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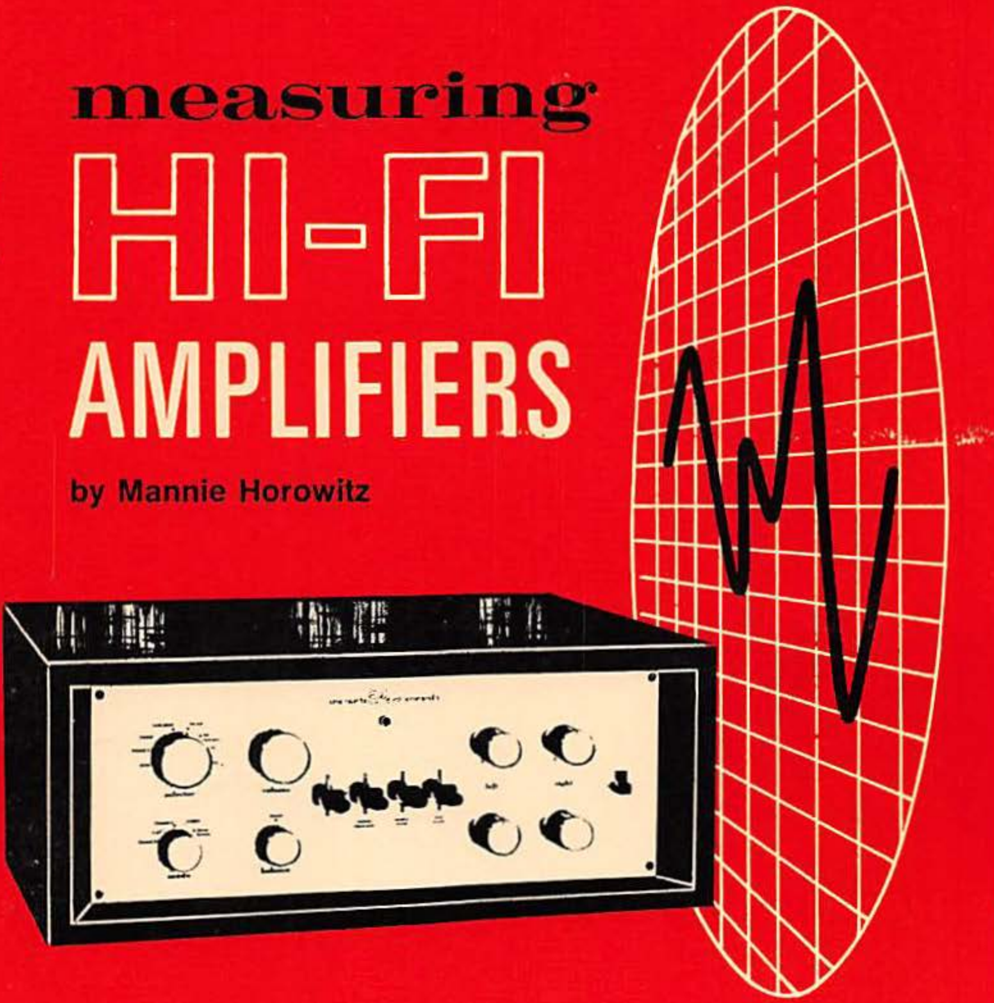
measuring HI-FI AMPLIFIERS

by Mannie Horowitz



MEASURING HI-FI AMPLIFIERS

by

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Preface

High-fidelity audio amplifiers have long been involved in the battle of specifications. Very few amplifiers are sold today that do not have a long list of data which define its characteristics. Practically all amplifiers have been checked and reported on by various test laboratories which have either confirmed or denied the published specifications.

The data presented to the consumer by the manufacturer or independent laboratory require the use of specialized instruments and careful measuring techniques. Although medium-quality instruments may be used to compare two different units with reasonable accuracy, excellent-quality instruments must be used for absolute measurements.

The types of instruments required and available for measuring audio equipment are discussed in the first chapter. A comparison of different types of instruments used to perform the various tests, as well as the characteristics and requirements of a good instrument, are also discussed. The meaning of various pieces of data as well as the laboratory methods used to obtain this information is covered at length. For example—how does peak power differ from rms power, and how do the two differ from music power? Does the peak-power or music-power specification have any significance, or is either one just a bigger number than rms power used primarily for advertising purposes? Is a damping factor of 20 more desirable than a damping factor of 4?

Answers to many questions are contained in this book, which is intended for the technician, engineer, or audiophile. For maximum benefit, a good working knowledge of electronics is desirable. Mathematics is used where required, but may be bypassed without loss of continuity. Detailed deviations are relegated to the 13 appendices at the end of the book, where they may be referred to as desired.

MANNIE HOROWITZ

Contents

Chapter 9

Chapter 10

CHAPTER 1

Instruments for the Laboratory

Several specialized pieces of equipment are used in the audio laboratory. These include a low-distortion audio signal generator, a square-wave generator, a harmonic-distortion meter, an intermodulation-distortion meter, and sometimes a wow-and-flutter meter. More general instruments found in practically every laboratory—audio or otherwise—are a vom, a vtvm, an a-c vtvm, and an oscilloscope.

The various manufacturers supply each of these instruments in different forms. The modes of operation and comparisons of the available types of instruments should help determine the reliability and limits of a particular unit.

SIGNAL GENERATOR

A source of audio signal is essential in any laboratory dealing with high-fidelity amplifiers. Because the characteristics of a sine wave are easily defined and measured, a generator producing waveshapes of this type has become the universal signal source.

