the Transmitter, and a Photodiode or Phototransistor is the Receiver
- High Voltage Isolation is in the Range of 5000V to 7000V
- Usually Not Linear -- Best for Digital or Frequency Information

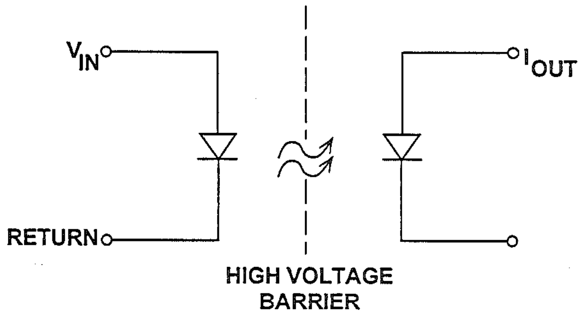


Figure 1.100

The Siemens IL300XC is an opto-coupler which contains two carefully matched photodetectors (they are actually matched at a sensitivity ratio of 0.7:1). By using one in a feedback system it is possible to linearize the response of the other when it is used as an isolation element. The linearity of such a system depends on accuracy of the matching of the photodetectors, and an optical design which ensures that both photodetectors receive the same amount of light from the LED.

Figure 1.101 shows an isolated temperature measuring scheme using such a device. The TMP-01 has a voltage output of 5 mV/K. Between -40° and +85°C its output is 1.16 - 1.8 V. This output voltage is applied to R1, and negative feedback in the OP-90 forces a current through the LED of the IL300XC such that a current flows in detector D1 to force pin 2 of the OP-90 to 2.5 V. Since D1 and D2 are matched at a ratio of 0.7:1 and operating in similar circuitry, 0.7 times this current will flow in D2 and, as resistor R1 is 0.7 times the feedback resistor R2 (which may be trimmed for exact calibration), the output of the second OP-90 will be equal to the output of the TMP-01. The circuit is simple and uses a single +5 V supply. The TMP-01 may be replaced with other voltage sources <2.5 V.

HIGH VOLTAGE ISOLATION USING SINGLE-SUPPLY

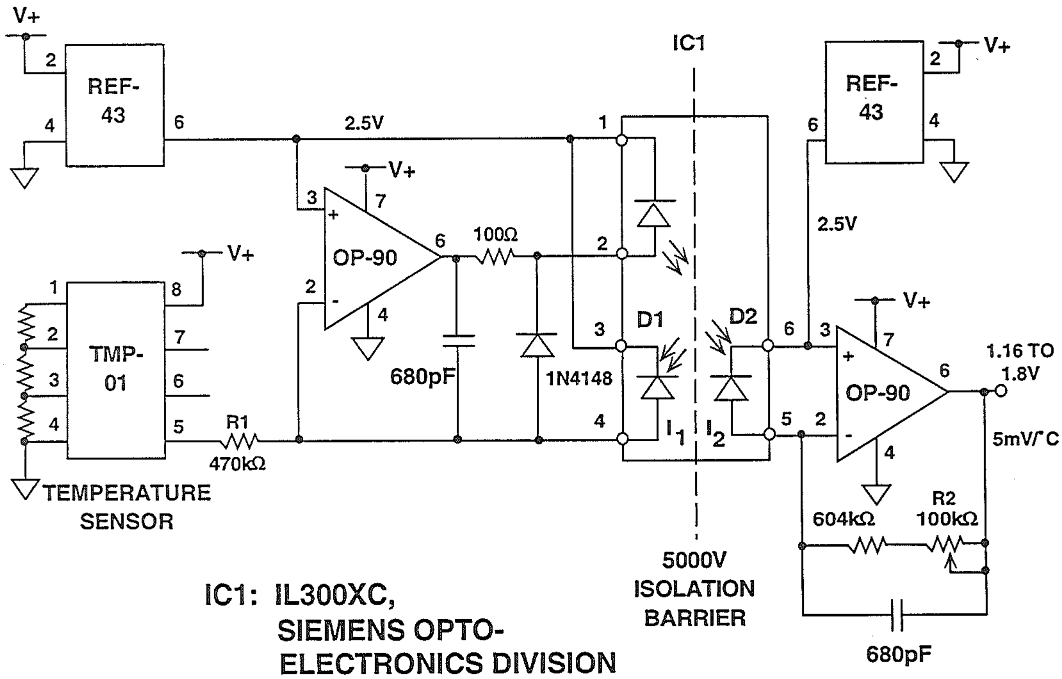


Figure 1.101

Voltage-Frequency Converters (VFCs) are valuable in systems requiring isolation. The analog signal is converted to a frequency in a VFC and may then be transmitted across an isolation barrier by very simple means with no risk of non-linearity and greatly reduced susceptibility to noise. At the receiver there are two options: the signal may be applied to a counter for a fixed period and the count read by a digital processor, the combination of VFC and counter acting as an ADC, or the frequency may be converted back to a voltage in a Frequency-Voltage Converter (FVC). FVCs are quite simple circuits and may generally be built with VFC ICs. (Reference 19)

