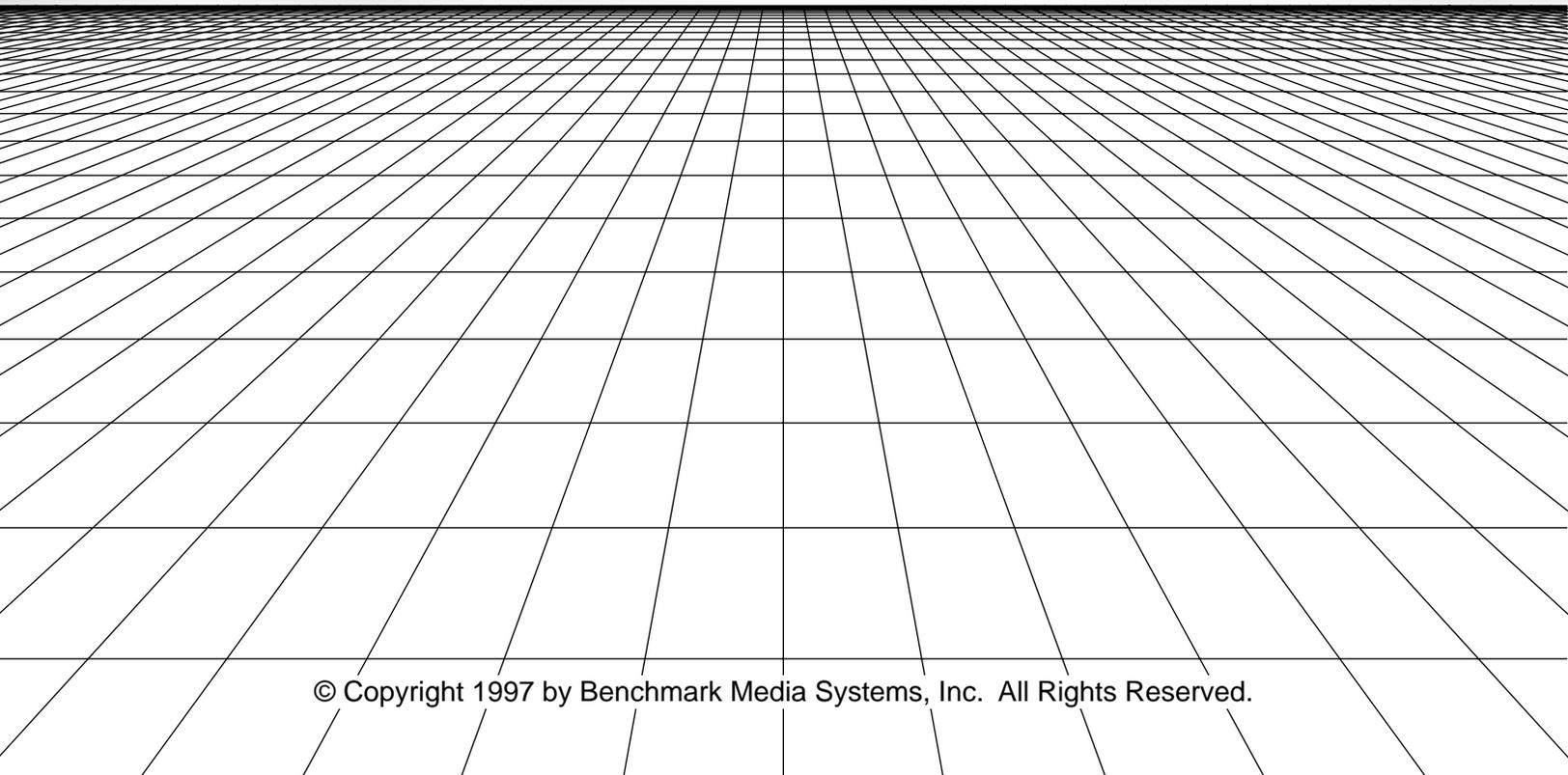
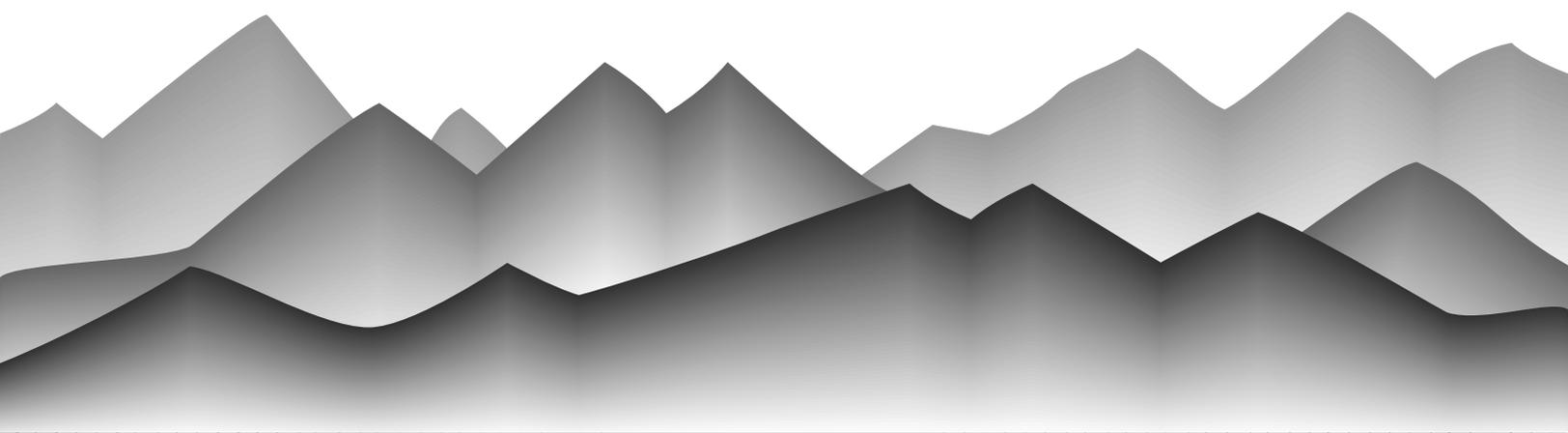


A Clean Audio Installation Guide

by Allen H. Burdick

A BENCHMARK MEDIA SYSTEMS, INC.
Application Note



A CLEAN AUDIO INSTALLATION GUIDE

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1.0 INTRODUCTION

The problems encountered in achieving clean audio are ones that each of us involved in audio must face with renewed interest. Initially, with the advent of inexpensive quality receiving equipment from Japan, the FM world was forced to “get it together”. Now with AM and TV stereo, HI-FI video tape, video disk systems in home theaters, the Compact Disk, R-DAT, Compact Digital Cassette, recordable disc systems, etc., creating an ever increasing public awareness of *very* high quality audio, we have no choice but to get it together as well!

Fortunately for us, the tools are available, and the price to be paid is usually that of time, some technical knowledge, occasionally some new test equipment, but most importantly, greater than average care with our installations. This paper will deal only with analog audio systems.

1.1 Overview

After choosing our quality audio equipment, there are five major areas, that must be properly addressed to achieve truly outstanding audio:

- 1) proper power systems installation;
- 2) inter-connection system methods;
- 3) RF immunity;
- 4) proper headroom and signal-to-noise ratios; and
- 5) correct operation within the limits of the system.