An oscillator must meet several obvious basic requirements to be suitable in high-fidelity test and design applications. These can be enumerated briefly as follows:

1. The distortion should be low. A maximum overall harmonic distortion of 0.1 percent between 20 Hz and 20 kHz is usually satisfactory. The percentage of distortion acceptable is only a function of the severity of the test to be performed. Furthermore, the distortion should not be substantially affected by the load presented to the generator.
2. The available frequencies should range from a few hertz to several hundred thousand hertz. This will provide the flexibility re-

quired to check the rolloff characteristics of the amplifier at both ends of the audio spectrum.

3. Frequency stability and calibration accuracy can be lumped together although they are in reality two individual characteristics. Exact frequencies are required when checking marginal and critical equipment, such as a tape deck. Large errors can also result in incorrect filter measurements.
4. A reasonably flat output over the complete frequency range is convenient. Absolute flatness is not required since the output can (and should) be carefully monitored with a wide-range a-c voltmeter or oscilloscope.
5. A low output impedance is a necessity. A range of 600 to 1000 ohms is perfectly acceptable. The frequency-selective networks should be independent of the output load.
6. Any hum present in the signal will be measured as distortion on a harmonic-distortion meter. Hum and noise must be maintained at a minimum. Hum specifications are usually stated as a number of db below the full rated output. Absolute hum voltages are not substantially affected by the setting of the output control. The best signal-to-noise ratios are achieved at or near the maximum setting of the output attenuator.

When extremely low-voltage signals are needed, a divider network is usually placed at the output of the generator. The network shown in Fig. 1-1 can be used effectively for the 2-mv, 1000-Hz signal usually required at the tape-head input of an amplifier. In this illustration it is assumed that the maximum output at the generator terminals is 20 volts.

In order to minimize hum and noise pickup from nearby sources, the leads of both resistors must be kept short, and the connecting cables should be well shielded. Low-loss cable as well as a low-value resistor in the bottom half of the divider network should be used to minimize high-frequency attenuation. The output should be monitored at the amplifier with an a-c meter. Hum may be due to the way

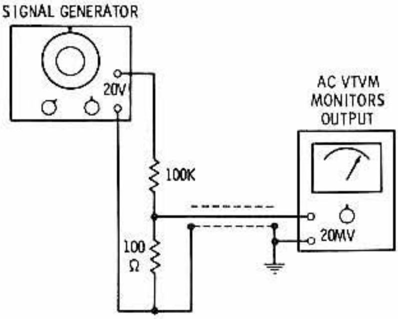


Fig. 1-1. Divider network arranged to get a 2-mv output with low noise and hum from a 20-volt signal source.

the signal generator and monitoring a-c voltmeter are connected in the circuit.

Let us assume that the signal generator is connected to the a-c meter, as shown in Fig. 1-1. Usually these instruments are designed so that the primary of neither power transformer is grounded to its respective chassis. It is also expected that the cabinets of both units will be "floating." This is not the case. The instrument's chassis and cabinets are at some actual fixed potential with respect to the line. The capacity between the transformer windings and the chassis is at a fixed potential with respect to the line because of the power-transformer secondary. The two chassis and cabinets (signal generator and monitoring a-c voltmeter) are connected together through these capacitive couplings to the common a-c line. If the two cabinets should touch, a loop is formed with the a-c line. The alternating current in this loop is induced into the signal leads, causing hum in the signal. Separating the two instruments will eliminate this loop and the resulting hum.

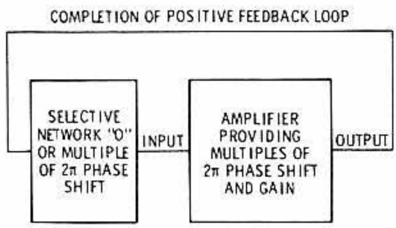


Fig. 1-2. Conventional audio-oscillator schematic.

Another possible source of hum can be the loop formed when the two chassis are connected, and when the commons of the two instruments are joined by leads. This also forms a complete loop susceptible to induced hum. Whatever the cause, the cabinets or chassis of both instruments should be separated physically—they must not touch each other.

The basic component of an oscillator is an amplifier, as shown in Fig. 1-2. Positive feedback around the amplifier causes the circuit to oscillate. A frequency-selective network inserted in either the amplifier or feedback loop determines the frequency of oscillation. Any properly arranged and proportioned amplifier circuit will oscillate if several criteria are satisfied.

1. At the frequency of oscillation, the amplifier and feedback network must have a phase shift of zero (or multiples of 2π).
2. The gain of the overall circuit (amplifier and feedback factor) must be equal to or greater than 1. This is referred to as the Barkhausen criterion.
3. The output voltage is limited by the nonlinearity of the amplifier.

The conventional audio oscillator uses the Wien-bridge circuit shown in Fig. 1-3 as the frequency-selecting network. The amplifier output, E_{in}, is fed to the input of the bridge. The output voltage, E_{out}, from the bridge is fed to the input of the amplifier, completing the positive feedback loop.

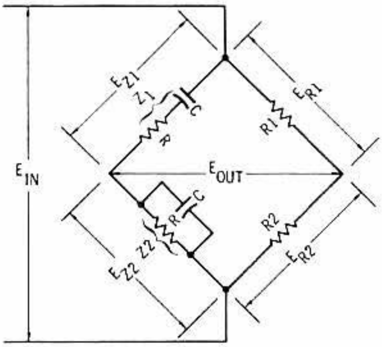


Fig. 1-3. Wien bridge used as frequency-selective network.

In order to satisfy Criterion 1, at the oscillating frequency, E_{out} must be in phase with E_{in}. That this is true can be surmised from the mathematical proof in Appendix A.