The AD650 is a charge-balance VFC which is well suited for use in isolated VFC applications (Figure 1.103). The input to the AD650 charges a capacitor in an integrator. When the integrator output pass a threshold it fires a precision monostable (one-shot) which switches an accurate 1 mA current source into the integrator for a precisely-timed interval, removing a precise amount of charge from the integrator. It is obvious that the rate at which the monostable fires (which is the output frequency) is proportional to the current flows into the integrator, and hence to the voltage at the input terminal.

ISOLATION USING VOLTAGE-TO-FREQUENCY (V/F) CONVERTERS

- Converts A DC Voltage to a Frequency
- Frequency Information is Immune to Offsets and Noise
- V/F Output Can Be Transmitted Over Non-Linear Transmission Media
- Long Distance Transmission Over Twisted Pair or Fiber Optic Links
- A Frequency-to-Voltage (F/V) Converteris Used as the Receiver
- Accuracy is Determined by the V/F and F/V Conversion Process

Figure 1.102

THE AD650 CONNECTED AS A V/F CONVERTER

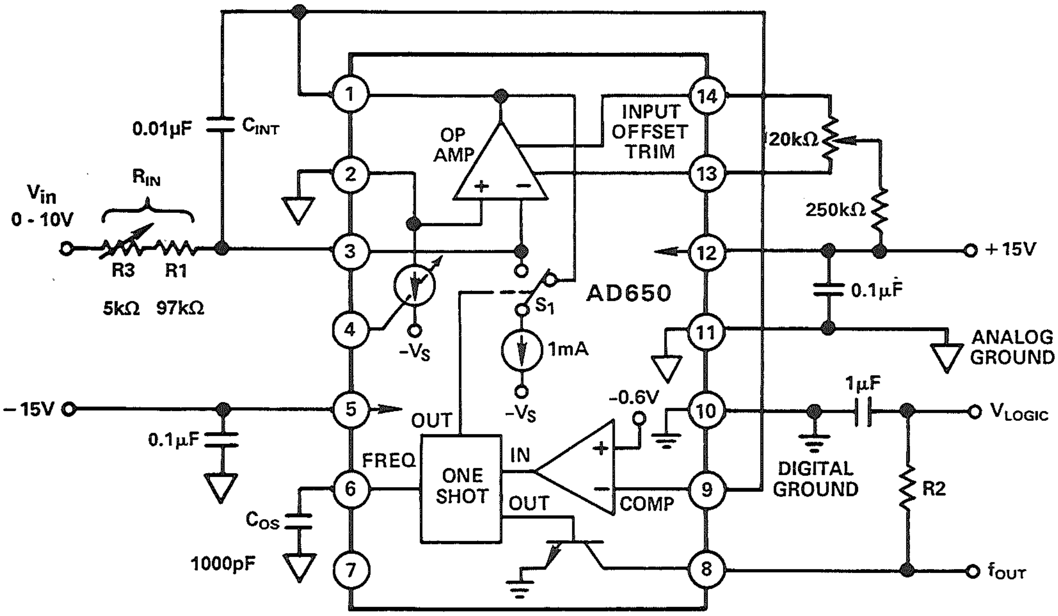


Figure 1.103

If the integrator of an AD650 is configured as a "leaky integrator" by connecting a resistor in parallel with the integration capacitor, and a pulse train is applied to the comparator to fire the precision monostable then the circuit will behave as an FVC: the more often the monostable fires the faster charge must leak out of the integration capacitor, and hence the larger the voltage on the integrator (see Figure 1.104).

