AN EMC DESIGNER'S GUIDE TO OPERATIONAL AMPLIFIERS

While the capabilities and limitations of op-amps can be fully defined for normal operations, their specific characteristics relevant to TEMPEST/EMC require serious consideration.

BRUCE C. GABRIELSON AND ROBERT L. WALTER IV, PE SYSTEMS, INC., ALEXANDRIA, VA

INTRODUCTION

Operational amplifiers (op-amps) are popular, easily used and widely applicable IC amplifiers. In addition to their capabilities and cost advantages, the unique value of op-amps is that their capabilities and limitations in normal applications can be fully defined by a few simple equations and rules. However, from the perspective of an EMC engineer, a multitude of conditions may require serious consideration.

This article is divided into two main device level areas — noise generation and susceptibility. Internal noise sources, regenerated noise, isolation, and circuit applications are discussed.

Intrinsic noise is that noise which is characteristic of the semiconductor material itself rather than the content of any impurities it contains. All noise power from each of the various intrinsic noise sources is additive. There are three types of intrinsic op-amp noise prevalent at low frequency, with one type (flicker) dominant below 100 Hz.

Flicker noise in a semiconductor is also called 1/f, excess, or modulation noise. In a resistor it is called contact noise. Flicker noise exhibits a $1/f^n$ power spectrum where n can vary as $.8<n<1.5$. Flicker noise density e_n is defined as

$$e_n = Kf^{-n/2} \qquad \frac{\mu V}{\sqrt{Hz}}$$

where K is a constant dependent on the type of material and its geometry, and f is frequency in hertz.

Popcorn noise (burst) is generated at semiconductor junctions as a result of modulation of the barrier height caused by metallic impurities in the semiconductor. It is greatest in high impedance devices such as op-amp input circuits. Popcorn noise can be reduced by improving manufacturing techniques and reducing the metallic impurities of the semiconductor device, and is characterized by the constant amplitude of the individual bursts. However, the duration of the popcorn noise varies from microseconds to seconds and the repetition rate is not periodic. The amplitude is generally 2 to 100 times thermal noise. Popcorn noise exhibits a $1/f^n$ power spectrum, where n is generally 2, and appears to be related to flicker noise.

Pink noise is the term used to describe any noise that has a 1/f power spectrum. The total flicker noise in bandwidth B falling between frequencies f_1 and f_2 has been shown by Lindquist[1] to be $E_n = 1.52KD$, where D is the frequency separation between f_1 and f_2 in decades. Pink noise, therefore, increases 3dB for every doubling of decade separation.

Shot or ***Schottky noise*** is caused by the fluctuations in the rate of arrival of charge at semiconductor junctions. Junction current for Schottky noise can be calculated from:

$$I_n = [2qI_{dc}B]^{1/2} \text{ amps}$$

where
- I_{dc} = quiescent junction current (amps)
- B = device bandwidth, and
- q = election charge.

White noise (thermal) is dominant

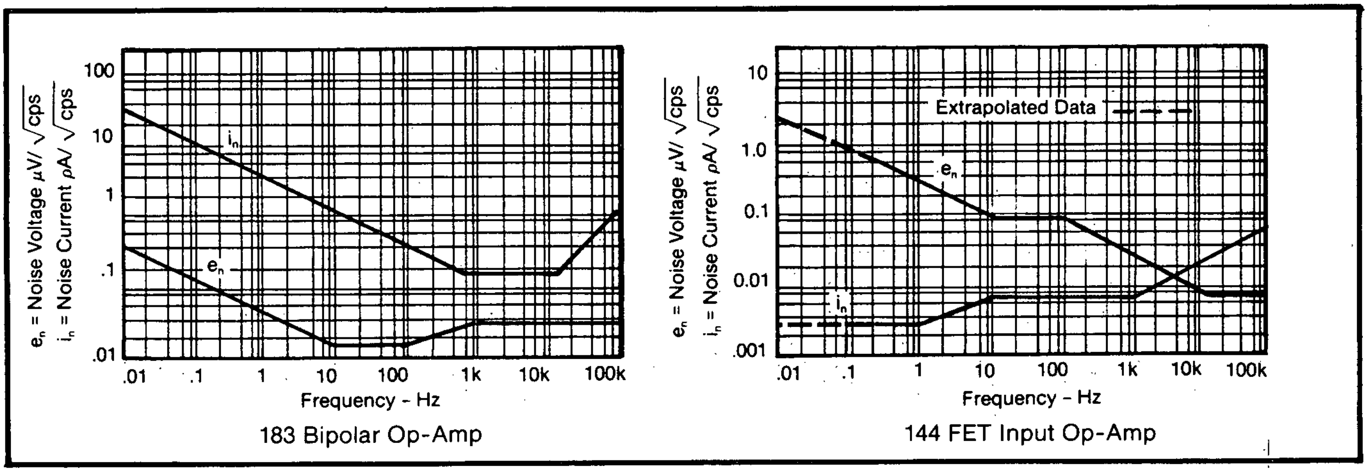

Figure 1. Voltage and Current Noise Densities.[1]

at frequencies above 100 Hz. White noise, also referred to as Johnson noise, is the result of resistances internal to the op-amp. Normal temperature variations have only a small effect on the value of thermal noise. For example, according to Ott, an increase in temperature from 17°C to 117°C produces only a 16% increase in thermal noise.[2]