From bridge theory, we know that when the bridge is balanced, there will be zero voltage at E_{out}. For the oscillator to operate, E_{out} must not be equal to zero. Therefore, the bridge is slightly unbalanced when it is used in an oscillator circuit. Ordinarily, $R_2/(R_1 + R_2) = Z_2/(Z_1 + Z_2) = 1/3$. When the bridge is slightly unbalanced, the ratio of $R_2/(R_1 + R_2)$ is made slightly smaller than $1/3$.

The circuit has been used fairly consistently in many commercial audio signal generators, as illustrated by the circuit shown in Fig. 1-4. Here the voltage E_{in} is fed from the output of V2 and the voltage E_{out}

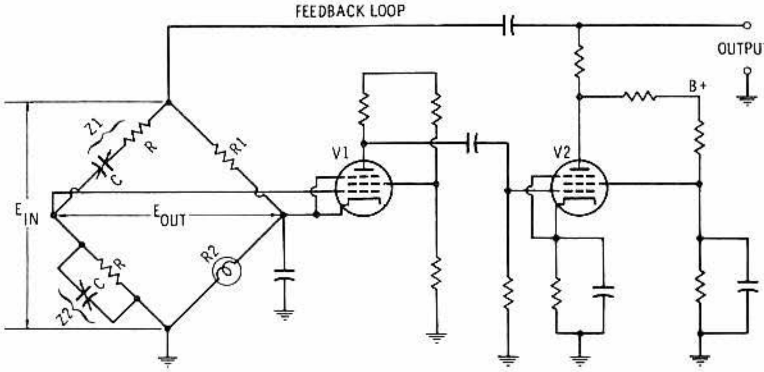


Fig. 1-4. Wien bridge used in commercial signal generators.

is applied between the cathode and the grid of V1. The two capacitors in the Z1 and Z2 arms are varied simultaneously to maintain r-c relationships with their respective resistors. In this way, phase shift is avoided while the required frequency is being selected. The R components of Z1 and Z2 are changed when different ranges are needed. A variable resistor in the form of a lamp filament is substituted for R2. This is done to maintain amplitude stability over the various switched ranges, as well as to guard against variations due to component aging. R1 serves a dual function. First, it completes the bridge circuit. Second, it is incorporated in a negative-feedback loop from the plate of V2 to the cathode of V1, reducing distortion and maintaining stability with tube variations. R1 is adjusted for the best waveform.

Another frequency-selective circuit which is used in many commerically available oscillators is the bridged-T network. Although it is characterized by low distortion and good stability, there has been some tendency toward the Wien-bridge configuration because of the somewhat more practical value of the resistors used. In reality, there is no basic advantage of one circuit over the other insofar as performance is concerned.

There are frequency limits imposed on the Wien-bridge oscillator by its very nature. At high frequencies, the resistors in the bridge arms are too small. They load the output tube excessively as well as introduce phase shift. The lower frequencies are limited by the practical size of the capacitors, C, and the resistors, R. Too high a resistor in the Z2 arm may overbias the grid of V1 because of grid-leak action, as well as make the circuit susceptible to stray-field pickup.

The sine-wave output from the signal generator can easily be converted into a square-wave signal, useful for many audio tests. The circuit shown in Fig. 1-5 is commercially feasible. On the positive half-cycle of the sine wave, grid current flows through R1 and R2. The voltage across R2 acts as a fixed bias for the diode formed by the grid and cathode of V1. The equivalent circuit of this is shown in the diode circuit in Fig. 1-5. When the signal is applied to the diode, only the positive peaks of the signal will be sufficient to overcome the bias voltage and cause the diode to conduct. The conducting diode will be a short for these peaks, resulting in a clipped positive portion. This clipped form will appear at the grid of V1, where it is amplified. The tube clips the negative portion when the high bias voltage drives it to cutoff.

The phase of the signal has been shifted 180 degrees in the plate of V1. The signal is fed to V2, which is cut off for a portion of the cycle, and the negative end of the signal is clipped. The rise time is good because the signal has gone through two stages of amplification and clipping.

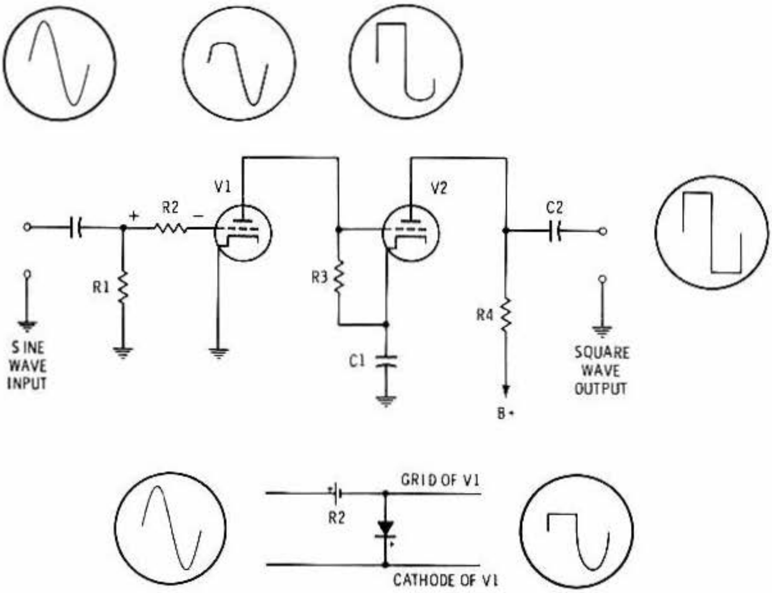
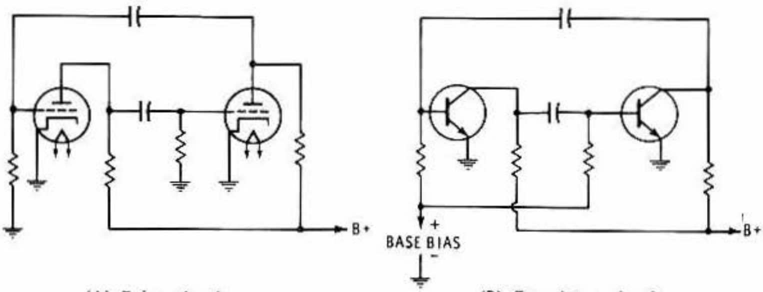


Fig. 1-5. Circuit used to convert a sine wave into a square wave.