2.0 POWER SYSTEM INSTALLATION

If all of the lighting and equipment in this world ran on DC as Edison had wanted, then the only concerns that we as audio professionals would face with our installations would be those of RF immunity, very low noise, low distortion levels, and wide bandwidth in our equipment. As we all well know, that is not the case. Probably the largest single source of audio “interference” we face is that of power line related voltages being added to the desired signals within our systems. Most of this problem is a result of *normal* 50 or 60 Hz voltage differences that exist between the chassis of equipment. (See figure 1)

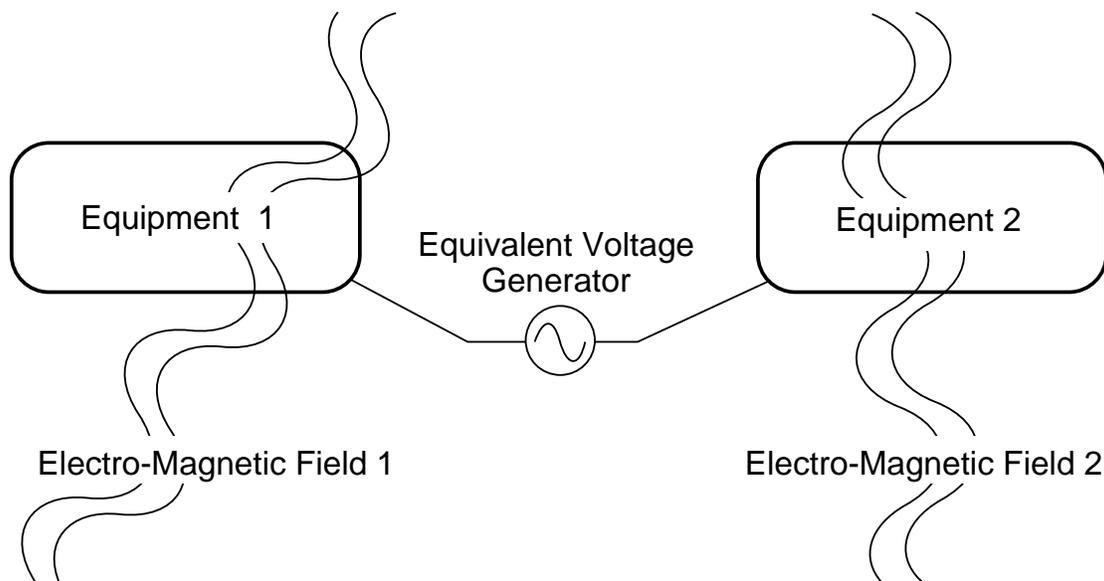


FIGURE 1 - TYPICAL VOLTAGE DIFFERENCE SITUATION BETWEEN CHASSIS

These differences are a result of the different electro-magnetic environments experienced by each piece of equipment. They come from nearby AC power lines and transformers, lighting equipment, etc., and from internal power transformer magnetic leakage. Additionally, higher frequency audio trash may be conducted into the equipment via the power line. Our objective, of course, is to reduce these potential interference sources to, or near to, zero volts.

Therefore, the following generally accepted guidelines should be observed in relation to power system installation:

2.1 Separate Feeders

The most ideal power installation uses a separate dedicated feeder for the audio equipment, where a separate transformer (if required) and panel are installed as close as possible to the equipment in use. This separate feeder should originate, if possible, at the point of power entry into the building. This will provide the lowest possible source impedance to the equipment. Likewise, the equipment grounding conductor should also originate from the “cleanest possible source.” (see details under 5.0 INTERCONNECT below)

2.2 Noise Generators

Keep all high power switching equipment off of the separate feeder (or phase) used for your quality audio. SCR lighting controls, motor switching, high power vapor lamps, etc., even florescent lights, generate a lot of ElectroMagnetic Interference. (EMI)

2.3 Quality Transformers

Use equipment that has input line filters, if possible, and quality power transformers to minimize conducted and leakage coupling. Occasionally a high isolation transformer, such as those made by Topaz and others, with ultra low coupling capacitance from winding to winding, may be necessary to eliminate high frequency conducted interference from the power line. The installation of high isolation transformers requires some careful thought to achieve the degree of isolation to which they are capable.

2.4 Separate Conduit

Provide individual, separate, steel (rigid, IMC or EMT) conduit runs for power and audio lines, separated physically as much as possible to avoid induced power related signals. Be sure to use ferrous metals for magnetic shielding effectiveness – that is, preferably not aluminum. If you find yourself in a situation where it is impossible to use conduit and virtually impossible to have a separate space for audio and power lines, such as on stage, where audio and power lines must cross, do so at 90 degree angles rather than letting them cross obliquely.

2.5 Safety or System Ground and Technical Ground

Tie all equipment together in a star configuration with insulated ground wires (to prevent random connections to other grounded objects) to reduce any potential differences between pieces of equipment to a minimum. We will discuss this in more detail under the subject of INTERCONNECT. (Section 5.0)

3.0 INTERCONNECTION SYSTEMS

While not yet universally adopted, the voltage sourced balanced interconnect system is really the only viable alternative for the audio professional. The single ended IHF interconnect system found in consumer and “semi-pro” equipment is not usable because of its inability to reject normal power line related voltage differences. Although we will do everything in our power to minimize these voltages by carefully tying the various pieces of equipment together, voltages will still exist to some degree at various frequencies.

The 600 Ω power matched system developed by the Bell System for the telephone industry is falling from favor, and should have long ago. With modern amplifier technology it is no longer necessary, nor desirable, to terminate audio lines with a “matched” low impedance, except with very long interconnects ($1/10\lambda$ or greater). The foil-shielded audio cable that we use today does not have a characteristic line impedance of 600 Ω but rather near 100 Ω , or below. The true audio transmission line is rare and typically will only exist with the telephone company, very large networks, stadiums, and various military installations. It is possible to see the effects of unterminated cable at one tenth of a wavelength, at the highest frequency of interest (20 kHz) - i.e. $\approx 3250'$ if the velocity-of-propagation constant is 0.65. In practice, with runs of 2000' or more, it might be best approached from a power matched transmission line perspective. But to do so, the correct impedance of the cable should be determined (usually between 50-90 Ω), and a proper interface provided.

3.1 Interconnect Bandwidth

The voltage sourced interconnect system with low, 50-60 Ω , source impedance and relatively high (100 k Ω or higher) input impedances, is becoming the accepted practice. An output impedance of 50 to 60 Ω has been found to be the optimum drive Z for today’s foil-shielded audio cables by ABC⁽¹⁾ in New York, and by Deane Jensen.⁽²⁾ The advantages are threefold:

- 1) Less power drawn at low frequencies from the source equipment (normally no 600 Ω loads), and therefore, less heat generated;
- 2) Lower (14 dB lower) noise pickup by interconnect lines because of the lower equivalent line impedance; and most of all,
- 3) Five to ten times the cable length that may be driven for a predetermined small signal high frequency interconnect cut-off.

To understand this last statement, one needs to see that the output impedance of a piece of audio equipment combines with the capacitance of the interconnect cable to form an R-C low pass filter.

For example, assume a circuit with 1000 feet of shielded pair (oft-times found with TV remotes) whose capacitance is 32 pF per ft between conductors, a drive impedance of 600 Ω and a bridging (100 k Ω) input. The high frequency small signal cutoff of the interconnect is 8.34 kHz! Not exactly HI-FI. (See figure 2)

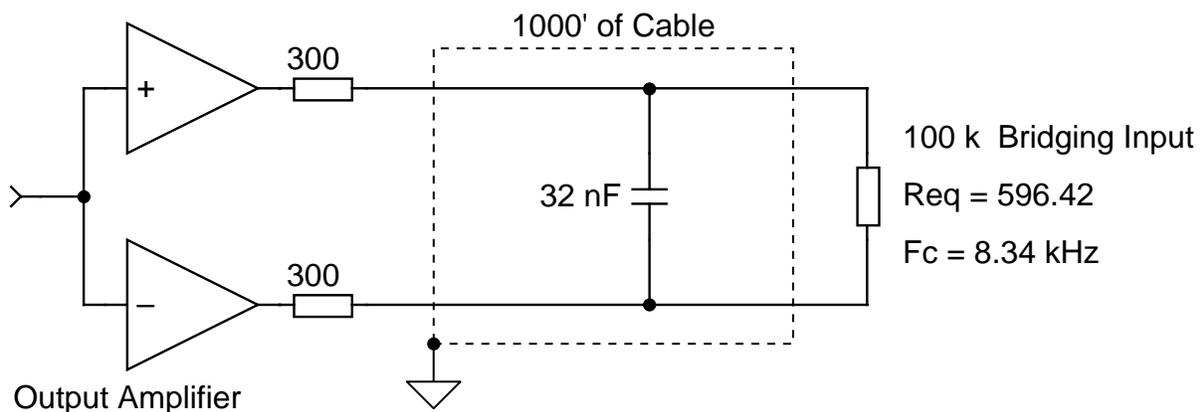


FIGURE 2 - INTERCONNECT LOW PASS FILTER

With the line terminated in 600 Ω , the situation improves to 16.68 kHz but is still nowhere near what we are looking for in the demanding world of high quality audio. By dropping the source impedance to 60 Ω , the small signal bandwidth of our interconnect moves out to 83.4 kHz, an even better situation. If we use 1000 feet of Mogami 2944A cable at 6 pF per ft then our small signal bandwidth will move further out to 442 kHz. This is a greatly improved condition, but it is still not the total story, as we shall see in a minute.

The above calculation for F_c is found by;

$$F_c = \frac{1}{2 \pi RC} \quad [1.0]$$

Where:

R is the output impedance (60 Ω),

C is the cable capacitance, and

F_c is the small signal, high frequency cutoff.