The power spectrum for white noise is independent of frequency, and, therefore, much flatter than the other noise types. The power spectrum is calculated by

$$Sn(\Omega) = 2kTR \frac{V^2}{\text{rad/sec}}$$

where
k = 1.38×10^{-23} $\frac{\text{Joules}}{°K}$ (Boltzman's constant)

T = temperature in °K, and
R = resistance in ohms.

White noise affects the usable bandwidth of the op-amp since practical circuits do not exhibit uniform gain within their passband and exhibit zero gain outside. Internal resistance (R) within the op-amp creates an open circuit RMS noise voltage due to internal circuit biasing. This voltage can be calculated as follows.

$$E_n = [Sn(0)\, 2\pi B]^{1/2} = [4kTR_{int}\, B]^{1/2} \text{ volts}$$

where
R_{int} = op-amp internal resistance
B = op-amp Bandwidth in Hz
k = Boltzman's constant (1.38×10^{-23} $\frac{\text{Joules}}{°K}$)
T = temperature in °K
4kT = 1.6×10^{-20} W/Hz at room temperature

Figure 1 from Smith and Sheingold[3] gives current and noise densities for AD 183 and 144 op-amps. As noted by Lindquist (pg. 720), flicker noise (voltage) dominates from 10 to 80 Hz, and shot noise above 1 kHz. Looking at the current density curve, flicker noise dominates to 1 kHz, with shot noise at the noise floor between 600 Hz and 15 kHz.

The usable equivalent bandwidth B is directly related to the number of poles in the circuit passband, and limiting bandwidth reduces thermal noise. The noise bandwidth is always wider than the circuit's 3dB bandwidth. Ott[2] has calculated the ratio of the noise bandwidth B to the 3dB bandwidth for circuits with several poles. (See Table 1.)

Number of poles	B/fo	High frequency rolloff (dB per octave)
1	1.57	6
2	1.22	12
3	1.15	18
4	1.13	24
5	1.11	30

Table 1. Ratio of the Noise Bandwidth B to the 3dB Bandwidth f_o.

As mentioned previously, the noise power from each intrinsic noise source is additive. The noise is incorporated as a part of the total noise power per unit bandwidth,

$$\frac{\vartheta_t^2}{R} \qquad \frac{VA}{\sqrt{Hz}}$$

and includes the 4kT thermal noise plus all other internal noise sources, the

$$\frac{\vartheta_n^2}{R} \qquad \frac{VA}{\sqrt{Hz}}$$

noise contributed by the input noise voltage, ϑ_n of the op-amp, and the Ri_n^2 noise power contributed by input noise current (i_n) of the op-amp. If a frequency greater than 100 Hz is considered, thermal noise dominates, and the noise power per unit bandwidth for an entire op-amp gain block equation becomes

$$\frac{\vartheta_t^2}{R} = \frac{P_B}{B} = 4kT + \frac{\vartheta_n^2}{R} + Ri_n^2 + \frac{E_n^2}{R_{int}}$$

where ϑ_t and ϑ_n are in $\frac{V}{\sqrt{Hz}}$.

i_n is in $\frac{A}{\sqrt{Hz}}$

$$4kT = 1.6 \times 10^{-20} \frac{VA}{Hz}$$

at room temperature.

If ϑ_n and i_n are constant within the bandwidth of interest, the resulting noise power (P_B) can be reduced to

$$P_B = B\frac{\vartheta_t^2}{R} + B\frac{E_n^2}{R_{int}}. \qquad VA$$

Barna[4] adds that if ϑ_n and i_n are not constant, the resulting wideband noise must be calculated by integrating over the bandwidth the product of the amplitude and the input noise power per unit bandwidth.

Noise power resulting from the external op-amp circuit was determined using the following gain block. It is important to point out that regardless of whether the circuit is inverting or non-inverting, the dc return paths to ground for the two inputs must be of equal or finite impedance for amplifiers that require a significant amount of input bias current.

