VALLEY PEOPLE REVISED TRANSFORMERLESS MIC PRE-AMP CIRCUIT

The circuit shown has been laboratory tested, and represents the TRANS-AMP LZ mic pre-amp in its fully optimized form. The circuit is capable of direct transformerless amplification of both dynamic or phantom powered condenser microphones, whose impedances range from 50 ohms to 600 ohms. Specifications stated are for the nominally standard 150 ohm microphones used in most professional installations.

Typical Performance

Gain Range = Adjustable (R14) from 10dB to 60dB (Gain = $1 + \frac{R15 + R16}{R13 + R14} \times \frac{R20}{R17}$)

Max. input level = +12dBv with circuit values shown
Noise Figure @ 60dB gain = .5dB (EIN = -130.3dBv re .775v)
Noise Figure @ 30dB gain = 3dB (EIN = -127.8dBv re .775v)
I.M. or THD = Below .01% at any setting or level below clipping
Bandwidth (full power or small signal) = 1Hz to 150KHz at any setting
Slew rate = 13v/μsec
Common Mode Rejection Ratio = Typically trimmable to in excess of 100dB
Output capability = +22dBv into 600Ω or greater (NE 5534 output amp)
+22dBv into 2K or greater (TL 071 output amp)

Circuit Description

The mic is supplied with Phantom Powering voltage, in the industry accepted manner by R1 and R2. These resistors should be 1% or better matched, for minimum generation of non-common mode hum from the Phantom supply.
Capacitors C1 and C2 serve to isolate the elevated phantom power voltage from the TRANS-AMP LZ input. Tantalum capacitors should be used here, for extended life and minimum leakage currents. A suitable capacitor is available from Kemet (T310 series - 18μfd/50v - $1.39 in 25 qty.) and from other manufacturers. Although 18μfd may appear a low value for coupling a 150 ohm source, it should be noted that the frequency response is determined by $f = \frac{1}{2\pi XcC}$ wherein Xc equals the 100K differential input impedance of the TRANS-AMP LZ. This works out to a low end roll off at .18Hz. The capacitors will increase the noise figure of the amplifier at sub-audio frequencies - a point of essentially no concern in practice Larger value capacitors would reduce this effect, but would become prohibitive in terms of size, cost, leakage current and hum induction. Noise level measurements should be taken through a 20Hz high pass filter because of the increased sub-audio noise.

Input Protection

R4, R5, and Z1-4 form an input protection circuit to protect the TRANS-AMP LZ input stages from high voltage surges which may be experienced when connecting and disconnecting microphones from the input. These components must not be ignored when the system is Phantom Powered, or permanent damage may result to the TRANS-AMP LZ.

The circuit clamps each input leg to ±6.8v maximum potential. The circuit has no effect on signal inputs below +21dBv.
If diode clamps to the power supply rails are substituted, it should first be ascertained that these rails may absorb the potential discharge from C1-C2 without damage or voltage spikes.

De-Thump

The input stage of the TRANS-AMP LZ is dc coupled, and should be offset to zero volts differential, if rapid gain changes are to be made with R14. If not so offset, large magnitude level shifts (thumps) will be heard, especially when adjusting in the high gain spectrum. The circuit of R11 and R12 provide this offset adjustment.
De-thumping may be alternatively accomplished by inserting a large value capacitor in series with R13. This is somewhat impractical, in most applications since a low frequency roll off will be introduced at $f = \frac{1}{2\pi X_c C}$ wherein Xc equals R13 + R14 at the selected gain. At 60dB gain setting a 7000µfd capacitor would be required for 1dB roll off at 20Hz.