THE AD650 CONNECTED AS AN F/V CONVERTER

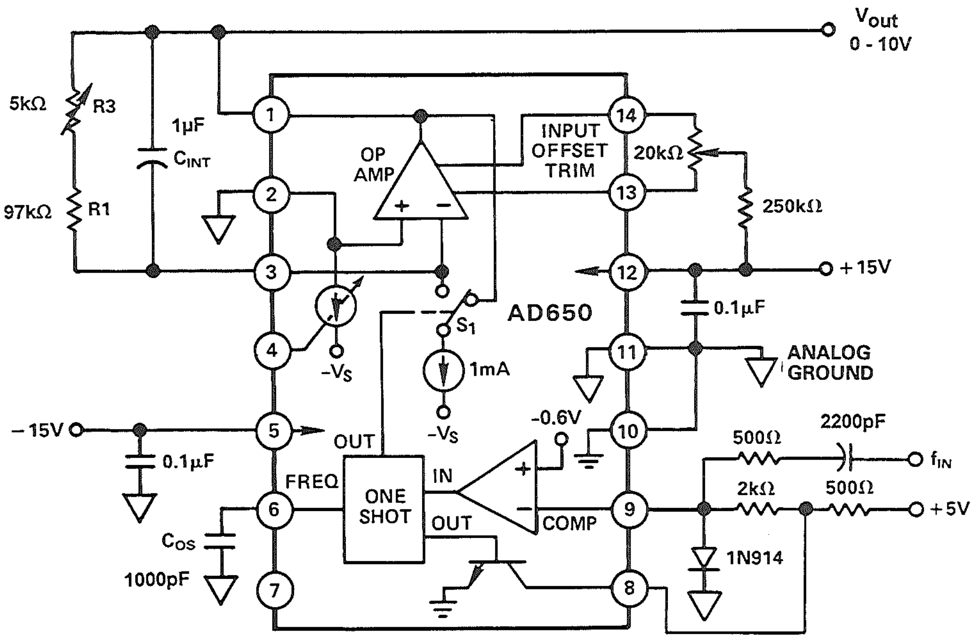


Figure 1.104

Such an FVC is very accurate at DC, but circuit constraints prevent independent adjustment of different time constants, and it is not very practical for demodulating VFC signals with wideband modulation. For such an application a phase-locked loop, using a VFC as an oscillator for linearity (this is very important), allows the design of much more flexible systems (Reference 19). Nevertheless a simple system using two AD650 devices and a 4N26, a common digital opto-isolator, can make a very efficient and accurate analog isolator (Figure 1.105).

AD650 IN A V/F, F/V OPTO-ISOLATED SYSTEM

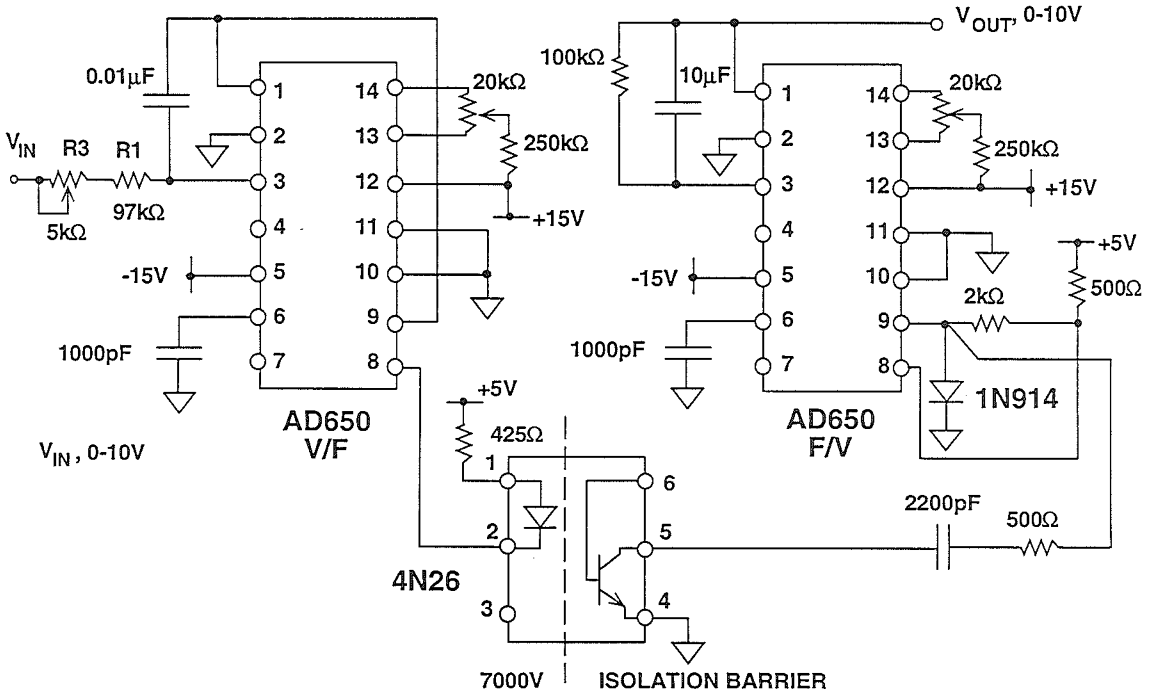


Figure 1.105

The accuracy of this circuit depends on the accuracy of the VFC and FVC conversions - the opto-isolator is not involved. Figure 1.106 shows the measured linearity of a prototype of the circuit, which is 0.006%, or better than 12-bits.