A typical inverting op-amp circuit is shown in Figure 2[2]. Gain for this circuit is

$$A = \frac{R_F}{R_I}. \quad \text{(inverting).}$$

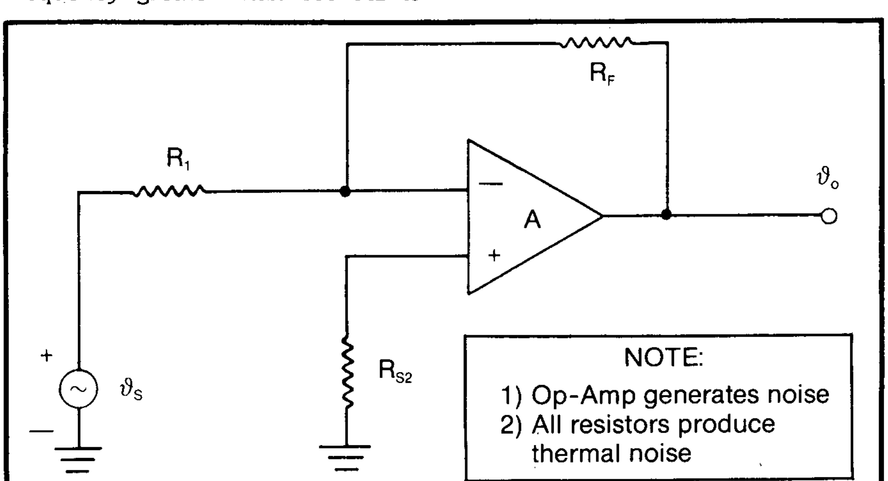

Figure 2. Typical Inverting Op- Amp Circuit.

This op-amp circuit can be reconfigured to the noiseless op-amp circuit of Figure 3 by combining R_F and R_1, and incorporating all noise sources externally. Only thermal noise above 100 Hz generated external to the op-amp has been included in this example.

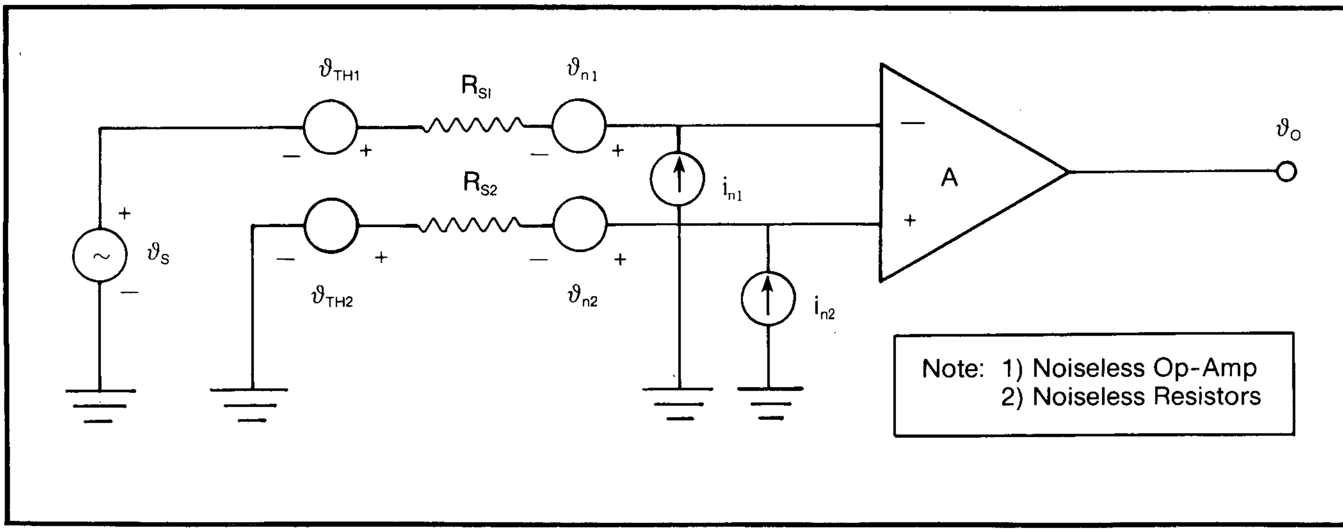

Figure 3. Noiseless Op-Amp Circuit.

ϑ_{TH_1} and ϑ_{TH_2} are noise voltages generated by the equivalent noiseless external resistors R_{S1} and R_{S2}, where

$$R_{S1} = \frac{R_1 + R_F}{R_1 + R_F} \text{ (noiseless)},$$

$$R_{S2} = R_2 \qquad \text{(noiseless)},$$

and the thermal noise generated by R_{S1} and R_{S2} is

$$\vartheta_{TH_1} = \sqrt{4kTBR_{S1}}$$
$$\vartheta_{TH_2} = \sqrt{4kTBR_{S2}}.$$

The total input noise to an op-amp circuit (gain block) can now be calculated specifically by adding each contribution shown in Figure 3.

POWER SUPPLY NOISE REJECTION

Noise coupled to the power supply lines of high Q filters is transformed by the B+ rejection of an amplifier. The Power Supply Rejection Ratio (PSRR) is defined as the ratio of the equivalent voltage change at the input of the amplifier to a change in the supply voltage on the B+ input.

$$\text{PSRR} = \left| \frac{1}{A_{DC}} \frac{dV_{OUT}}{dV_{SUPPLY}} \right|$$

where A_{DC} is the op-amp gain at dc.