Our tests indicate that an 80 to 90 Ω source yields a maximally flat response. Lower source impedances will result in high frequency peaking because of series inductance in the cable, not accounted for in our simplified equivalent circuit. However, the best compromise between bandwidth and high end response, even though some peaking takes place at the very highest frequencies (above 100 kHz), is 60 Ω .

3.2 Interconnect Slew Rate

In addition to the small signal bandwidth limitation set by the interconnect low pass filter, an additional problem is created by the cable's capacitance. If we are to drive our cable to any appreciable voltage swing at high frequencies, we will need current from our amplifiers to feed the cable capacitance. The amount of current that is needed is in direct proportion to the total cable capacitance and output slew rate at the highest frequency to actually be sent over the cable. Slew rate is the rate of voltage change per unit time. We are, in our usage of the term, referring to the "linear" slew rate of an amplifier - that is, the amplifier is not allowed to go into saturation. Linear slew rate is defined as:

$$SR = 2 \pi f V_p \quad [2.0]$$

Where:

SR = Slew Rate = $\frac{dv}{dt}$, a change in voltage with respect to time,

f = the minimum upper frequency for full voltage response, and

V_p = the peak output voltage from the amplifier.

The amount of current that is required to feed the cable, or in fact any capacitance, is given by;

$$I = C \frac{dv}{dt} \quad [2.1]$$

3.2.1 System References

At this point, we need to digress a minute to define our system reference nomenclature. In this paper we will be using the term dBu. 0 dBu is defined as 0.7746 volts. This is the same voltage that would be found on an audio transmission line operating at 0 dBm on a properly sourced and

terminated “600 Ω” line. The use of this voltage as a reference is very desirable because of the vast accumulated experience in 600 Ω systems and because of the readily accepted volume indicator commonly called a “VU meter”. The VU meter is of course a *voltage* measuring device, where 0.7746 volts will give a “0” indication when fed with a steady tone.

While the dBu may not be a broadly recognized standard, it is an official standard among the Nordic countries in Europe. It is found in Nordic N-10 standard. Also, its common usage, particularly in Europe, causes us to accept it as the most logical way to define the voltage reference that relates to the power matched system. Occasionally, in some of our older documentation, the term dBv will be seen. This has the same meaning as dBu. Other authors will use dB/.7 or dB/0.775 to indicate the same voltage reference. One Japanese manufacturer uses the term dBs. And now back to our cable problem.

3.3 A Cable Problem

If we have 1000 feet of cable as above and we expect to be able to pass 30 kHz at the maximum output amplitude of our distribution amplifier (we will use the maximum output for most equipment of recent design of +26 dBu min., and of +30 dBu max.), what is the required drive current from our DA that will enable the above criteria to be met? Remember, 0 dBu is a *voltage* reference of 0.7746 volts.

$$\text{dBu} = 20 \log \frac{V}{0.7746} \quad [3.0a]$$

Therefore:

$$V = \left(10^{\frac{\text{dBu}}{20}} \right) \times 0.7746 \quad [3.0b]$$

Solution 1

From the above, +26 dBu is equivalent to 15.46 volts RMS output. The peak output voltage is 1.414 x the RMS value, or 21.85 volts peak. Therefore:

$$\text{SR} = 2 \times 3.14 \times 30,000 \text{ Hz} \times 21.85 \text{ V}$$

$$\text{SR} = 4.119 \times 10^6 \text{ volts per sec}$$

(or 4.119 volts per micro-second)

Solution 2

+30 dBu out = 24.50 volts RMS, peak output of 34.69 volts. Therefore:

$$\text{SR} = 6.529 \times 10^6 \text{ volts per sec.}$$

$$\text{SR} = 6.529 \text{ V}/\mu \text{ sec}$$

Now that we know the slew rate needed to meet our desired conditions, let’s calculate the current necessary to feed the line. We recall from our earlier work that 1000 feet of cable has a capacitance of 32 nanofarads (32 pF per foot), and,

$$I = C \frac{dv}{dt} \quad [3.1]$$

Therefore:

@ +26 dBu out,

$$I = 32 \times 10^{-9} \times 4.119 \times 10^6$$

$$I = 131.8 \text{ milliamps}$$

@ +30 dBu out,

$$I = 32 \times 10^{-9} \times 6.429 \times 10^6$$

$$I = 208.9 \text{ mA}$$

Multiple cables fed from a DA, of course, multiply the total current drain from a unit.

Let's suppose that we wish to drive our line with SMPTE time code and that a 100 kHz full voltage output is deemed necessary to preserve waveform integrity.

$$SR = 2 \times 3.14 \times 100,000 \text{ Hz} \times 21.85 \text{ volts}$$

$$SR = 13.73 \times 10^6 \text{ volts per sec}$$

@ + 30 dBu

$$SR = 21.76 \times 10^6 \text{ volts per sec}$$

In this case our line length must be limited to considerably less than 1000 feet. If we want a full output to 100 kHz, then we really need a small signal interconnect bandwidth of at least 300 kHz. This limits the total cable capacitance to 8.84×10^{-9} Farads. At 32 pF per foot, the maximum cable length we can use is 276 feet. The current required is:

@+26 dBu out,

$$I = 8.84 \times 10^{-9} \text{ Farads} \cdot 13.73 \times 10^6 \text{ volts per sec}$$

$$I = 121.4 \text{ mA}$$

@ = 30 dBu out,

$$I = 192.4 \text{ mA}$$

If longer cable lengths are required, low capacitance cable is the only alternative. Mogami 2944A shielded pair at 6 pF/ft allows almost 1500 feet of cable to be used and still meet the above criteria.

3.4 A Second Cable Problem

If a piece of equipment uses an NE5532 as an output amplifier, what is the maximum cable length that can be driven to the full output amplitude capability of the equipment at 30 kHz?

Solution

From the manufacturer's data sheet we find that the output current limit of the NE5532 is ≈ 40 mA. Let's assume the use of ± 15 volt power supplies and, therefore, the maximum output amplitude capability of the device is $\approx +26$ dBu.

We already found that +26 dBu at 30 kHz represents a slew rate of 4.110×10^6 volts per sec; rewriting an earlier equation:

$$C (\text{max.}) = I \frac{dt}{dv} \quad [4.0]$$

Therefore:

$$C = 40 \times 10^{-3} \text{ mA} \cdot \frac{1}{4.119 \times 10^6} \text{ volts per sec}$$

$$C = 9.711 \times 10^{-9} \text{ Farads}$$

$$C = 9.711 \text{ nF (max.)}$$

At 32 pF per foot we find a maximum of 303.5 feet of cable is permissible. With Mogami 2944A @ 6 pF per foot we find 1619 feet of cable may be used. If the equipment is capable of + 30 dBu out, then only 6.127 nF, or 191.5 feet of cable or 1021 feet of the Mogami 2944A, may be used.

3.5 System Frequency Response

All of the above assumes a 30 kHz interconnect full voltage capability. That is our recommendation for minimum system performance. It is also our firm recommendation that a 200 kHz small signal interconnect bandwidth capability be the design goal to achieve the flattest response and minimum phase shift at high frequencies for good overall system performance.

It is important to recognize the difference between the small signal bandwidth of an interconnect and its slew rate limitations.

Small signal bandwidth sets the 3 dB cutoff of the interconnect filter, which in turn describes the flatness and phase response back at 20 kHz. It must be remembered that every element of an audio system will contribute its 3 dB cutoff and associated phase shift to the overall performance of the system. Every element must be viewed as one section of a large multi-pole low pass filter; and while one element may have adequate response to 30 or 40 kHz, it is the *cumulative* effect of these filter sections that is of major concern. At first glance, the proclaimed need for a wide bandwidth of 200 kHz in both the equipment and the interconnect may seem outlandish. However, when you realize that television stations and networks, for instance, may have from ten to twenty or even more pieces of equipment in the audio chain, each contributing its cutoff characteristics, we begin to realize the magnitude of the problem in achieving adequate high frequency performance through the *system*.

If a piece of equipment or an interconnect has a 100 kHz bandwidth and has a simple single-pole, 6 dB per octave roll off (often only true for interconnects), the following chart can be used to estimate the system roll off.