V/F, F/V LINEARITY

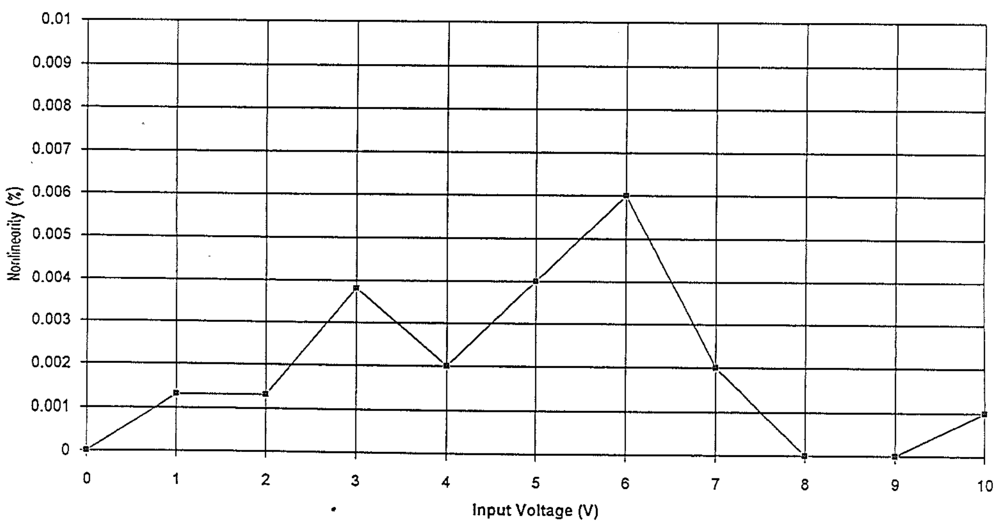
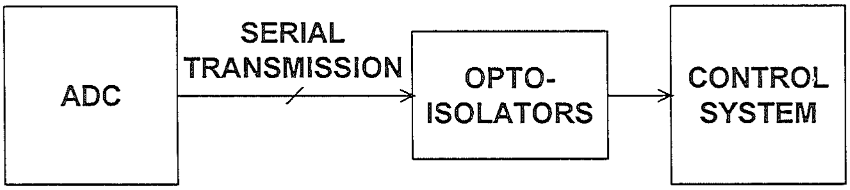


Figure 1.106

Despite the preceding discussion of analog isolation techniques, there is no doubt that the most accurate technique is analog to digital conversion *before* transmission across the isolation barrier (VFC-FVC is, after all, one version of this technique). If the signal will eventually be required in digital form the technique is particularly attractive, especially as there is no serious limit on the distance the data may be transmitted (Figure 1.107).

FOR HIGH ACCURACY ISOLATION, DIGITIZE FIRST



- Accuracy Limited Only By ADC
- Digital Transmission Relatively Immune To Noise

Figure 1.107

In the past the limitations of ADCs may have discouraged this approach, but inexpensive modern high-resolution, low-power ADCs with serial data output and cheap digital opto-isolators may make the technique attractive even when the output signal required is analog and a digital-analog conversion is necessary after transmission.

Despite evolution in the techniques available the need for galvanic isolation will remain for the foreseeable future. Each technique has its advantages and disadvantages and engineers must choose the most appropriate for the particular application.

ISOLATING TRANSDUCERS SUMMARY

TECHNOLOGY	ADVANTAGES	DISADVANTAGES
AD20x-Family Transformer Coupled	Fully Self-Contained 3-Port Isolation High Linearity Wide Bandwidth (20kHz)	Lower Breakdown Voltage
Opto-Isolators	High Voltage Immunity	Separate Power Needed External Amplifiers Required Poor Linearity
V/F and F/V Isolation	High Linearity Long Distance Transmission High Noise Immunity	Separate Power Needed DC Inputs Only
Digitize First	Highest Accuracy Best Linearity High Noise Immunity Long Distance Transmission	Separate Power Needed

Figure 1.108

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