The lowest (best) PSRR occurs at dc, and deteriorates with increasing frequency of ripple. Compensation also can affect PSRR by preventing some internal transistors from operating in a fully on or off condition, or by changing the bias of internal small signal amplifiers. PSRR decreases at about 20dB/decade. This indicates that high frequency powerline noise can affect op-amp performance and stability. Therefore, power supply inputs require protection. For example, if 0.1V of signal appears on the power supply input at the center frequency of a high Q bandpass filter, with a low PSRR of 40dB, and the filter gain is 68dB, the total B+ amplification at this frequency would be 28dB. Therefore, 0.1 volts at the B+ terminal would produce more than 2 volts at the output.

RC decoupling is a commonly used method to filter out high frequency wideband noise. However, since load current changes may modulate the voltage at the amplifier's input, a preferred approach would be to use ferrite beads.

Ferrite beads attenuate EMI signals without affecting dc or low frequency signals on dc lines. The bead absorbs noise power and dissipates the power in the form of heat. Since the inclusion of a bead effectively increases line impedance to higher frequencies, the combination of a bead with a capacitor would provide a damping value equal to the bead impedance multiplied by the capacitive reactance. The op-amp bandwidth and current requirements will determine the allowable RC time constant for the powerline filter.

A tendency for mistaking internally generated low frequency op-amp noise for inadequate power supply regulation is common especially when using externally compensated devices, which are subject to the input noise ϑ_n previously discussed. Therefore, in high closed loop gain applications, internally compensated op-amps have lower PSRRs.

COMMON MODE NOISE REJECTION

Theoretically, if equal amounts of RF energy enter both inputs of an op-amp simultaneously, equal rectification and equal offset will occur at each input terminal. The common mode noise appears as the offset voltage between each input and the op-amp ground. Transverse mode noise would occur if the inputs were unequal. For an ideal amplifier $\vartheta_{out} = A\vartheta_d$ as shown in Figure 4.

In reality, a small fraction of ϑ_c will effect ϑ_{OUT} as

$$\vartheta_{out} = A\vartheta_d + A_{CM}\,\vartheta_C$$
$$\left| A_{CM} \right| << \left| A \right| .$$

The Common Mode Rejection Ratio, CMRR, is defined as

$$\text{CMRR} = 20 \log \left| \frac{A}{A_{CM}} \right| .$$

In general, CMRR is the greatest at dc.

The CMRR is commonly determined when using feedback amplifier circuits, particularly when using differentiated amplifiers, as shown in Figure 5. In this case, an inverting configuration is used.

Arpad Barna[4] has shown that if

$$\frac{R_S}{R_P} \neq \frac{R_I}{R_F} \text{ and } A_{CM} \approx 0,$$

CMRR can be calculated as

$$\mathrm{CMRR} = 20\mathrm{dB}\left| \frac{\log\left(1 + \frac{R_F}{R_I}\right)}{\left(1 + \frac{R_S R_F}{R_P R_I}\right)} \right|$$

CMRRs are difficult to calculate, and should be measured where practical.

ACTIVE FILTER NOISE

One of the primary applications of op-amps is in active filter design. Even though the filter design calculations appear simple, they seldom consider the real physical properties of the active components used in the filter synthesis. As shown previously, the total noise power associated with the gain block is equal to the sum of the noise sources in the measurement bandwidth. In other words, if $\Delta\vartheta_{1n}$ is the input noise voltage due to the three port network, and if $\Delta\vartheta_{1a}$ is the noise voltage generated within the op-amp, then the total rms noise present is

$$\Delta\vartheta_1 = (\Delta\vartheta_{1n}{}^2 + \Delta\vartheta_{1a}{}^2)^{1/2}$$

where $\Delta\vartheta_{1a}$ is dependent on e_n and i_n.

The figure-of-merit for filters is the signal-to-noise ratio, SNR. SNR is defined in dB

SNR = 20 Log (RMS Signal/RMS Noise).

In terms of the noise voltage, SNR becomes

$$\mathrm{SNR} = 20\,\mathrm{Log}\,[\vartheta_1 / (\Delta\vartheta_{1n}{}^2 + \Delta\vartheta_{1a}{}^2)^{1/2}].$$

This equation is significant because it shows that the SNR of a filter is independent of op-amp gain. As will be discussed shortly, if a real filter is used as part of a circuit, gain does have some effect when noise is present.

An alternate figure-of-merit is sometimes used to describe filters. This is the noise figure, NF, a measure of the relative levels of $\Delta\vartheta_{1a}$ and $\Delta\vartheta_{1n}$. The noise figure is defined as

$$\mathrm{NF} = \mathrm{SNR}_0 - \mathrm{SNR}_I \text{ in dB}$$

where SNR_I = input SNR independent of internal noise.
SNR_0 = output SNR dependent on e_n and i_n.