Single pole response @ -6 dB/octave	Number of Filter Sections in Chain & Resultant Response				Measured Frequency
	2	4	8	16	
3 dB @ 100 kHz	-6 dB	-12 dB	-24 dB	-48 dB	@ 100 kHz
1 dB @ 50 kHz	-2 dB	-4 dB	-8 dB	-16 dB	@ 50 kHz
0.2 dB @ 25 kHz	-0.4 dB	-0.8 dB	-1.6 dB	-3.2 dB	@ 25 kHz
0.05 dB @ 12.5 kHz	-0.1 dB	-0.2 dB	-0.4 dB	-0.8 dB	@ 12.5 kHz

TABLE 3.1 AUDIO CHAIN FILTER RESPONSE

Manufacturers of audio equipment have for years mistakenly considered an upper bandwidth of 20 to 30 kHz to be totally adequate for their equipment. This narrow viewpoint, of course, fails to see their equipment as an element in a long chain, and potentially the limiting element. While we will actually never “use” - that is, put a signal into that upper portion of the 200 kHz bandwidth - it must exist to achieve the necessary 30 to 40 kHz *system* bandwidth.

We have often been told that “an audio chain is only as strong as its weakest link” – would that it *were* as strong as its weakest link.

Slew rate limitations, on the other hand, are large voltage swing limitations and again, have to do with the actual current output required of a stage that is driving a capacitance. It is important that no amplifier be allowed to slew limit. To do so produces high frequency intermodulation distortion. If an amplifier can provide adequate current to a cable to allow full output swing to 30 kHz at low THD, the chances are practically nil that it will ever slew limit with normal audio.

3.6 Interconnect Rules

So then, to maximize the interconnect performance in a system:

- 1) Move to a low output impedance of 60 Ω (balanced) on all equipment. Many pieces can be modified by changing their build-out resistors, however, current demands must be considered.
- 2) Keep the cable runs as short as possible.
- 3) Where long lines are unavoidable, use low capacitance cable (Mogami 2944A is normally in stock at Benchmark Media Systems, Inc.); also, Belden Data Cable at 12 pF per foot is a good alternative to the Mogami 2944A.
- 4) Make sure the device that is driving a long line can supply the necessary current at high frequencies.

One word of caution concerning distribution amplifiers is in order. A number of manufacturers of DAs have been changing over to a 60 Ω output impedance. However, in many cases the conversion consists of simply removing 300 Ω build out resistors and replacing them with 30 Ω resistors without regard to the possible output current demands that can occur.

If we establish as a criteria the need to be able to support a third of a DA’s outputs in a short circuit condition, remembering that a DA is an *insurance* policy, then, for a DA that uses resistive splits and 10 outputs, 3 shorted outputs represents a 10 Ω load on each of the two output legs. As a result, upwards of 3 amps peak output current capability is needed from the DAs. This drive capability is in addition to that required to feed cable capacitance. It is a rare DA that can handle this type of current demand. One shorted 60 Ω output should be reflected as only a 0.02 dB drop in amplitude at the other outputs operating at maximum output amplitude. If a distribution amplifier will not provide this degree of isolation, the question must be asked, “Is this truly a distribution amplifier?”, for is not the whole purpose of a DA to provide isolation?

3.7 Amplitude Advantages

Another benefit of the 60 Ω output impedance is the 0.8 dB amplitude difference between a bridging input and residual 600 Ω inputs. The advantage of this may not be immediately obvious, but let’s consider a not unusual TV situation where the nominal system reference (O VU) is +8 dBu. Next, we must realize that the normal peak to average ratio (crest factor) of most audio is 8 to 10 dB, but, in fact, may be as high as 16 dB, or even higher on very percussive

material. Now, let us assume that we are feeding a bridging input from a 600 Ω system (the output *voltage* will be 6 dB higher without a 600 Ω “termination”). And lastly let’s add up our amplitudes:

$$+8 \text{ dBu (system avg.)} + 16 \text{ dB (peak to avg.)} + 6 \text{ dB (for no termination)} = +30 \text{ dBu peak out}$$

When you understand that a unity gain differential (op-amp) input stage running from ± 15 volt supplies (typical for many pieces of equipment) clips at about +21 dBu, and the input clip of a properly designed input stage operating from the same ± 15 volt supply rails is +26 to +27 dBu, you begin to see the need for reducing that last 6 dB term.

4.0 COMMON MODE REJECTION (CMR)

The desirable feature of a balanced or differential input is its ability to reject a signal which, referenced to “ground,” has the same amplitude and phase on both inputs. This feature is known as common mode rejection (CMR). Whether the input stage is an active or transformer type, the amount of CMR that is needed is directly dependent on the amplitude of the common mode signal. If, for example, the power line related voltage difference between two chassis was measured, and found to be 200 mV (a long run in a network facility, perhaps), this is -11.76 dBu; and if you wanted the no signal output of your input stage to be pure noise of, say, -90 dBu, then the amount of CMR that you need is the difference between these two amplitudes - that is, 78.24 dB at the frequencies of interest.

You need to know, however, that not all differential input stages are created equal! The amount of CMR that is built into a circuit is a *large* variable. To achieve the degree of common mode rejection to which most op-amps are capable requires some very careful design, P.C. layout, and adjustment. Low frequency CMR is achieved by making the resistive portion of the diff-amp gain determining network a *balanced* bridge. Some manufacturers use 5% resistors to form the differential input stage; more use 1% metal film resistors but do not use trimmers to precisely balance the bridge. The better manufacturers use both 1% resistors and a stable trim resistor to achieve high degrees of CMR. (See figure 3)

The average CMR that can be expected with 5% resistors is ≈ 26 dB and with 1% resistors is ≈ 40 dB, whereas a carefully trimmed input stage is capable of better than 100 dB of CMR at low frequencies.

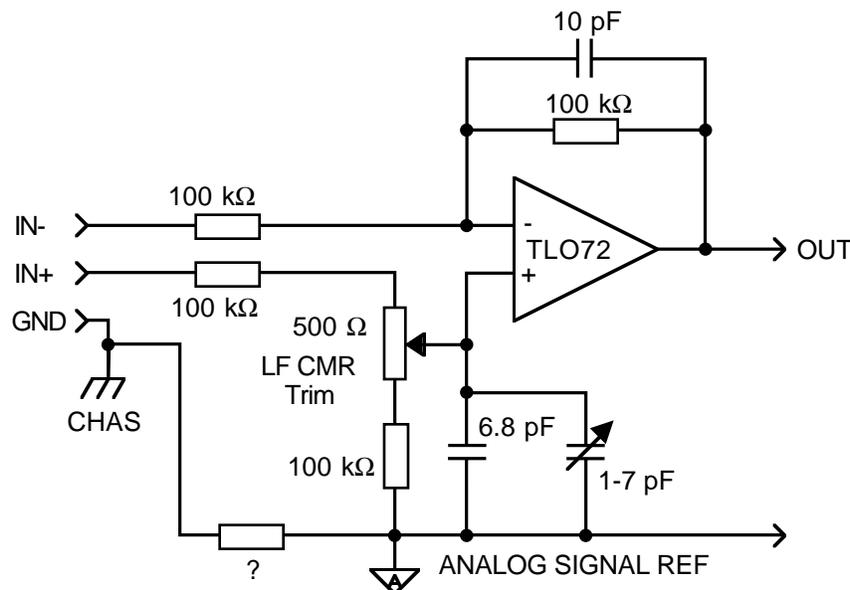


FIGURE 3 - DIFFERENTIAL INPUT AMPLIFIER

To achieve any kind of high frequency common mode rejection requires stray capacitances be trimmed out as well. Practically speaking, 75 dB of CMR is achievable out to 20 kHz by adding a capacitive trim, and even greater CMR is achievable (provided the op-amp is capable) when very careful RF style P.C. layout is practiced. The precision balance of the resistor capacitor bridge around the op-amp makes the common mode signal exactly equal at the *amplifier's* inputs, and that common mode signal is then rejected by the well balanced differential input stage within the op-amp. Unity gain buffers (not shown) in front of the differential amplifier provide the significant added benefit of equal input impedances.

Likewise, with transformer inputs, a high degree of CMR (sometimes called “longitudinal balance” after the days when the input stage transformer center taps were grounded) is necessary to reach the level of performance that we desire. While it is possible to achieve 90 dB of CMR in a transformer, most transformers on the market won't begin to come close to that figure. The input transformers made by Jensen Transformers and Reichenbach Engineering are notable exceptions, with typical 100 Hz CMR of 105 dB, 1 kHz CMR of 85 dB, and 10 kHz CMR of 65 dB. This kind of CMR performance requires extremely careful manufacturing procedures.