Gain Setting Resistors

The TRANS-AMP LZ gain is set by the ratios of R15, R16, R13, and R14, and is determined by the formula: differential gain = $1 + \frac{R15 + R16}{R13 + R14}$. The total circuit gain is equal to the TRANS-AMP LZ gain plus the output amp gain (loss). The maximum gain is set by R13, while the minimum gain is established by R14 (the gain adjust pot).
It should be carefully noted that, for thermal noise purposes, R13 and R14 are effectively in series with the microphone, and must be kept as low in resistance as possible. For instance, at the gain point where R13 + R14 equal the mic impedance of 150 ohms, the noise figure is necessarily degraded by 3dB. With the values shown, this occurs at a net circuit gain of 24dB. Fortunately, at this relatively low circuit gain, the actual output noise figure is of no consequence in achieving ultra low noise performance.
Indiscriminately raising the values of R13, 14, 15, and 16 can, however, bring the noise performance into an area of degradation.

The Gain Set Pot

A higher value potentiometer can be substituted for R14, allowing lower minimum gains (below unity) with resultant higher allowable input levels (to +21dBv). The practical limitation here, however, is a loss of adjustability at the high gain end, with commercially available potentiometers. The problem lies in the fact that, at the high gain end, the pot must have decent resolution in the 0 to 10 ohm region. Our experience has not turned up a suitable pot with such characteristics. Alternatives might be:

A. A custom tapered pot (i.e. 5K reverse compound audio taper)
B. A switchable 2 pot scheme (i.e. 1K high gain / 10K low gain)
C. A switch in series with R14 which opens completely for very high input levels (to +21dBv) and provides a net circuit loss of 5dB

Output Coupling

Since the TRANS-AMP LZ produces a differentially opposed ouput swing, its

maximum output swing is 6dB greater, for a given power supply voltage, than a single ended output amplifier. (i.e. +28dBv or 55 volts peak to peak for ±18volt supplies)
Since the output amplifier, in the circuit shown, is limited to an output swing of +22dBv, a loss amounting to 5.2dB is introduced in the output coupling circuit, in order to equalize the clipping points and maximize the signal to noise characteristics of the total circuit.
Higher output swings can be realized by substituting a higher voltage output amplifier operating on higher supply voltages (i.e. ±36 volts), and omitting the 5.2dB loss. Extreme consideration must be given to slew rates, however, in employing elevated levels, since a 26v/µsec device would be required in order to not degrade transient response.

The NE5534 Circuit

If the circuit is to drive 600ohm loads, the signetics NE5534 op-amp (about $2.00) is suggested, as it can deliver +22dBv into 600ohms, at a slew rate of 13v/µsec. (170KHz full power bandwidth)
A slight inconvenience with the 5534 is that it is not stable at gains below 3, unless over-compensated (with a resulting degradation of slew rate). In the circuit shown, the values of R17 through R22 are adjusted such that the gain of 3 stability criteria is met, while maintaining the desired -5.2dB signal transfer. Thus the circuit is stable, without over-compensation. The extremely low noise voltage of the NE 5534 allows superior noise performance, in spite of the increased operating gain.

The TL071 Circuit

Where the driven load is 2Kohms or higher, the Texas Instruments TL071 series (with counterparts from National LF series) are equally effective. (cost about $ .50)
They too exhibit 13v/µsec slew rates. Additionally, these devices are stable to unity gain. With the omission of R19, and the value change of R22 (to 10Kohms), the noise performance is essentially identical to the NE5534.
Other devices, such as 741S, LM318, etc. and discrete op-amps, may be employed with good results.

The CMRR Trims

Extreme Common Mode Rejection Ratio requires careful balancing of common mode paths. This is particularly true with the circuit configured for Phantom Power capability, as the series elements R4, R5, C1 and C2 are capable of introducing errors in the magnitude of signals reaching the TRANS-AMP LZ inputs.
It must be remembered that an error as small as .01% between signal levels on the two inputs can reduce the achievable CMRR to 80dB. Such an error can be introduced, at 60Hz, by a 3% mismatch of C1 and C2, unless compensated for. In the circuit shown, 3 CMRR trims are provided (R6, R9 and R21) which serve the following functions:

A. R21 cancels the error produced in R17 through R22
B. R9 cancels the error produced by C1 and C2
C. R6 cancels the error produced by R5 and R6, together with the small CMRR error of the TRANS-AMP LZ itself

Properly adjusted, the circuit provides CMRR's well in excess of 100dB (typically 115 to 125dB), throughout the audio range, with excellent CMRR extending into the RF region.