When the filter is used as part of a circuit, coupled noise presented to the filter will be treated like any other signal, and may be amplified. The phenomenon is most apparent when using high Q filters. In designs that use a single filter to generate two poles, the amplification factor of noise is proportional to the square of the filter Q. An improvement can be made if a biquad or state variable filter is used that has relative noise amplifications proportional to Q.

NOISE SOURCES AND OP-AMP SUSCEPTIBILITY

There are three major sources of noise which affect op-amp susceptibility: intrinsic, man-made, and natural. Intrinsic noise was discussed previously. Man-made noise is the noise generated by transmitters, motors, etc. Natural noise is the result of lightning or sun spots. Each of these sources can lead to circuit level susceptibility through conducted or radiated paths. Figure 6 describes noise sources over various frequency ranges that can affect op-amp performance.[5]

Op-amps are the most common type of linear integrated circuit. Due to their high input impedance and their linear high gain nature, they may also be the most susceptible IC to RF energy, conducted or radiated, and

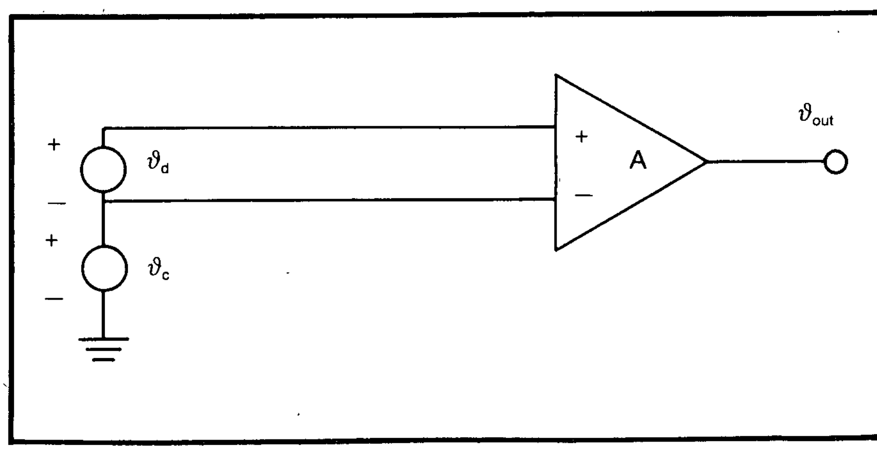

Figure 4. Circuit to Evaluate CMRR.

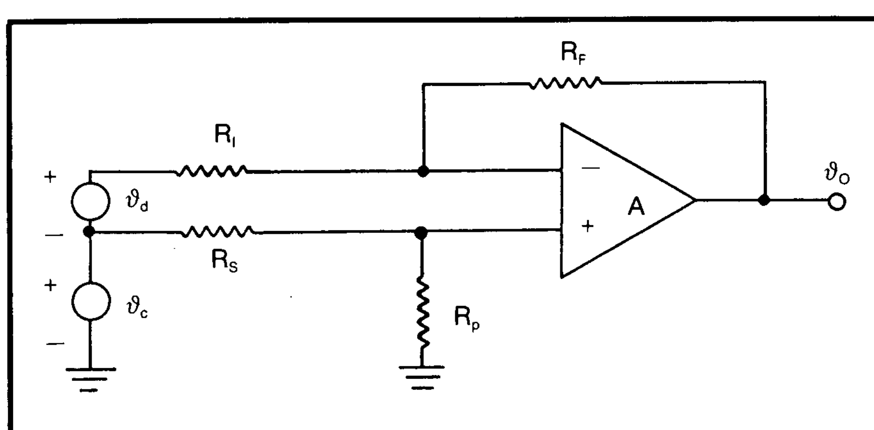

Figure 5. Equivalent Circuit for Determining Diff Amp CMRR.

Noise sources external to the op amp

Source	Nature	Causes	Minimization methods
60-Hz Power	Repetitive Interference	Power lines physically close to op-amp inputs. Poor CMRR at 60 Hz. Power transformer primary-to-secondary capacitive coupling.	Reorientation of power wiring. Shielded transformers. Single point grounding. Battery power.
120-Hz Ripple	Repetitive	Full-wave rectifier ripple on op-amp's supply terminals. Inadequate ripple consideration. Poor PSRR at 120 Hz.	Thorough design to minimize ripple. RC decoupling at the op amp. Battery power.
180-Hz	Repetitive EMI	180 Hz radiated from saturated 60-Hz transformers.	Physical reorientation of components. Shielding. Battery power.
Radio stations	Standard broadcast AM through FM	Antenna action anywhere in system.	Shielding. Output filtering. Limited circuit bandwidth.
Relay and switch arcing	High frequency burst at switching rate	Proximity to amplifier inputs, power lines, compensation terminals or nulling terminals.	Filtering of hf components. Shielding. Avoidance of ground loops. Arc suppressors at switching source.
Printed-circuit-board contamination	Random low frequency	Dirty boards or sockets.	Thorough cleaning at time of soldering, followed by a bake-out and humidity sealant.
Radar transmitters	High frequency gated at radar pulse repetition rate	Radar transmitters, from long-range surface search to short-range navigational—especially near airports.	Shielding. Output filtering of frequencies >> PRR.
Mechanical vibration	Random > 100 Hz	Loose connections, intermittent metallic contact in mobile equipment.	Attention to connectors and cable conditions. Shock mounting in severe environments.
Chopper frequency noise	Common-mode input current at chopping frequency	Abnormally high-noise chopper amplifier in system.	Use of balanced source resistors, bipolar input op amps instead of a chopper amplifier or use of a premium low noise chopper amplifier.