In terms of input types, it is generally preferable to use an active input stage (provided it has been correctly designed) rather than a transformer, because of its lower cost, generally wider bandwidth, freedom from low frequency distortion, hum susceptibility, and sometimes frequency response errors and ringing. If, however, the common mode signal cannot be reduced below two or three volts, as is often the case with telco feeds, then a transformer input should be used because of its almost unlimited (1600 volt) common mode input voltage range capability. 2 to 3 volts is the practical common mode limit for active inputs. However, with some fancy amplifier tricks, an active input stage can handle up to 200 or 300 volts common mode input.

5.0 INTERCONNECT

And now for the nitty gritty. If you adopt the following rules for the interconnection of your equipment, hum free audio can usually be achieved the first time around.

5.1 Balanced Inputs

All equipment *must* have well-trimmed balanced inputs, preferably high impedance. This is absolutely necessary, as we have just seen, to reject the residual power line related voltage differences between pieces of equipment. For equipment that does not have balanced inputs from the factory, use an interface box such as the Benchmark IFA-1. Alternatively, install internal balanced inputs such as the Benchmark DIA-1 or DIA-2. This is sometimes preferable because of the control that you have over the interface grounds.

5.2 Balanced Outputs

All equipment *should* have balanced low impedance outputs. However, satisfactory performance from unbalanced outputs can sometimes be obtained using these rules of interconnection, provided the existing output is low impedance (50 Ω or less) and the line is *not* a long line in a high RF field.

!!!CAUTION!!!

<p>The output impedance of most consumer and semi-pro equipment may range from 1 kΩ to 10 kΩ. This is purposefully done to protect you, dear reader, and the equipment, against the use of Y cables used to sum output signals to mono.</p>

Without the large build out resistor (the actual output stage is usually an op-amp with a 0 Ω output impedance) the output amplifiers would be fighting each other and oscillation is the result. This is the *only* advantage for increasing the output impedance. As we have seen earlier, there are very significant disadvantages to a high output impedance. In addition to the havoc that is wreaked on frequency response by the high output impedance and cable capacitance, another major problem will occur. If a differential input is used as the line receiver from such an output, the high output impedance will severely unbalance the diff-amp network and ruin the wonderful CMR that we paid dearly to achieve. The CMRR of a balanced receiver connected to an unbalanced source is determined by the ratio of the input impedance of the receiver to the output impedance of the source. It is better to play it safe, if at all possible, and add balanced, low impedance outputs where they do not exist. At a bare minimum, reduce the source impedance of the unbalanced output to $\approx 30 \Omega$. Then add the same resistance to the output ground side. These two resistors should be matched as closely as possible, i.e. $\approx 0.01\%$. This will provide a balanced output impedance to the differential input, even though the signal itself is not balanced. In order to really maintain the CMRR of the input diff-amp, an instrumentation type input stage is needed with identical input impedance on each leg of the input stage. (See figure 4)

5.3 Star Ground

Interconnect the analog reference from all pieces of equipment with a separate insulated ground wire (#14 or larger is preferable) to form a “star” or single point ground system. The hub of this star could be a 1/4" thick copper plate perhaps 4 by 6 inches in size and located physically at the center of the audio signal hub, i.e. an audio console, distribution amplifier or routing switcher system. The copper plate needs to be insulated from any other ground. An additional wire must be added to this star to connect the copper plate to the system’s power ground located within the power box, or, better yet, go directly to the power company’s local ground point, typically a ground stake. Be sure to use your own clamp.

At times, this “star” configuration must be created using the third wire (safety ground) of the power cord. This is, most assuredly, *inferior* to a separate conductor from the power system star ground, the reason being that the “ground reference” third wire may have a power related signal induced into it by the parallel runs of current carrying mains. Not only that, but ferrous conduit, while ideal from a shielding standpoint, will raise the impedance of the ground wire, and render it ineffective at all but power line frequencies. However, if the equipment does not have an analog signal “ground” reference within the chassis that can be made to float from the chassis, as is often the case, then this may be the only recourse. Do not be tempted to eliminate the third wire safety ground. It has been included to protect your life in the event of a power line component failure! Third wire safety grounds should have a separate wire that returns to the power box from every isolated type outlet (Hubbel #IG5362, or General Electric #GE8300-IG). Do not use the metal conduit as a substitute for the separate third wire. This separate third wire arrangement forces a star configuration. These third wire grounds should tie together in the power box via a copper buss bar that is floating from the box, which in turn is tied to the power system ground point via a 4-0 cable. (See Ref. 1 for more information)

In any case, all signal references must be tied together with a separate wire other than the drain of the shielded pair used to interconnect the signals, and, hopefully, these references will be tied together with only *one* external wire.

Ideally, all *chassis* will be tied via a brute force system by the equipment racks in which they are installed, and the *signal references* will be tied together via a star wire system. Better equipment manufacturers provide a terminal strip with the signal reference on one terminal and the chassis ground in the adjacent terminal in their equipment to allow for proper grounding practice.

With large systems, such as in network facilities where there are many systems that must be interconnected, a star of stars is the correct approach to take. The hub of the stars is again the central point of the signal distribution, usually the distribution amplifier or routing switcher system. The time taken in carefully thinking through this signal reference system will pay large dividends.

5.4 Shield Wires

Connect all inputs and outputs together without connecting the drain (or shield, if you prefer) at one end. The reason for not tying both ends to the shield is that we do not want any power line related currents to flow between equipment through this wire and risk inducing a signal into our audio inputs. Edgar Lee Howard⁽¹¹⁾ has shown that a technical advantage exists for tying the shield of an interconnect cable at the send end rather than at the receive end. The potential advantage is a reduction in coupling of high frequency noise to the audio signal due to “line to shield” capacitive unbalances.

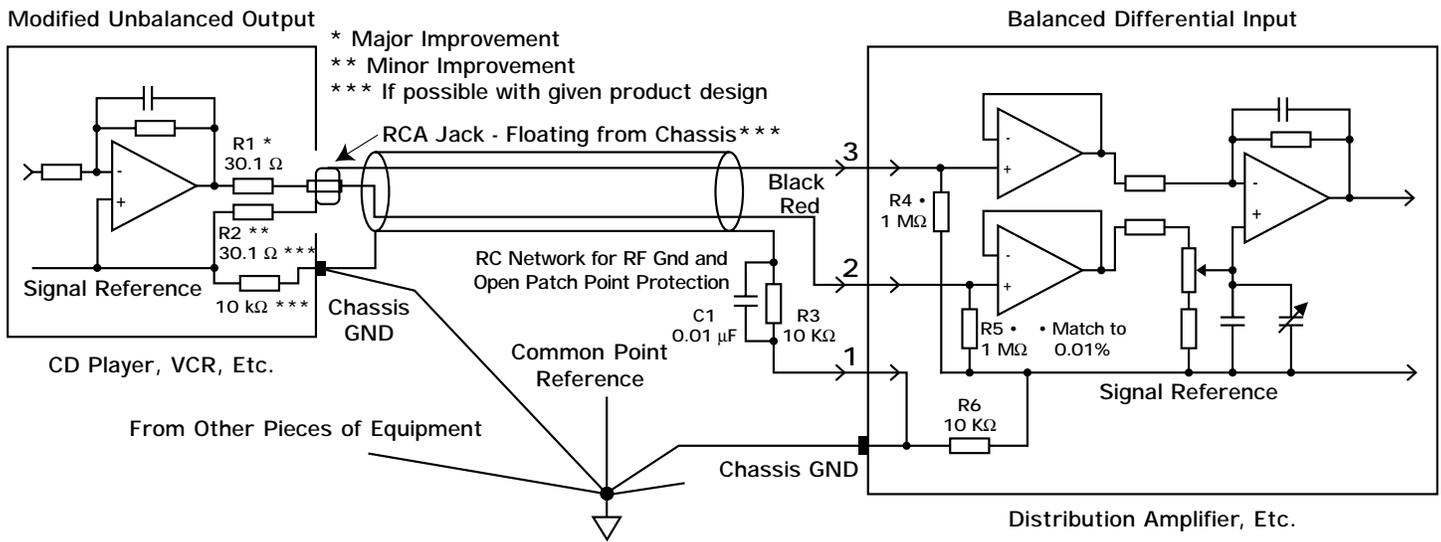


FIGURE 4 - FORWARD REFERENCING

5.4.1 Forward Referencing

Balanced outputs, of course, are connected in the normal one-to-one manner. However, unbalanced outputs must be connected so that the signal is forward referenced. (See Figure 4) This is accomplished by connecting the non-inverting (+) input to the output terminal and the inverting (-) input to the “ground” (signal reference) terminal of the unbalanced output. Now, any difference of potential that still remains between the two pieces of equipment will be ignored by the input as a common mode signal.