RF Suppression

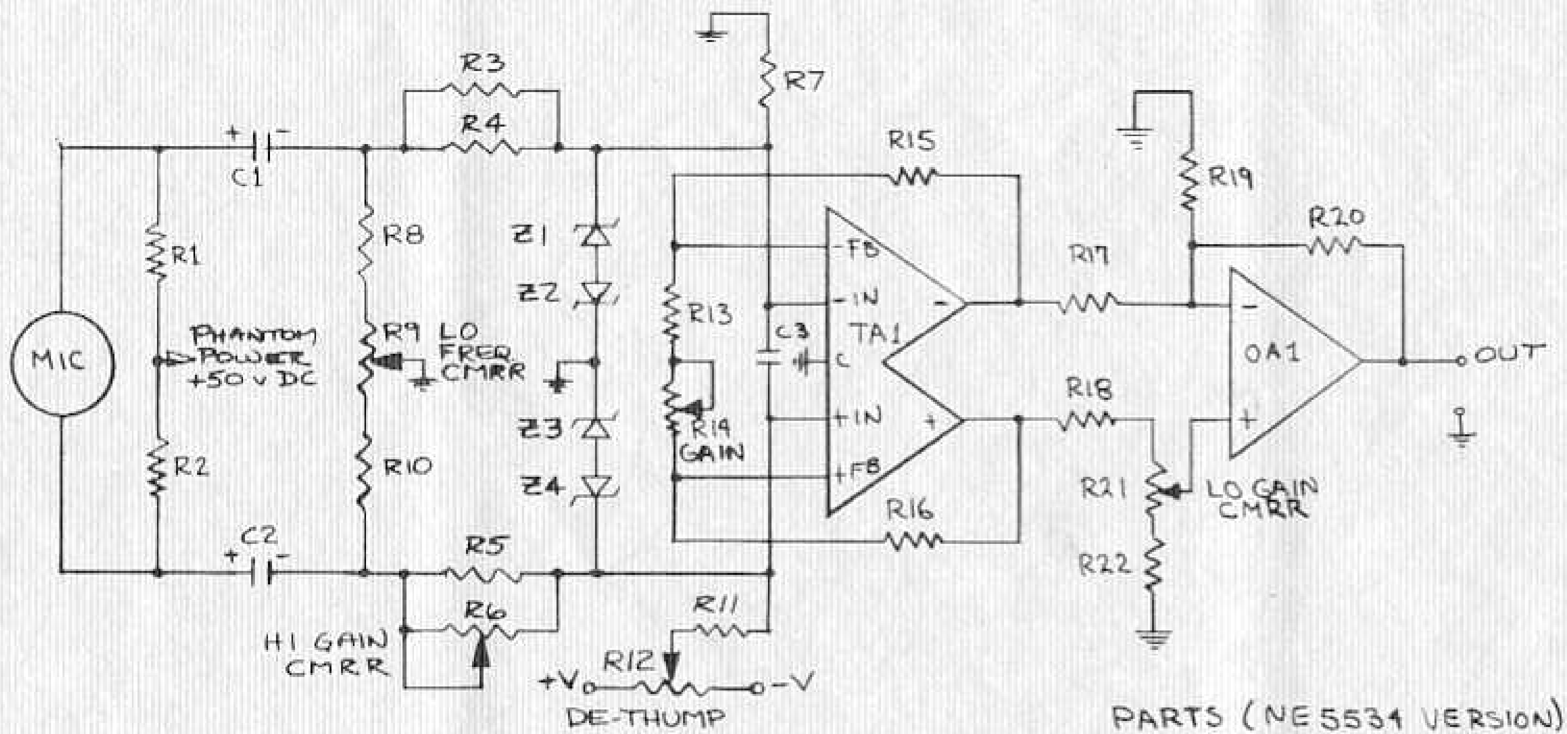
Capacitor C3, together with R4 and R5, provides additional RF Suppression, by limiting the input bandwidth to about 200KHz. C3 also serves to insure stability with capacitive/inductive input sources. The circuit is typically stable with C3 as low as 560 pfd, in which case the bandwidth is extended beyond 1MHz.