Table 2. Common Noise Sources and Protection Techniques.[5]

coupled to either of the op-amp input terminals. The RF energy appears as an offset voltage at the device's inverting or noninverting input. The RF equivalent source impedance, frequency, power level, and the characteristics of the block will determine the resulting op-amp susceptibility. Additionally, some device types, such as CMOS, are less susceptible than other types. Table 2 describes common noise sources and protection techniques employed on operational amplifier circuits.[5]

The McDonnell Douglas Astronautics Company report (MDC E1929) presents a comprehensive examination of op-amp susceptibility.[6] This report describes the coupled levels necessary to cause either operational failures (output changes) or device failures (catastrophic). Operational failures are relevant to system tolerances, while device failures were identified as being related to three specific types, all thermal in nature. These failure types included bond wire, junction, and metallization failures, and are due to heat resulting from I^2R power dissipation. RF power levels and pulse duration are the most significant factors. The report noted that circuits using high signal levels and high operating currents experienced less than expected upset to unwanted rectified signals.

Packaging styles were also evaluated. Measurement results indicate no significant differences were apparent between the absorptive or reflective loss properties of the different package styles — flat pack or TO-5. Additionally, using lossy material in DIP packages provided little or no protection at around 220 MHz where devices are typically most susceptible.

Figure 7 describes the worst case susceptibility level of an op-amp obtained by applying an offset voltage at ϑ_{11}[6]. The output voltage is related to the offset voltage by

$$\vartheta_{OUT} = -\vartheta_{IN}\frac{R_F}{R_1} - \vartheta_{11}\left(\frac{R_F}{R_{IN}}+1\right).$$

According to the MDC report, for offsets less than .05 volts, the offset voltage is approximately proportional to the minimum RF power level, while the offset voltage is approximately

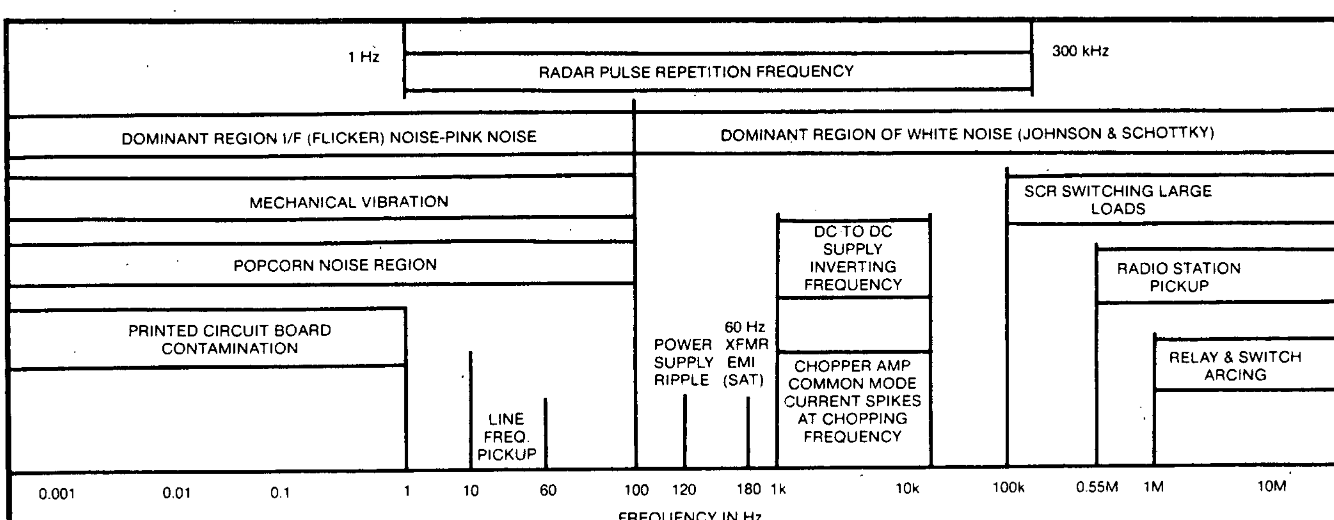

Figure 6. Noise Sources Which Perform Op-Amp Performance.[2]

proportional to the square root of the RF power level for offsets greater than 0.20 volts.