5.5 Jackfields

When using jackfields, be sure to use a fully balanced jackfield even when connecting to unbalanced outputs so that forward referencing can be maintained.

The jackfield presents an interesting problem in light of the above shield connection recommendations because the shield lines of the jacks are not usually switched. If you wire the jackfield so that the drain wires tie across from output jack to input jack as a part of the “normaled” wiring and do not buss the grounds together, everything is “according to Hoyle” ...until you patch. Then, all at once, you have grounds tied together that you did not want.

Perhaps the best way to approach the wiring of a non ground-switching jackfield is to tie all drain wires at the equipment ends, both input and output, but not at the jackfield. Then, at the

jackfield buss all grounds together and tie them to the system ground as though it were a separate piece of equipment, using a separate insulated ground wire.

However, the best approach to the problems of ground loops with patch bays is to use ground switching jacks. The following devices will switch the ground when a patch cord is inserted: ADC JJ-075; Switchcraft MT-336A; Audio Accessories 322A - Mil JJ-075. While it is not common to run microphone lines through a patch bay, occasionally it is necessary. In such a case, it is our firm recommendation that ground switching jacks be used.

6.0 RF INTERFACE

Often, RF related problems will remain after the above procedure has been followed. The presence of RF related problems may often be heard as an increase in high frequency noise, a “gurgling” sound, or as outright detection of the RF signal by a PN junction in an active device or by a poor solder joint. RF field intensities at a transmitter site may easily range from 1 to 100 V/Meter; as a result, the interconnect scheme described above may need modification to accommodate RF problems. With the advent of the wireless society we are experiencing, most cities now have significant RF field intensities - virtually everywhere.

6.1 Open Shield Ends

The open ended (one end) shield recommended above can cause the interconnect cable to act as an antenna. One possible cure for this is to tie the open end of the shield to its respective chassis ground through a 0.01 μ F high quality ceramic capacitor (such as Erie Red Cap or other monolithic type). Note the C1/R3 network in Figure 4. This, in essence, provides a ground at RF frequencies while leaving the shield open at audio and power frequencies. However, all capacitors have their own self-resonant frequency, and you may need to parallel two or three capacitors of different values in order to have an overlapping of curves and full effectiveness at high frequencies. The 10 k Ω parallel resistance assures a ground tie for the shield if this line should pass through a patch bay with a switched ground circuit. This is a place where it is possible for the line to have an ungrounded shield under patch conditions.

6.2 Op-Amp Configurations

The three basic op-amp gain configurations - non-inverting, inverting, and differential - each have different RF sensitivities. The non-inverting unity gain buffer amplifier is the most sensitive, and the well-balanced differential input stage is the least sensitive. Manufacturers sometimes use an instrumentation amplifier input stage which often consists of two unity gain non-inverting input buffer amplifiers followed by a well-trimmed differential input amplifier. This amplifier type will generally work well in all but the highest RF environments. In extremely high RF environments it may be necessary to replace the input buffers with jumpers and have a lower (10 k Ω or thereabouts) input impedance for the sake of RF stability.

6.3 LC Filters

At both AM and FM frequencies, series resonant L-C networks can be placed from each line of a balanced pair to chassis ground to bypass incoming RF. These devices should be placed just inside the chassis and directly at connector terminals, observing good RF wiring practice. At FM frequencies self-resonant (parallel) chokes may be placed in series with the incoming lines, and then the lines should be shunted with capacitors (that yield a maximum of 2 Ω reactance to ground at the RF frequency of interest) after the choke. A typical self-resonant choke for 100 MHz would be a single layer solenoid approximately 1/8" in diameter and 1" long. It is tuned by varying the spacing between windings and thus changing the distributed capacitance. One can expect approximately a 45 dB reduction of RF energy from such a network.⁽⁹⁾

6.4 Common Mode Filters

A very helpful but not well known device in audio is the common mode filter. When inserted in a balanced line, this filter is effective in troublesome RF environments. A typical filter will consist of a common mode choke, which in turn consists of two highly symmetrical windings on a common toroid core, two 1000 pF capacitors, and two 10 K Ω termination resistors that tie to chassis ground. The filter shown below will have a differential bandwidth of greater than 200 kHz when driven from a low impedance source, but a common mode bandwidth of only 26 kHz. This is primarily accomplished by the common mode choke itself. When the choke sees a differential signal (equal amplitude and opposite polarity), the magnetic fields created in the core cancel and the inductor effectively disappears from the circuit. However, when the signal is of equal amplitude and of the same polarity, as is the case with the interference we wish to remove, both lines see the L-C low pass filter. As a two pole device with a cutoff rate of 12 dB per octave, the filter is down 60 dB at 1 MHz and thus prevents RF from reaching the active electronics. (See figure 5)

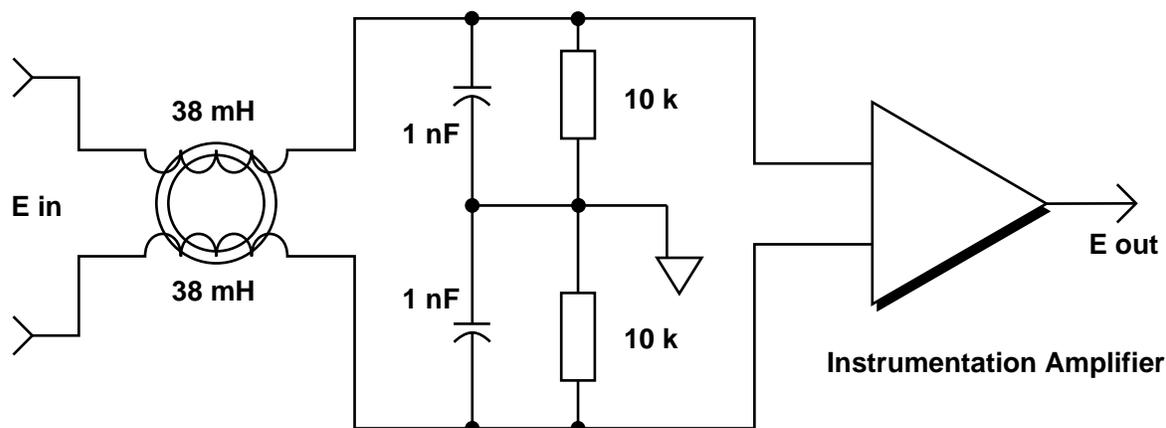


FIGURE 5 - COMMON MODE FILTER

6.5 Shielding Materials

Additionally, further shielding may be necessary. Separate ferrous conduit for the audio cabling can be very effective with RF as well as with power line radiation. Tightening up the shielding of specific audio packages may be necessary. Numerous EMI type seals and gaskets may be obtained from either:

Tecknit, Inc.
129 Dermody Street
Cranford, NJ 07016
(908) 272-5500

or

Chomerics, Inc.
77 Dragon Court
Woburn, MA 01888
(617) 935-4850

6.6 Screen Rooms

In a few extreme cases with which the author is familiar, entire studios had to be constructed inside RF tight “screen rooms”. If you need to construct a screen room for your facility, use steel screen, if you can locate it, for all interfering frequencies including the FM band and below, since it is more lossy and thus is more effective than copper, not to mention the expense. Copper is necessary for the microwave frequencies, and there exists a gray area in between. You will need to solder the seams, and when your room is finished, be sure to use power line filters physically located at the point of power entry into the room to clean up the incoming mains.

7.0 SIGNAL-TO-NOISE RATIOS

It is our recommendation that all audio systems be run at a maximum “nominal 0 VU” audio voltage amplitude of +4 dBu. As mentioned before, with ± 15 volt power supplies most op-amps will clip around +21 dBu (sine wave - RMS). This yields a headroom or overload factor of 17 dB, which we saw earlier is a bare minimum for live audio. Peak to average ratios of 16 dB have been measured on speech, and we at Benchmark have seen at least this high a ratio on percussive music. It is our design philosophy to take a 6 dB loss at the input of our equipment, thus yielding an input clip point of +26 to +27 dBu (up to +30 dBu with 20 volt supplies) and internally operating the circuits at a nominal average amplitude of -2 dBu, yielding a headroom factor of 23 dB. We then achieve the 6 dB gain makeup needed at the output stage. This allows both input and output clip points of +26 to +27 dBu and a more desirable headroom factor without much sacrifice in noise performance. The output noise floor of carefully designed equipment can reach -93 dBu or better. This yields an average signal-to-noise ratio of 97 dB and peak signal-to-noise ratio (dynamic range) of 120 dB. Many Benchmark products actually have a dynamic range greater than 131 dB. We recommend, therefore, that all of your equipment be capable of this high an input and output amplitude.