Other and more direct means are frequently employed to get a square wave. The popular multivibrator circuit shown in Fig. 1-6 is quite common. Here, one tube or transistor conducts while the second tube or transistor is driven to cutoff. The frequency is determined by the time the voltages across the capacitors take to leak off through the associated grid- or base-circuit resistors, and by the voltage required to cut off the amplifying devices. Symmetrical signals are obtained if both r-c pairs are equal. Excellent rise time and good square waveforms can be obtained using either configuration.



(A) Tube circuit. (B) Transistor circuit.

Fig. 1-6. Multivibrator circuit used to generate a square wave.

HARMONIC-DISTORTION METER

The details of the Wien-bridge circuit can be applied to harmonic-distortion meters as well as to audio oscillators. The operation of a conventional distortion meter is straightforward. The complete signal is fed from the audio amplifier under test to a meter, and the voltage is measured. The fundamental is eliminated from the signal under test, with the result that only harmonics remain. These are now measured on the same voltmeter. The ratio of the harmonics to the total voltage is the amount of distortion in the signal being tested. The Wien bridge is often used as the selective network for the elimination of the fundamental component.

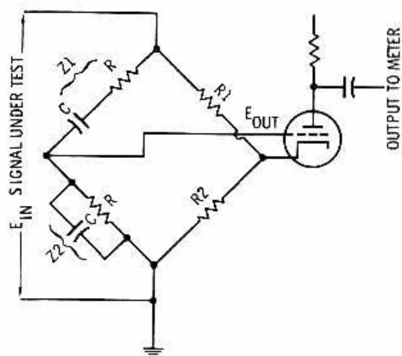


Fig. 1-7. Wien bridge used in distortion analyzers.

The circuit in Fig. 1-7 shows how the Wien bridge eliminates the fundamental frequency while passing the harmonics. Unlike the circuit used in a signal generator, the bridge is completely balanced for the fundamental frequency, $f_o = 1/2\ \pi$ rc. Then $Z1/Z2 = R1/R2$. Under this condition, $E_{out} = 0$ for the fundamental (f_o) frequency. It is not zero at all other frequencies because of phase shift in the bridge. Thus, these harmonic voltages are passed on to the next tube. It is interesting to note that E_{in} is frequently obtained from a cathode follower and that there is a considerable amount of feedback around the bridge circuit. These are important in increasing the rejection of the fundamental frequency.

A variation of this is shown in Fig. 1-8, where a phase splitter drives the circuit. As indicated in the discussion of the oscillator, the ratio of Z1 to Z2 is 1:3. Thus, if voltages of proper bucking phase, but of this 1:3 amplitude ratio, are fed to this network, there will be a null between the junction of Z2 and ground at the fundamental frequency, f_o, only, while all other frequencies will be passed. This method is inferior to the bridge circuit because the null cannot be quite so pronounced.

Using an instrument employing either circuit can be misleading unless the resulting harmonics are observed on an oscilloscope. The meter measures everything except the fundamental. The reading will include hum along with the harmonics. The significance of the hum, as well as the frequency of the undesirable harmonics, can be observed and evaluated on the scope.

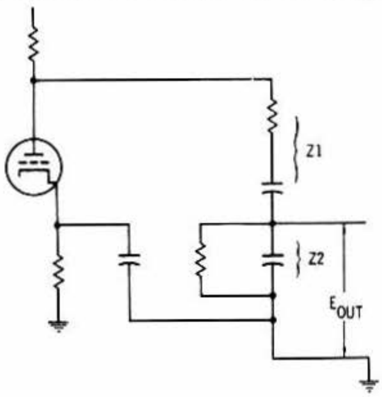


Fig. 1-8. Phase splitter drives dividing network to get proper voltage ratio and zero output when W = 1/RC.

The characteristics of a good harmonic-distortion meter are many, but a few are quite significant to the operator.

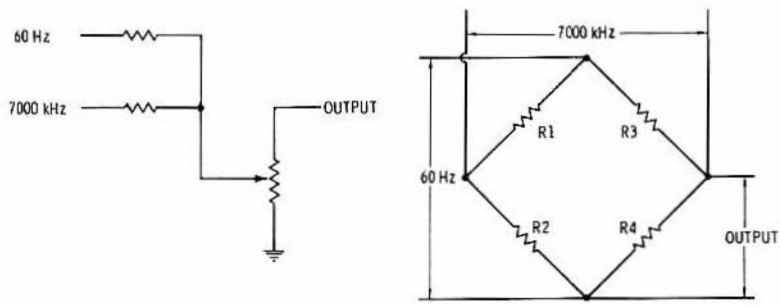
1. The instrument must be capable of measuring harmonics of fundamental frequencies from 20 Hz to 20 kHz. The voltmeter must then be capable of linear response to at least 60 kHz.
2. The fundamental frequency must be reduced considerably. For measurements with up to 0.1-percent distortion, the fundamental must be down about 80 db. However, the closest harmonics, such as the second, should not be reduced by more than 3 db. This is best achieved when the bridge circuit in Fig. 1-7 is used.
3. The instrument must introduce only negligible amounts of distortion and hum.
4. It should be sensitive enough to read 0.1-percent distortion on a 1-volt signal with reasonable accuracy.