PUBLISHED PARAMETERS

Not all the parameters related to op-amp performance described in this article are supplied in the manufacturers' specification sheets. Often, a series of curves related to various parameters over operating frequencies or supply voltages are the only information provided. Two typical curves are shown in Figure 8.[7]

One subject not yet discussed is settling time. Figure 9 from Burr-Brown describes settling time, defined as the total time required, after an input step signal, for the output to settle within a specified error band (normally .01% of the input) around the final output value.[7] This value is important from an EMI viewpoint, since even though the voltage overshoot has little effect on op-amp performance, it is continually a source of noise which affects the performance of other nearby circuitry.

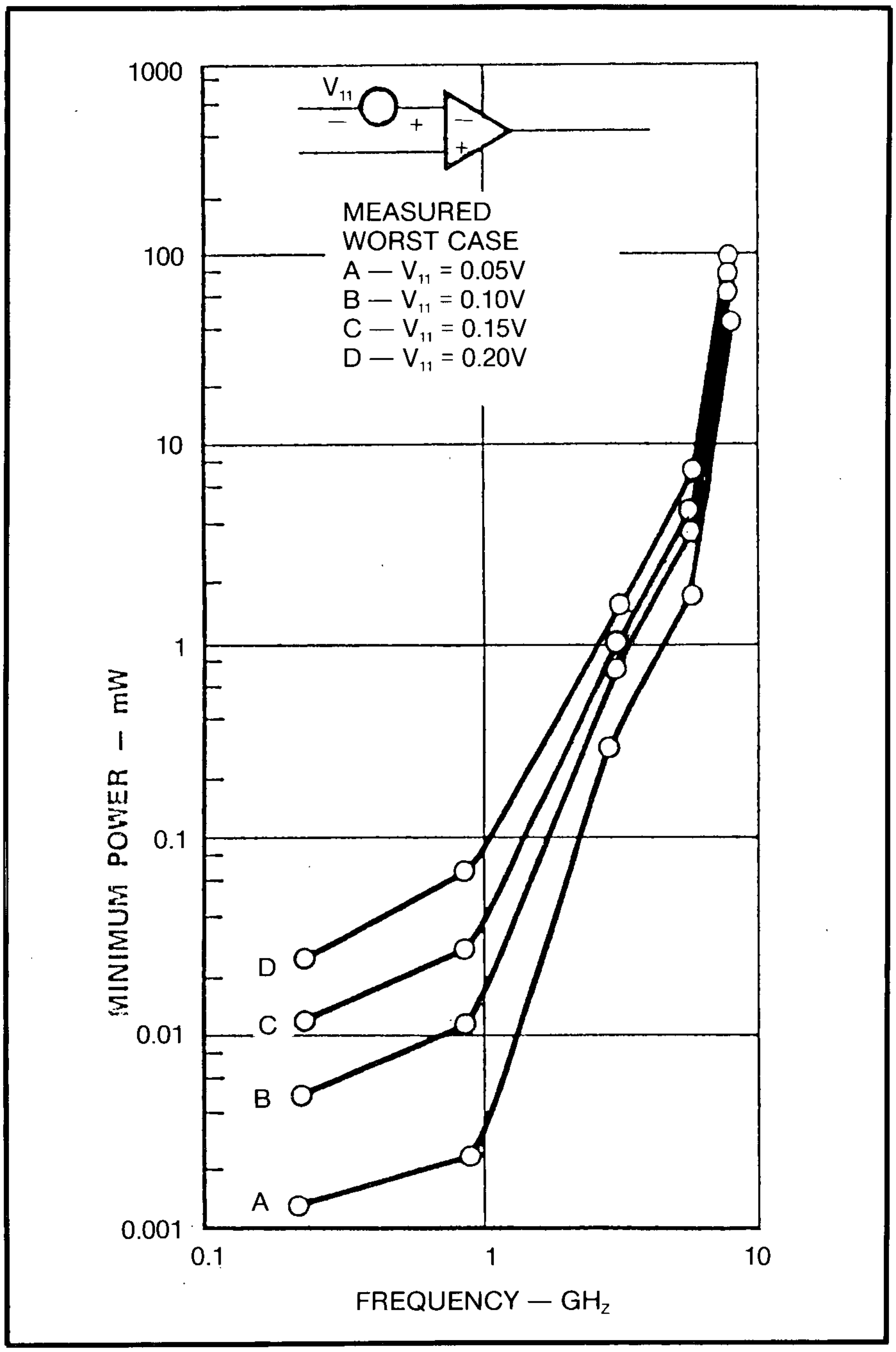

Figure 7. Worst Case Susceptibility Values for Op-Amps.[6]