Adjustment/Set Up

Although the trims may certainly be adjusted with test equipment, a faster, and sometimes more effective method, simply involves a signal source (osc) and monitor speakers.

Procedure:

1. Set R6, R9, R11 and R21 to center range
2. With Phantom supply off, short the two mic terminals together, and apply 1KHz, at a -10dBv level, (250mv Rms) to the shorted inputs
3. Set R14 to Minimum Gain Position
4. Adjust R21 for minimum 1KHz at output of circuit
5. Rock R14 between Min. and Max. Gain, while adjusting R12 for minimum "Thump"
6. Set osc to 60Hz, adjust R9 for minimum output
7. Set osc to 1KHz, adjust R6 for minimum output
8. Repeat 3 - 7, if necessary
9. Seal trimpots with fingernail polish, for vibration resistance



IMPROVED TRANS-AMP MICROPHONE PRE-AMP
GAIN = 10dB TO 60dB
MAX INPUT = +12 dBv

PARTS (TL071 SERIES VERSION)
SAME AS NE5534, EXCEPT:
OA1 = TEXAS INSTRUMENTS TL071CP
OMIT R19
R22 = 10K 1%
MAX OUTPUT = +22dBv @ 2KΩ

PARTS (NE5534 VERSION)
OA1 = SIGNETICS NE5534
TA1 = TRANS-AMP
Z1-Z4 = ZENER 6.8v/400mW
C1,C2 = 20ufd 50V TANTALUM
R1, R2 = 6.81K 1%
R3 = 100 Ω 5%
R4, R5 = 10Ω 1%
R6 = 1K TRIM
R7, R11 = 2.2 Meg 5%
R8, R10 = 68K 5%
R9, R12 = 50K TRIM (R9: 10 TURN)
R13 = 2.2Ω 5%
R14 = 1K REVERSE AUDIO WITH LOW END RESISTANCE
C3 = .005ufd
R15, R16 = 2.21K 1%
R17, R18 = 18.2K 1%
R19 = 5.62K 1%
R20 = 10K 1%
R21 = 100Ω TRIM
R22 = 3.57K %

NOTES
1. MAX SUPPLY = ±18 VOLTS
2. DECOUPLE SUPPLIES WITH .1ufd MYLAR + 2.2ufd TANT
3. MAX OUTPUT = +22dBv @ 600Ω
4. FOR EQUAL SIZE RESISTORS USE 1/4 W 5%s AND 1/8 W 1%s.