7.1 System Setup

It now becomes very important that you *carefully* set up your system. Set it so that all of the gains allow the various pieces of equipment and indeed *every stage* within the pieces of equipment to reach their clip points at the same time. This is done by taking the majority of the gain needed from the amplifier stage that has the lowest noise figure, the point where the system’s dynamic range is established, the mic preamp. As simple as this sounds, it is the key to an outstanding system, provided the installation has been performed correctly.

7.2 Noise Primer

We at Benchmark Media manufacture microphone preamplifiers with a one dB noise figure so you may enjoy the highest signal-to-noise ratio possible. This section will include the MP-4 mic preamp, the circuit used in the MPS-400/420 microphone preamplifier systems, as an example.

Noise figure is a measure of how well an amplifier amplifies the intended signal without adding noise. In the case of the MP-4 the amplifier adds only one dB of noise to that of the original signal for amplification factors greater than 40 dB. The noise figure is referenced to the Johnson noise of the resistive portion of a transducer’s source impedance.⁽¹²⁾

Johnson noise may be calculated from:

$$E_n = \sqrt{4 kTRB} \quad [5.0]$$

Where:

k = Boltzman’s Constant = 1.38×10^{-23}

T = temperature of resistance in degrees Kelvin
(room temperature $\approx 300^\circ$ K)

R = resistance = microphone source impedance

B = bandwidth = 19,980 Hz (20 Hz - 20 kHz)

From the above formula, we see that the noise of a 150 Ω resistor at room temperature is 222.9 nanovolts or -130.82 dBu, whereas a 200 Ω resistor has a noise voltage of -129.57 dBu.

7.3 Amplification of Noise

Any amplifier, while amplifying the desired signal from a microphone, will also amplify the Johnson noise from the source resistance. Therefore, the output noise of a totally noiseless amplifier operating @ 50 dB of amplification from a source resistance of 150 Ω at room temperature would be -80.82 dBu. The MP-4's performance under these conditions is \approx -80 dBu. At its minimum amplification (18 dB), the MP-4 has an output noise floor of -94 dBu. The noise increases slowly as the amplification is increased to 40 dB where the output noise is approximately -88 dBu. From this point on, the noise will increase directly with the increase in amplification. Now that we have the tools, let's see how this understanding of noise applies to two specific microphones. Knowing the source resistance, self noise, in the case of a condenser microphone with its own amplifiers, and sensitivity, you can then evaluate the performance of any microphone with the MP-4 under various sound pressure level (SPL) conditions. In the following examples, you will notice the addition and subtraction of dB quantities that are derived from SPLs and voltages. This is possible since sound pressure level is a voltage scalar quantity.

7.4 The Electro-Voice RE-20

The sensitivity of an RE-20 is 1.09 mV (-57 dBu) at 94 dB SPL input. If we use a voltage amplification of 58 dB, the output voltage of the preamp at 94 dB SPL input will be +1 dBu, and the noise of the MP-4 will be approximately -72 dBu (-130 dB EIN of the preamp +58 dB gain). If the sound pressure level is sufficient to give us an output of +4 dBu, then the average signal-to-noise ratio capability is 76 dB (+4 dBu - [-72 dBu]). The sound pressure level necessary to achieve an average output of +4 dBu is +97 dB SPL (94 dB SPL + [+4 dBu - {+1 dBu}]). The peak SPL that the system can handle is +123 dB SPL using \pm 20 volt supplies, since they allow a maximum output amplitude of +30 dBu (30 dBu - [+4 dBu] + 97 dB SPL). At this amplification and SPL, the preamp will reach its output clip point and yield a dynamic range of 102 dB (+30 dBu - [-72 dBu]).

7.5 The Sennheiser MKH-40

Let's examine the performance of the Benchmark MP-4 with the high output Sennheiser MKH-40-P48. This microphone has a very high sensitivity of 25 mV per Pascal (10 dynes per $\text{cm}^2 = 1$ Pascal = 94 dB SPL), and very low self noise (a condenser microphone with internal electronics) of 12 dBa. The self noise, therefore, is 1.99 μV "A" weighted. This translates to \approx -111.8 dBu. Using 18 dB of amplification at the MP-4, the combination output noise is approximately -93 dBu (20 kHz noise bandwidth). At the reference SPL of 94 dB the output of the microphone is 25 mV or -29.82 dBu. Add 18 dB of amplification and the output amplitude from the preamp. The output voltage is now -11.82 dBu @ 94 dB SPL. To get +30 dBu out (just below the clip point), the additional SPL needed is; the preamp clip point minus the output voltage at the reference SPL or:

$$+30 \text{ dBu} - (-11.82 \text{ dBu}) = 41.82 \text{ dB}$$

additional SPL to be added to the reference. Adding this to the reference yields:

$$+ 94 \text{ dB SPL} + 41.82 \text{ dB} = 135.82 \text{ dB SPL for peak clip.}$$

Thus, we have an average signal to noise ratio of 97 dB (+4 -[-93]). A peak signal to noise ratio, that is - dynamic range, of 123 dB (+30 -[-93]) at a peak acoustic input of 135.82 dB SPL. This is only possible using \pm 20 volt supplies with the MP-4. This input SPL is just above the 134 dB SPL/0.5% THD point of the microphone. 109.82 dB SPL average is required at this amplification for an output of +4 dBu. Now let's apply what we have just learned to a practical problem.

7.6 An SPL Problem

If a 20-bit digital A to D converter (the Benchmark AD2004) has a “dynamic range” (peak signal-to-average broadband [20 kHz brickwall] noise ratio) of 109.2 dB, an input clip point of +24 dBu into an balanced input, what is the lowest peak sound pressure level that the Sennheiser MKH-40 microphone can receive and still maintain the full dynamic range of the converter? Also, what amplification is necessary from the MPS-420 microphone preamplifier system to achieve this? The MPS-420 uses the MP-4 preamplifiers with ± 18 volt power supplies.

Solution

First, since the ± 18 volt supplies of the MPS-420 yield a balanced output clip point of greater than +28 dBu and the input clip point of converter is +24 dBu there will be no limitation in reaching Full Scale Digital code when driven from the MPS-420. Since the dynamic range of the converter is 109.2 dB, (Full scale digital to 20 kHz noise floor) then the converter’s noise floor is:

$$+24 \text{ dBu} - 109.2 \text{ dB} = -85.2 \text{ dBu}$$

If the microphone self noise is -111.82 dBu then the maximum amplification that we can use is (converter noise floor - self noise), that is:

$$-85.2 - (-111.82) = 26.62 \text{ dB}$$

Actually, if we make both the converter and the output of the MPS-420 have the same noise voltages, they will add and we will experience a 3 dB loss in dynamic range. Hence, it is well to keep the MPS-420’s output noise voltage 3 dB lower than the noise voltage of the converter (even at 3 dB lower there will be a 1.8 dB loss in range). Therefore, let’s select a maximum amplification factor of:

$$26.62 \text{ dB} - 3 \text{ dB} = 23.62 \text{ dB}$$

We can round off to $A = 24$ dB. With 24 dB of amplification, the peak input voltage that the MPS-420 can receive is, the FSD clip point of the converter - gain of the mic preamp = peak input level from mic, which is:

$$+24 \text{ dBu} - 24 \text{ dB} = +0 \text{ dBu}$$

From the microphone sensitivity figure given by the manufacturer, we find that @ 94 dB SPL into the mic, we have an output voltage of 25 mV = -29.82 dBu. Therefore, we can have a 29.82 dB higher SPL than the reference, where we reach the clip point of the converter. Converter input clip - sensitivity = additional SPL over sensitivity reference:

$$+0 - (-29.82) = 29.82 \text{ additional SPL}$$

Therefore, the peak SPL (electronics clip point) is:

$$94 \text{ SPL reference} + 29.82 \text{ additional SPL before clip} = 123.82 \text{ dB SPL}$$

This converter is professional device and normally set up to take a balanced input level of +4 dBu to give a -20 dBFS on the LED meter of the converter, and +24 dBu for Full Scale Digital code This provides 20 dB of headroom. Therefore the average SPL that we should have at the microphone is:

$$123.82 \text{ dB SPL peak clip} - 20 \text{ dB headroom} = 103.82 \text{ dB SPL average input}$$

In this example we have worked with the commonly accepted definition of dynamic range i.e. “The difference between Full Scale Digital code and the broadband noise floor of the converter.” Whereas in reality the dynamic range will extend significantly below the broadband noise floor! This is because the human hearing mechanism acts much like an FFT with averaging capability that allows us to hear below the broadband noise. We, at Benchmark, have measured and listened to tones that were at -135 dBFS, 25.8 dB below the broadband noise floor of the AD2004, (we added digital gain of 60, or more, dB after the converter). Our ability to hear below the broadband noise floor is not insignificant and we will illustrate why it is, in a moment.