A better method of measuring distortion requires the use of a wave analyzer. Here, a voltmeter measures the relative harmonic components in the signal. Thus, the amount of second- or third-harmonic component is checked, rather than the composite sum of all components. In this type of instrument, only one frequency component at a time is fed to, and read on, the voltage-measuring instrument.

Although harmonic-distortion characteristics are practically always stated in the list of audio-amplifier specifications, intermodulation-distortion measurements have gained in significance because of the excellent correlation with actual listening tests.

To test intermodulation distortion, two signals of different amplitude and frequency are passed through the amplifier under test. These two signals will appear at the output of an undistorted amplifier. If nonlinearity does exist in the unit under test, the two input signals will heterodyne and produce sum and difference frequencies along with the original two frequencies.

The conventional intermodulation-distortion meter supplies these two signals in a 4-to-1 amplitude ratio. The frequencies commonly used are 60 Hz and 7000 Hz respectively. Two methods are generally used to combine these frequencies in the analyzer before they are fed to the amplifier. The first mixes the two frequencies in the type of adding network shown in Fig. 1-9A. The combined signal appears across the potentiometer. A second method, shown in Fig. 1-9B, uses a bridge for mixing.



(A) Adding network. (B) Bridge network.

Fig. 1-9. Network for mixing two frequencies.

The latter configuration is usually preferred because in the first method the signals can interact with each other. A second factor in favor of the balanced bridge is that it isolates the two signals. Balance is maintained when all resistors retain the proper relationship to each other. The signal must be attenuated by means of a T-pad at the output, so that R4 will maintain its resistance regardless of the amplitude required at the output. The bridge will then be balanced at any output level.

Operation of the rest of the instrument is straightforward. The combined signal is connected to the input of an amplifier. The output from the amplifier is sent through a high-pass filter, eliminating the 60-Hz component. If the amplifier has introduced distortion, the remaining 7000 Hz will appear to be modulated by a low frequency. The amplitude of the total modulated 7000-Hz signal is measured. The signal is then detected with only the modulating frequency remaining. This is in turn measured and compared with the amplitude of the 7000 Hz. The ratio of the two is the percentage of intermodulation distortion.

This section of the analyzer is quite important. The accuracy of the measuring instrument and the quality of the filters determine the validity of the measurements. Low-hum circuitry is necessary to eliminate any stray signals. The analyzer input impedance must be relatively high (500,000 ohms will do) to avoid putting an excessive load on any preamplifier under test.

THE OSCILLOSCOPE

The oscilloscope is a visual voltmeter used to observe the output of the equipment under test. Unfortunately, low percentages of distortion are not too obvious on the scope screen, and the measuring instruments described previously must be used. A scope display can be used as an effective indication of some amplifier characteristics. It can also supplement the information gained from other instruments.

The heart of the oscilloscope is the cathode-ray tube. In this tube, a potential difference between the cathode and one of the other electrodes starts the electrons in motion toward the fluorescent screen. The groups of deflection plates are arranged so that each pair is perpendicular to the other. A potential applied to these individual pairs of plates deflects the electron stream. The horizontal pair of plates deflects the electron stream vertically toward the more positive of the two plates. The vertical pair of plates provides the equivalent horizontal deflection. The electron stream hits the screen, causing the fluorescent material to glow. Under proper conditions, the waveshapes applied to the deflection plates will appear on the screen.

The amount of deflection on the screen is determined by the sensitivity of the cathode-ray tube. This is usually stated in inches or centimeters of deflection per volt. The sensitivity of the scope can be increased by providing voltage amplification. Amplifiers employed in a scope useful in audio testing procedures must provide linear response from d-c to about 500 kHz. They should also respond faithfully to square waveforms and present no phase shift over much of the range. The sensitivity should be great enough to show the hum components present in a piece of audio equipment—10 millivolts per inch should be a satisfactory sensitivity after amplification.

The vertical amplifier is the more critical of the two, although the horizontal amplifier must also be undistorted and have a wide frequency range. The horizontal amplifier need not have flat response at the extremes. To display a two-dimensional signal on the screen, a varying voltage must be applied to the horizontal as well as to the vertical plates. A sawtooth signal is applied to the horizontal amplifier so that the vertically applied signal may be swept across the screen. This sawtooth must be linear and variable to about 100 kHz to be capable of displaying any signal significant in audio tests.

In general, a good scope has a thin, bright trace; the sweep oscillator is easy to synchronize with the incoming signal and will usually provide a signal of known amplitude to enable easy calibration of the screen in volts per inch or centimeter of deflection.

Variations have appeared by the dozens. One of the most useful is the dual-trace scope for displaying two signals at one time. This is particularly applicable in stereo tests. A dual-trace scope usually has a self-contained multivibrator in the form of an electronic switch. The scope alternately sweeps two signals applied at the input and displays them in sequence on the screen. The persistence of the fluorescent material makes both traces appear as though they are being viewed simultaneously at different vertical positions on the face of the cathode-ray tube.

Dual-trace scopes can also be made using special dual-beam tubes. There are two completely independent sets of electron guns and deflection plates. Two independent amplifiers are built into the oscilloscope. One amplifier sets up the voltages at one pair of deflection plates, and the second amplifier is for the second pair of deflection plates.