TRANS-AMP™ LZ

APPLICATION NOTE 2 (A.N. 2)
2-78

THE TRANS-AMP LZ AS A TRUE BALANCED DIFFERENTIAL CURRENT SUMMING AMPLIFIER

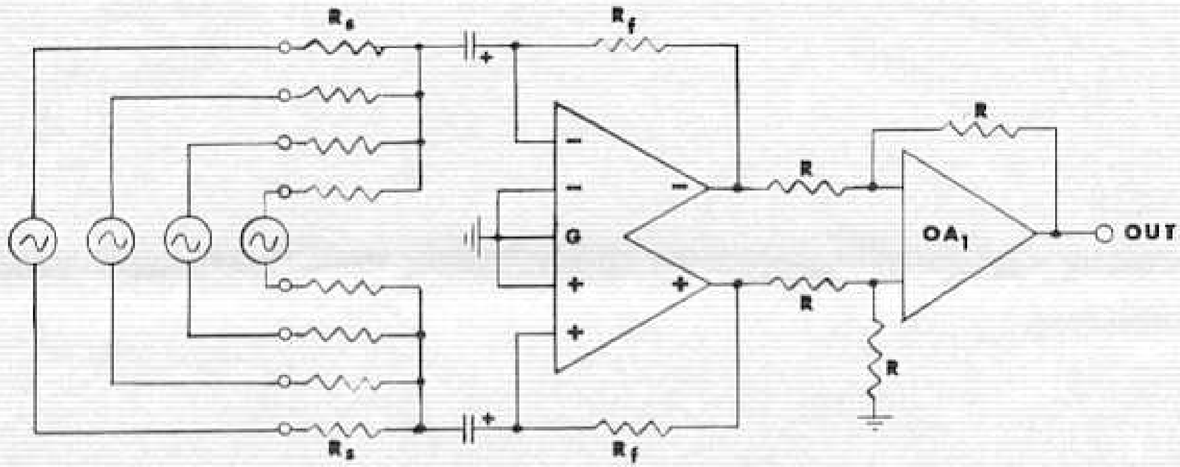


FIGURE 6

As shown if Figure 6, the TRANS-AMP LZ may be configured as a summing amplifier simply by grounding its input terminals and treating its feedback terminals as virtual ground current summing points. In this application, coupling capacitors are required to prevent differences in bias voltages present at the +FB and -FB terminals of the TRANS-AMP LZ to cause DC currents to flow through the sources.

For the circuit shown:

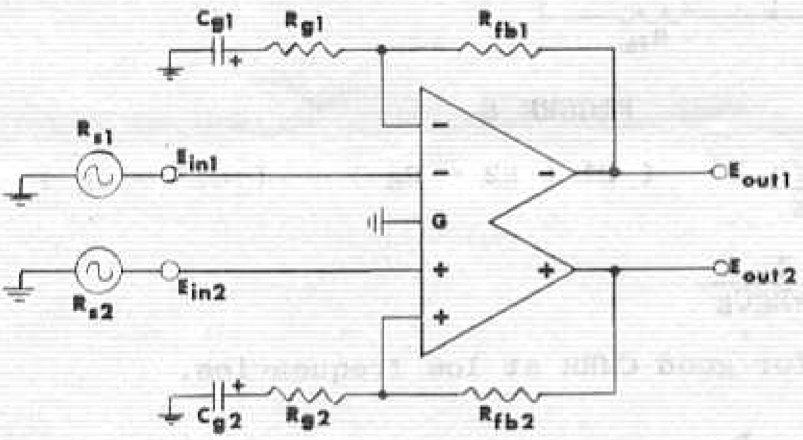
Summing gain = Rf/Rs
Differential input impedence, per input pair = 2Rs
Full power bandwidth: 180kHZ

Specifications, re 100 inputs, Rf = Rs = 10kohm

Noise figure = 1.5dB Broadband, 20HZ to 100kHZ
Buss output noise = 91dBv re .775vrms, 20HZ - 20kHZ
Bandwidth = .8MHZ
Distortion = under .01% IM or THD to 20kHZ
CMRR = Determined by tolerances of Rs and Rf

SINGLE ENDED APPLICATIONS

The TRANS-AMP LZ has many applications as a dual independent amplifier, where the true balanced input feature is not required. In such service, the input noise voltage is decreased by 3dB, with respect to the balanced configuration. Channel separation is typically 90dB, and good power supply rejection is achieved. The basic circuit for a two channel unbalanced amplifier is shown below in Figure 7.