Now, the above calculations *do not* take into account the ambient acoustic noise of our recording environment. They are meant to demonstrate how, with a given microphone, the preamp gain can be optimized to meet the common definition of dynamic range in the converter. However, this also clearly demonstrates the need, in this digital recording age, to use recording environments with *extremely* low ambient acoustic noise levels, if we really expect to even come close to realizing the converter’s full dynamic range.

For example, If the peak SPL shown above, +123.82, is the highest that will be expected during a recording session, and the dynamic range of the converter is 109.2 dB then the converter’s noise floor is at an equivalent ambient SPL of 14.62 dB plus 1.8 dB for the noise addition factor = 16.42 dB SPL. Now considering our ability to resolve signals that may be 26 dB below this broadband noise floor, we run out of positive sound pressure level given by our current definitions (0 dB SPL is 20 μ Pa, [micropascals] the threshold of hearing, 1 kHz - 4 kHz, in youths). And who has measured a recording venue with an ambient sound pressure level that even begins to approach 16.5 dB SPL?

If our sound source can, and normally does, produce higher peak SPLs than the 123.82 figure calculated above, we may reduce the gain of the microphone preamplifier from +24 to +18 dB (the minimum gain of the MPS-420) which will raise the SPL clip point to 129.82 dB. The average SPL will move up to 109.82. A recording venue ambient noise of +20 SPL might be able to be tolerated. Beyond a gain reduction to +18 dB, we will have to use an attenuator on the input of the preamp to get further gain reductions and this is not desirable as it will cause us to run into the self noise of the preamplifier. The next move we can take to increase the peak SPL input to the microphone is to choose the +28 dBu Full Scale Digital input setting on the AD2004. The peak clip output point of the MPS-420 operating with \pm 18 volt supplies exceeds +28 dBu. This buys us an additional 4 dB to approximately +134 dB SPL peak. To handle recording sessions with SPLs that are higher than this will require using microphones that have less sensitivity than the Sennheiser MKH-40 and recalculating the figures.

No matter how we try, however, we can’t get away from the need for *extremely* quiet recording venues. Pray tell, how will we take full advantage of the coming 24-bit converters?

7.7 The Limiting Factor

The amplification required for many microphones will typically be 40 dB or greater, and the preamplifier section will usually be the limiting factor in the output noise of a console or other electronics prior to any recording or transmission medium. Therefore, the majority of amplification needed, consistent with desired headroom, should be taken from the mic preamp, since it will have the lowest noise figure of any of the amplifying stages.

7.8 Conclusion - Setup In Practice

Proper attention to these two areas; 1) clip points made equal and; 2) maximum gain from the mic preamp; will optimize the signal-to-noise ratio of your system. Unfortunately, this author has encountered *very* few audio systems that were set up correctly. This, of course, results in a lowered signal-to-noise ratio (lots of noise), poor headroom (lots of distortion) and frustration (*lots* of it).

8.0 PROPER OPERATION

Many audio operators find themselves fighting a perplexing and confusing battle with audio amplitudes. This often is because of the inability of the VU meter to show peaks of program material, which are the source of overmodulation, tape saturation, amplifier and converter clipping. The VU meter, while being a universal measurement device in the Americas, Japan, France, and Australia,⁽¹⁰⁾ finds less application in the rest of the world because of this problem. The problems occur when adequate headroom has not been designed into the system. The late Hans Schmid of ABC in New York has shown⁽²⁾ that use of the peak program meter will give clearly superior indication of program material without lowering average amplitudes, but with a new control over excessive program peaks. In order to observe peak program amplitudes, it is desirable to convert existing VU meters to operate as a PPM as well as VU by the use of a conversion card, or add parallel PPMs along side the VU meters. As a bare minimum, at least add peak overload indicators of some form to any equipment without that capability.

9.0 TESTING

Overall noise and frequency response measurements should be taken. It is our firm conviction that if the above procedures are followed, wide band ($f_3 = 100$ kHz) audio systems (depending upon the number of pieces of equipment) right up to the band limited device, such as an A to D converter, stereo generator, or pre-emphasis network, are feasible. This allows the band limited device to set the overall system bandwidth. This, as we have seen, allows for a flatter frequency response (not counting intentional response shaping) rather than allowing an accumulation of 3 dB roll-off points to upset the high frequency response and a proof-of-performance test in broadcast systems, for instance. It allows low phase shift out to 20 kHz and good transient response, provided the various pieces of equipment have proper feedback compensation.

9.1 Square Wave Testing

We recommend that you do extensive 10 kHz square wave testing of the system by placing a function generator at the start of the audio chain and then look at every output with your scope. We recommend that all response type measurements be made at -12 dBu. This conforms to the new CCIR recommendations.⁽¹⁰⁾ For square wave testing, it ensures that no slew rate limitations will confuse the picture of system bandwidth. The -12 dBu level makes sure the amplifiers under test do not go into saturation during this test; otherwise, any accuracy in the conclusions is null and void. Check for overshoot or ringing, which would indicate under-damping of amplifier circuits, and thus potential instability and RF susceptibility. Do this test with each piece of equipment separately and then as a system. A small amount of high frequency related overshoot from an interconnect may be seen due to 60 Ω drive impedance. As discussed above, the 60 Ω figure is actually a compromise in favor of bandwidth and is slightly lower than necessary for a "maximally flat" response. There will be some HF peaking. Approximately 85 Ω would give the maximally flat response, but is restrictive of bandwidth. When you reach a stage that has significant bandwidth limitation, noted by the strong rounding of the corners of the square wave, it is time to move the generator up past this point in order to get a true indication of what is happening in the following stages. (See Ref. 7 for more information)

9.2 Distortion Testing

Once you are satisfied with the frequency response, noise floor, and transient response of your system, it is well to do some overall distortion measurements. While low frequency total harmonic distortion measurements have merit, in bandwidth limited systems, high frequency THD measurements are almost meaningless. The filters in an A to D converter or stereo generator do a nice job in removing the distortion products of anything above 10 kHz and destroy the accuracy of any measurements much above 2 kHz; hence everything may measure better

than it actually is. If slewing induced intermodulation and/or transient intermodulation distortion exist within the system, it will only occur at the higher frequencies. The best way in a bandwidth limited system to detect their presence is with the CCIF twin tone IM distortion measurement. By using 14 and 15 kHz tones mixed 1:1, a 1 kHz IM product is easily detected if SID/TIM exists in an FM broadcast system. A 19 and 20 kHz pair may be used to evaluate A to D converters or other equipment in the audio chain. We believe that every broadcast facility should perform twin tone IM distortion measurements to see the truth about their high frequency audio performance.

9.3 Useful Tools

Several tools useful in testing your system apart from your ears:

- 1) The telephone pickup coil, Radio Shack #44-533: with a pair of headphones and a microphone preamplifier, it is possible to do a thorough electro-magnetic survey of the space in which you intend to place your equipment. See the Benchmark Media Systems, "The Audio Microscope™" application note for an amplifier setup that will do the job.
- 2) A battery operated wide band AC. voltmeter, with which you can measure the voltage differences between various pieces of equipment.

The ultimate in test systems, from our viewpoint, is the new Audio Precision, *System-Two*, Audio Precision, PO. Box 2209, Beaverton, Oregon 97075, 1-800-231-7350. This device's abilities would take pages to describe. See a review of the older System One by the late Hugh Ford in "Studio Sound", June, 1986. In addition to all of its distortion measurement capability, it has the ability to do an FFT analysis and thus give amplitude measurements of all the power line harmonics that may be present in your audio. This will allow you to discern between hum that is primarily from internal power supplies (second harmonic), and that which may be induced from the power mains (fundamental and third harmonic).

10.0 CONCLUSION

If you perform these steps with care, you will have a system that is capable of outstanding performance. It should have a residual noise floor not too far from that of the equipment with the highest output noise. In a broadcast facility, the satisfying of these criteria and its proof by the above measurements will allow you to complete a more meaningful set-up of your compressor-limiter, after which additional experimentation such as L-R processing can be performed. Hopefully, at this point, the engineering department and the production people will find harmony in their Clean Audio.

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