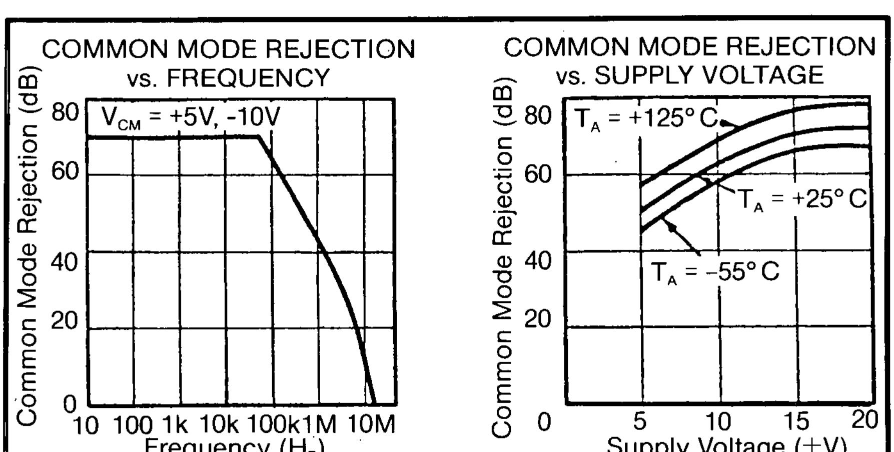

Figure 8. Typical Op-Amp Performance Curves.[7]

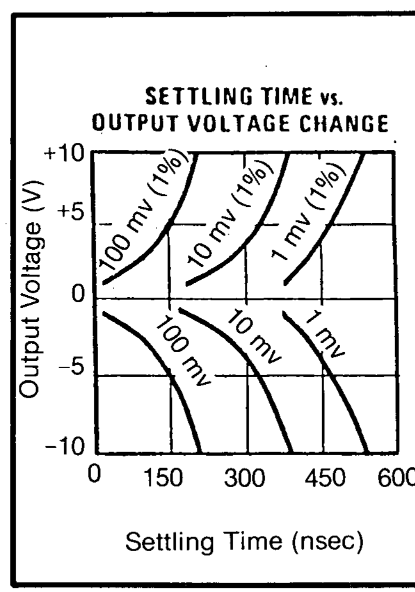

Figure 9. Typical Settling Time Curve.[7]

Table 3 describes some typical characteristics for various op-amp types.[5]

CONCLUSION

As the most common type of linear integrated circuit, op-amps are widely

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Type	107	702A	741	2741	9406
Description	Internally compensated, Monolithic	Monolithic	Internally compensated, Monolithic	Field-effect transistor input, Hybrid	Internally compensated, Hybrid
DC Amplification	160,000	4000	200,000	30,000	1000
Maximum input bias current	75 nA	5 μA	0.5 μA	100 pA	10 μA
Maximum input offset current	10 nA	0.5 μA	0.2 μA	50 pA	1 μA
Maximum input offset voltage	2 mV	2 mV	5 mV	5 mV	10 mV
Maximum temperature coefficient of input offset voltage	15 μV/°C	10 μV/°C	30 μV/°C	25 μV/°C	100 μV/°C
DC common mode rejection ratio	96 dB	100 dB	90 dB	70 dB	80 dB
Differential input resistance	4 MΩ	40 kΩ	2 MΩ	100 GΩ	7 kΩ
Supply voltage rejection ratio	15 μV/V	75 μV/V	30 μV/V	100 μV/V	3 mV/V
Corner frequencies	5 Hz 10 MHz	1 MHz 4 MHz 40 MHz	10 Hz 10 MHz	10 Hz 1 MHz	1.5 MHz 150 MHz
Slew rate	0.5 V/μs	50 V/μs	0.5 V/μs	5 V/μs	360 V/μs

Table 3. Typical Op-Amp Characteristics.[2]

used in a variety of applications where TEMPEST and EMC considerations are important. A thorough understanding of the entire spectrum of performance characteristics is essential for proper applications during the design phase of a program. This article should help in providing the necessary background to meet the challenge posed by this interesting integrated circuit. ■

REFERENCES

1. Lindquist, Claude S. "Active Network Design with Signal Filtering Applications." Long Beech, CA: Steward and Sons, 1977.
2. Ott, Henry W. "Noise Reduction Techniques in Electronic Systems." New York: Wiley Interscience, 1976.
3. Smith, L. and Scheingold, D.H. "Noise and Operational Amplifier Circuits." Analog Dialog, Volume 3, Number 1, Analog Devices, Cambridge, MA.
4. Barna, Arpad. "Operational Amplifiers." New York: Wiley-Interscience, 1971.
5. Electronic Design, "Keep Your Op-Amp Circuits Quiet." pp. 88-93: 27 September 1975.
6. "Integrated Circuit Electromagnetic Susceptibility Handbook." Report MDC E1929 McDonnell Douglas Astronautics Company, St. Louis, MO. 1 August 1978.
7. Burr-Brown. "1979 General Catalog." Tuscon, Arizona: Burr-Brown Research Corporation, 1978.