DUAL UNBALANCED AMPLIFIER

FIGURE 7

Gain = $1 + \frac{Rfb}{Rg}$ Low frequency 3dB point = $\frac{1}{2\pi(Rg)(Cg)}$

High frequency 3dB point = Bandwidth ÷ Gain (see bandwidth vs FBR graph)

High frequency bandwidth determining resistance = $\frac{(Rfb)(Rg)}{Rfb + Rg}$

Effective source impedance for noise purposes = Rs + $\frac{(Rfb)(Rg)}{Rfb + Rg}$

Equivalent input noise, see graph (EIN vs source impedance)

EXAMPLE

Assume Rfb = 9.9k, Rg = 100ohm, Cg = 200ufd, Rs = 600ohm

Gain = $1 + \frac{9900}{100}$ = 100 = 40dB Low 3dB point = $\frac{1}{2\pi\ (100)(200)(10^{-6})}$ =7.96Hz

High frequency bandwidth determining resistance = $\frac{(9900)(100)}{9900 + 100}$ = 99ohm

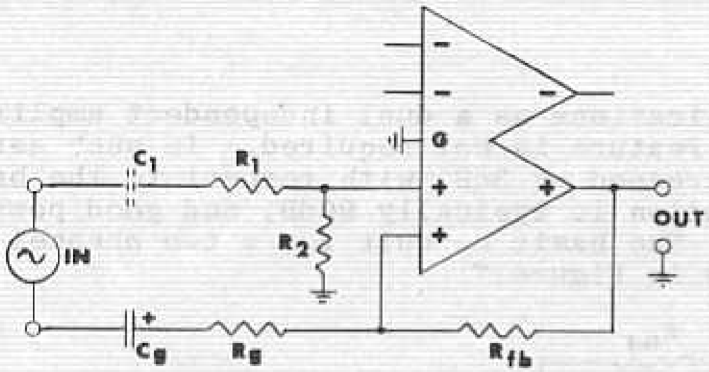
From single section graph, unity bandwidth $\simeq$ 70MHz

High frequency 3dB point = $\frac{70MHz}{100}$ = 700kHz

Effective source impedance = 600 + $\frac{(9900)(100)}{9900 + 100}$ = 699ohm

E.I.N. (from graph EIN vs source impedance) $\simeq$ -124dBv

OTHER DUAL CHANNEL CIRCUITS



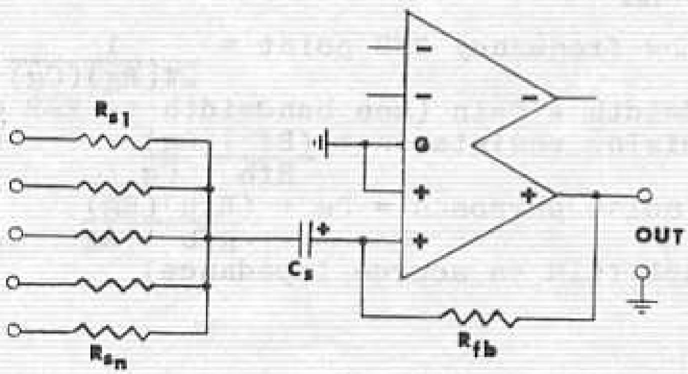
DIFFERENTIAL (GROUND SENSING) INPUT AMPLIFIER

FIGURE 8

Gain = $\frac{Rfb}{Rg}$ ($\frac{R2}{R1} = \frac{Rfb}{Rg}$) (R1 + R2 = Rg) (C1* = Cg)

Low 3 dB point = $\frac{1}{2\pi RgCg}$

* C1 may be required for good CMRR at low frequencies.