A-C VOLTMETER

The a-c voltmeter was mentioned throughout this discussion in several different applications. It is the basic instrument in any test setup. All available a-c voltmeters fall roughly into two groups. The most common type consists of a wide-band a-c amplifier. Its output is rectified and fed to a d-c meter movement. Although these meters are calibrated in sinusoidal rms voltage, they are actually sensitive to the average values.

A second and less expensive type peak-rectifies the signal. The resulting d-c voltage is then fed to a d-c amplifier or bridge. Once again the scale is calibrated in rms, but this time the unit is sensitive to peaks. A variation of this is the peak-to-peak reading voltmeter. Both units are useful in the laboratory. The latter types are relatively insensitive and are frequently incorporated as a portion of a general vtvm used for checking d-c voltages and resistance along with the a-c ranges.

The former type usually consists of several amplifier stages. Feedback is provided through the meter circuit, around all of these stages. This contributes considerably to stability, linearity, and wide frequency range. The power supplies must be well regulated to avoid reading variations due to line-voltage fluctuations.

The second type of meter can also have many variations. One form of peak-reading meter circuit is shown in Fig. 1-10. Here, the applied voltage is rectified and the ripple passes through resistor R during positive portions of the cycle. Capacitor C smooths out this ripple so that only the peaks remain. This circuit may be used as the rectifying

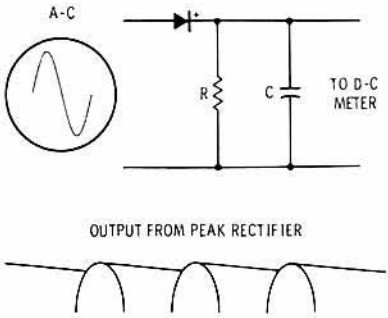
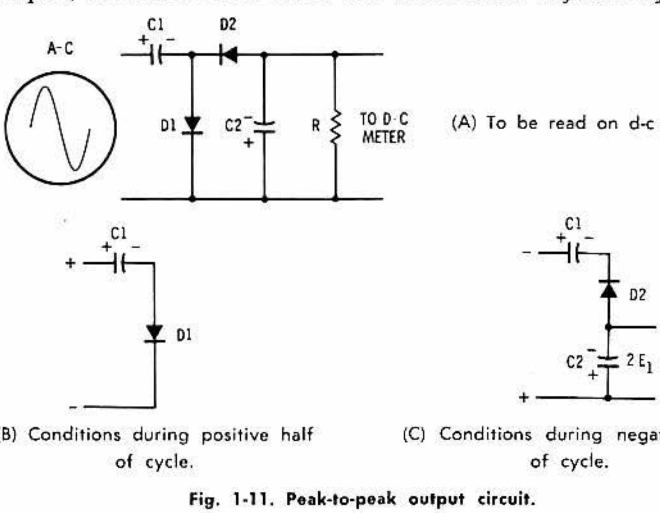


Fig. 1-10. Peak-reading circuit for use with d-c meter.

circuit for the a-c ranges, and can be included as part of any d-c vtvm.

The peak-to-peak reading meter is most useful in audio work. A circuit of this type appears in Fig. 1-11. The operation is fairly obvious. During the positive half-cycle, diode D1 conducts and a d-c voltage is built up on C1, as shown. During the negative half-cycle, the negative portion of the cycle is added to the voltage across C1, through D2. The sum of these appears across C2 and R, and in turn is fed to the d-c meter.

The meter movement can be designed into a d-c amplifier bridge circuit, as shown in Fig. 1-12. When the currents through both tubes are equal, the meter reads zero. The currents are adjusted by varying



(A) To be read on d-c meter.

(B) Conditions during positive half of cycle.

(C) Conditions during negative half of cycle.

Fig. 1-11. Peak-to-peak output circuit.

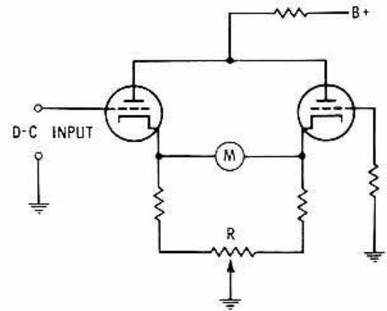


Fig. 1-12. A d-c reading meter in a balanced circuit.

resistor R in the cathodes. This resistor adjusts the relative bias on the two tubes and, consequently, the relative plate (or cathode) currents. Once adjusted, an unknown d-c input signal upsets the bias on the first triode only, resulting in a deflection of the meter.

Both types of meters are useful. The average-reading type is usually more accurate and more stable. For this reason, it should be used for monitoring the input signal to, and the output signal from, the amplifier under test. The peak-to-peak reading type is useful for measuring peak amplifier outputs. One note of caution: ordinary vom's are frequently provided with a-c scales. Often these scales are nonlinear because the rectifiers in the vom are frequency-sensitive. Thus, the vom should not be used where accuracy is essential.

THE FLUTTER METER

The flutter meter is not an audio instrument in the conventional sense of the term. However, because tape recorders and phono turntables are important parts of an audio system, this type of instrument has become most common in the audio laboratory.

Flutter and wow are variations in the speed. For testing tape recorders, a steady single frequency, usually 3000 Hz, is recorded on tape. A speed variation appears as a frequency variation around 3000 Hz, which is not unlike in the frequency-modulated signal. When the test tape is played on a tape machine, the output signal of the machine is passed through a filter where all frequencies except the 3000 Hz pass through. The remaining signal is then f-m detected and read on an ordinary a-c voltmeter.

Once again, there are several important characteristics of a good instrument. First, the filters must be sharp. The circuit must eliminate all extraneous amplitude variations to prevent them from affecting the final reading. Filters should be provided to separate the wow (slow-speed variations) components from the flutter (rapid-speed variations) components. As always, the power supply must be extremely well regulated to avoid reading variations with line fluctuations.

Other instruments which are useful and are basic components in most laboratories are the tube tester, the capacitance bridge, the inductance bridge, and the vom. Discussion of these might be of interest, but their application to actual audio measurements on audio equipment is strictly limited.

CHAPTER 2

Frequency-Response Specification

The first characteristic most likely to be recognized by the audiophile in relation to an audio amplifier is frequency response—which can be defined as the relative gain of the unit over a range of frequencies. The significance of this yardstick has not waned with time, but other amplifier characteristics have assumed a place of equal importance.

The importance of a flat frequency characteristic requires little discussion. It is quite obvious that for accurate sound reproduction, all frequencies should be given "equal opportunity." Any frequency presented to the input of an amplifier should be amplified the same amount as any other frequency simultaneously presented at the same input. There are several important exceptions to this ideal.

The output from an equalized phonograph or tape-head preamplifier is not uniform; records and tapes are recorded to adhere to a specific curve wherein some frequencies range. During playback, the amplifier must compensate for these frequencies in order to provide an overall flat response from the source (phonograph record or prerecorded tape), the transducer, and the amplifier. In this chapter the measurements of frequency response from the tuner input of the preamplifier through the power-output section will be discussed. The characteristic must be reasonably flat when only these sections are considered.

Another consideration is the frequency range desired from the amplifier in question. While many units will have a flat response to several octaves on either side of the audio spectrum (assumed here to be 20 Hz to 20 kHz), some amplifiers are designed for limited bandwidth in the interest of increased stability and reduced noise. The latter factor is especially true in transistorized units, where bandwidth limitations are required to keep noise measurements comparable with actual audible noise reproduction. Frequency response is usually measured in decibels (db) although it can also be measured in terms of voltage or power. In the latter cases, the numbers would become astronomical.

DECIBEL

The decibel is defined by the simple equation:

$$db = 10 \log_{10} P_o/P_i \qquad (2\text{-}1)$$

where P_o is the output power from an amplifier and P_i is the input power. Putting this equation into another form, with the logarithmic base being 10, yields:

$$db = 10 \log P_o - 10 \log P_i \qquad (2\text{-}2)$$

During the frequency-response check, the voltages (V_i) fed to the amplifier must be maintained at a constant level for all frequencies. It is assumed that the input impedance (R_i) of the amplifier is not frequency-sensitive. The latter condition can be assured by feeding the signal from the low-impedance source. The input power, P_i, is thus constant at all frequencies because it is equal to V_i^2/R_i, two constants. The term 10 log P_i in the next equation, 2-3, can be replaced by a constant. We will call this constant K.

In these tests, all measurements revolve about the 10 log P_o term. In the actual test procedure, the K term is adjusted for a specific power reading at the output of an amplifier, for some frequency in the middle of the audio range. The central frequency is usually 1000 Hz or 400 Hz. Ten times the log of the output power at all other frequencies is compared with this reading at 1000 or 400 Hz. In the following discussion, 1000 Hz is used as the reference frequency, "db (1000 Hz)" is the gain at 1000 Hz, and "db (100 Hz)" is the gain at 100 Hz. Similarly, "10 log P_o (1000 Hz)" is 10 times the log of the output power at 1000 Hz. At 100 Hz, the expression becomes "10 log P_o (100 Hz)."

The equation for gain at 1000 Hz is:

$$db\ (1000\ Hz) = 10 \log P_o\ (1000\ Hz) - K \qquad (2\text{-}3)$$

As an example, let us find the difference in gain (in db) at 100 and 1000 Hz. First, write the counterpart of the above equation 2-3 for 100 Hz:

$$db\ (100\ Hz) = 10 \log P_o\ (100\ Hz) - K \qquad (2\text{-}4)$$

The db variation at 100 Hz from the reading at 1000 Hz is found by subtracting equation 2-3 from equation 2-4 if the gain at 100 Hz is greater than the gain at 1000 Hz, or subtracting equation 2-4 from equation 2-3 if the gain at 1000 Hz is greater than the gain at 100 Hz.

Subtracting equation 2-4 from 2-3, as follows:

$$\text{db}\ (1000\ \text{Hz}) = 10 \log P_o\ (1000\ \text{Hz}) - K$$
$$- [\text{db}\ (100\ \text{Hz}) = 10 \log P_o\ (100\ \text{Hz}) - K]$$

results in the following expression.

$$\text{db}\ (1000\ \text{Hz}) - \text{db}\ (100\ \text{Hz}) = 10 \log P_o\ (1000\ \text{Hz}) - 10 \log P_o\ (100\ \text{Hz}) \qquad (2\text{-}5)$$

The input term drops out in the final equation. The resulting equation involves only the deviation of the log of the output power at 100 Hz from the log of the output power at 1000 Hz.

Another way of expressing the difference in gain at 1000 Hz and 100 Hz, is:

$$\Delta\ \text{db} = 10 \log \left(\frac{P_o\ (1000\ \text{Hz})}{P_o (100\ \text{Hz})} \right) \qquad (2\text{-}6)$$

where Δ db is the difference in gain between 1000 and 100 Hz.

The measuring circuit at the output of an amplifier takes the form shown in Fig. 2-1. The output power is developed across a load resistor, R_L, and measured on a wide-frequency-range a-c voltmeter. The power across the resistor is, of course V_o^2/R_L, where V_o is the output voltage reading on the a-c meter.