UNBALANCED SUMMING AMPLIFIER (ACN), INVERTING AMPLIFIER

FIGURE 9

Gain $\frac{Rfb}{Rs}$

Low 3 dB point = $\frac{1}{2\pi RsCs}$

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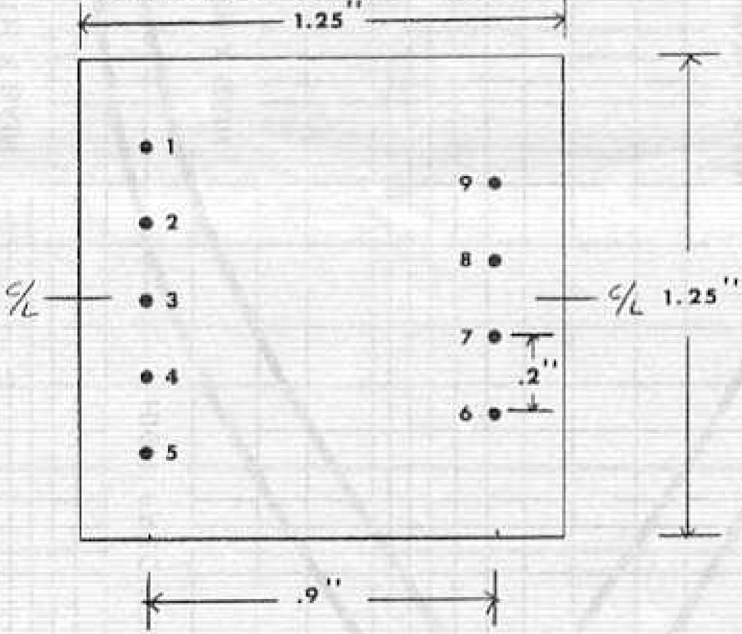
TRANS-AMP™ LZ

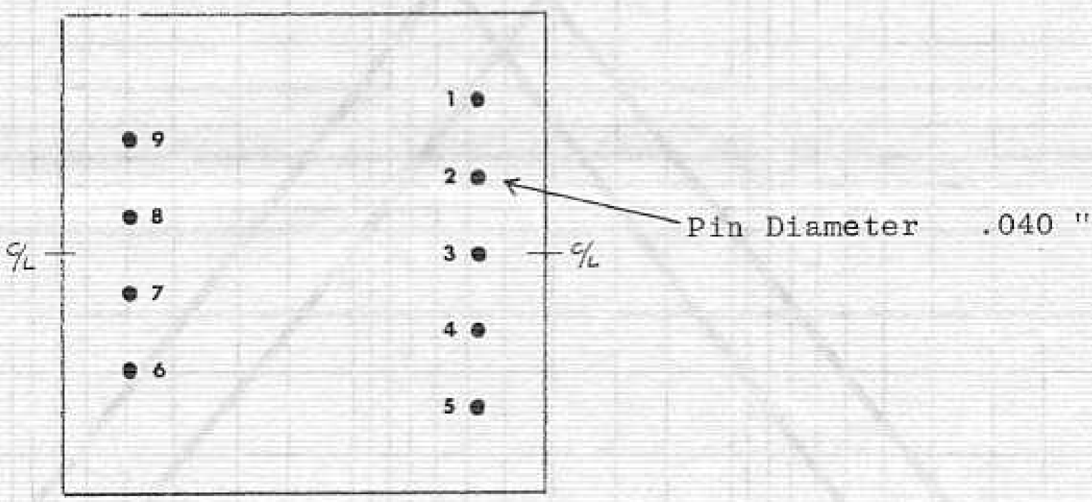
ACCEPTANCE SOCKET LAYOUT



PIN OUT

\# 1 - – F.B.
\# 2 - – IN
\# 3 - GND.
\# 4 - + IN
\# 5 - + F.B.
\# 6 - + OUT
\# 7 - + VOLTS
\# 8 - – VOLTS
\# 9 - – OUT

TRANS-AMP (Bottom View)



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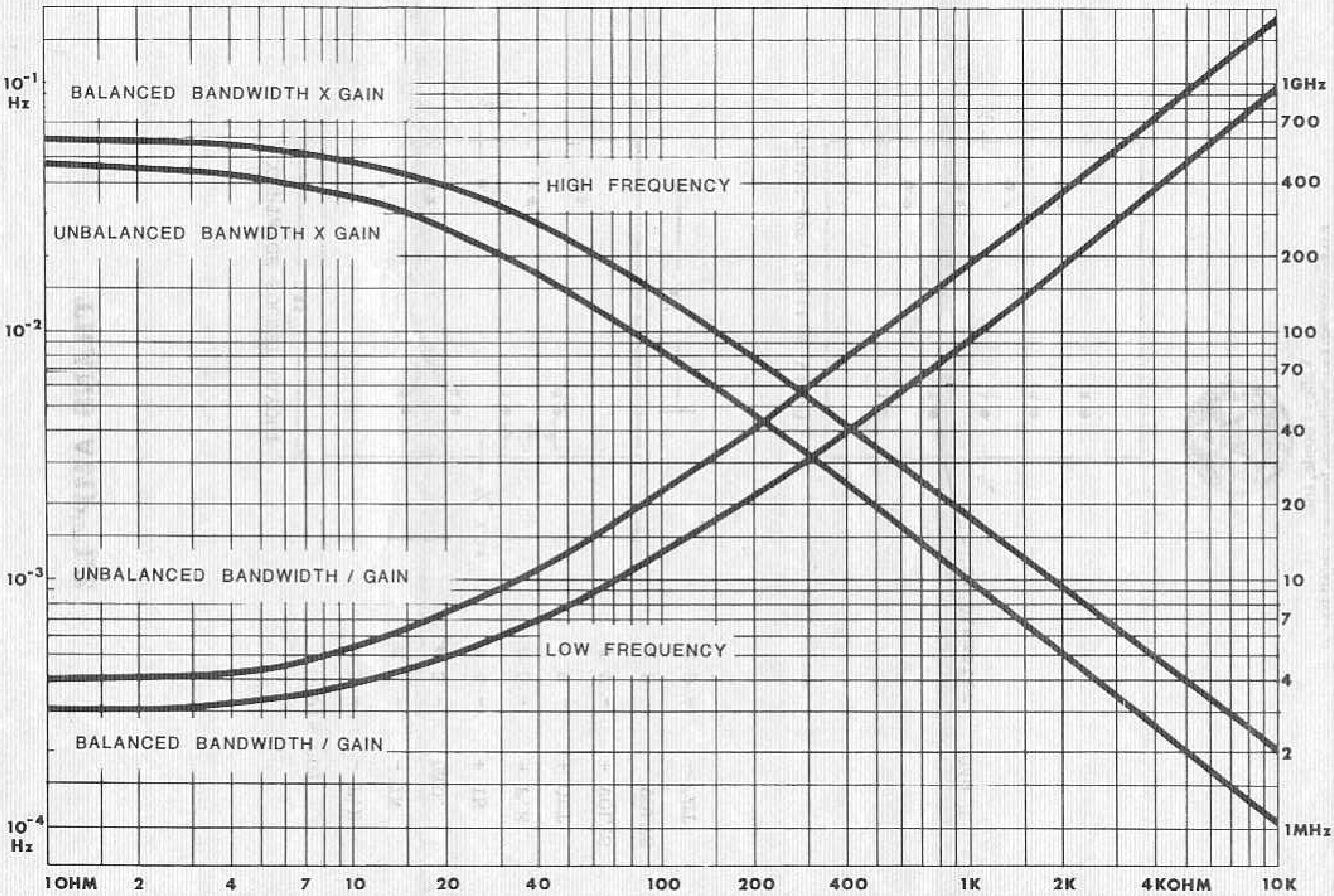

HIGH FREQUENCY BANDWIDTH x GAIN AND LOW FREQUENCY BANDWIDTH / GAIN VS. FEEDBACK PORT IMPEDANCE