A straightforward procedure consists of measuring the output voltages at 100 and 1000 Hz, calculating the power at each frequency from V_o^2/R_L, and substituting these into equation 2-6 to determine the db difference at the two frequencies. Converting the equation to read directly in voltage would be much simpler, saving two calculations.

Consider the output power at 1000 Hz to be equal to $P_o\ (1000\ \text{Hz}) = V_o^2\ (1000\ \text{Hz})/R_L$ and the output power at 100 Hz to be equal to $P_o\ (100\ \text{Hz}) = V_o^2\ (100\ \text{Hz})/R_L$. Substituting these into equation 2-6 would yield:

$$\Delta\ \text{db} = 10 \log \frac{V_o^2\ (1000\ \text{Hz})/R_L}{V_o^2\ (100\ \text{Hz})/R_L}$$
$$= 10 \log \left(\frac{V_o\ (1000\ \text{Hz})}{V_o\ (100\ \text{Hz})} \right)^2$$
$$= 20 \log \frac{V_o\ (1000\ \text{Hz})}{V_o\ (100\ \text{Hz})} \qquad (2\text{-}7)$$

Equation 2-7 can be used, assuming that R_L at 1000 Hz is equal to R_L at 100 Hz. This is generally true if the load resistor used in the test is noninductive. This formula does not hold true if a speaker load is used, for the load varies with frequency. All tests on amplifiers are made assuming a constant load at the output for all frequencies.

In equation 2-7, db is expressed as a ratio of two voltages. If one voltage is known, Δ db can be calculated for any other voltage from the equation. These db values, representing different relative voltages, can be printed on the meter face and read directly as in Fig. 2-2.

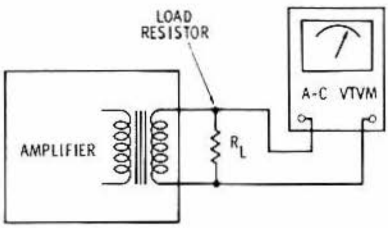
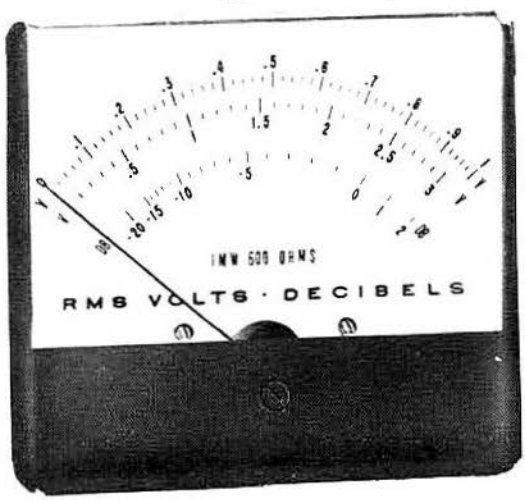


Fig. 2-1. Output-measuring circuit.

Reading db variation on this scale is obvious. Set the output for 0 db at 1000 Hz on a convenient range. Read the deviation from this 0 db at any other frequency, directly on the scale. If the voltage is on the next higher range, add 10 db to the original reading; if you must switch to the next lower range, subtract 10 db. Every time you switch from the original reference range, you either add or subtract 10 db per range, depending on whether the output is higher or lower than the original. If you use other than the 0 db as the reference voltage, all other readings must be referred to this new reference as if it were 0 db. Thus, if −2 db were the reference reading at 1000 Hz,



Courtesy Eico Electronic Instrument Co., Inc.

Fig. 2-2. Standard meter face.

a −4-db reading at 100 Hz indicates a loss in gain of 2 db, and a +2-db reading at 10 Hz indicates an increase of 4 db.

Several factors may be observed when comparing the voltage and db scales. Doubling the voltage is the same as a 6-db increase, while cutting the voltage in half is a 6-db decrease. Doubling the doubled voltage indicates a second 6-db increase, or a total of 12 db more than the original. Doubling the original voltage three times indicates an 18-db increase over the original reading (6 db + 6 db + 6 db). A voltage factor of 10 is a change of 20 db. A value of 26 db (20 db + 6 db) indicates a voltage multiplication of 20: a multiplication by 10 is 20 db, and a multiplication by 2 is 6 db, and $2 \times 10 = 20$, or

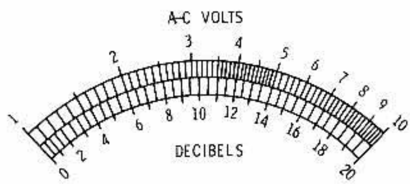


Fig. 2-3. Logarithmically expanded scale.

6 db + 20 db = 26 db. While numbers are multiplied, the db factors are added.

Another type of a-c meter, extremely popular in the audio field, uses a suppressed-zero movement, as shown in Fig. 2-3. The voltage scale does not start with zero and is essentially logarithmic. Each time the range is switched, it represents a change of 20 db rather than the 10 db of the instrument used with Fig. 2-2.

THE MEASURING CIRCUIT

As indicated, the first step in measuring the relative gain or frequency response is to maintain a constant input voltage at all frequencies. As shown in Fig. 2-4, a meter is connected at the input to the amplifier to monitor the voltage fed from the signal generator. The output from the generator should be readjusted or checked each time the frequency is changed, to maintain the input to the amplifier (as read on the input meter) constant at all frequencies.

Feed the signal from the oscillator to an unequalized input on the amplifier. This is usually marked "tuner" or "auxiliary." Adjust all controls on the amplifier to get an optimum flat response. If a preamplifier is involved, the tone controls, loudness or contour controls, and the scratch and rumble filters are all set so that there is no compensation introduced. Turn all level controls to their maximum output position.

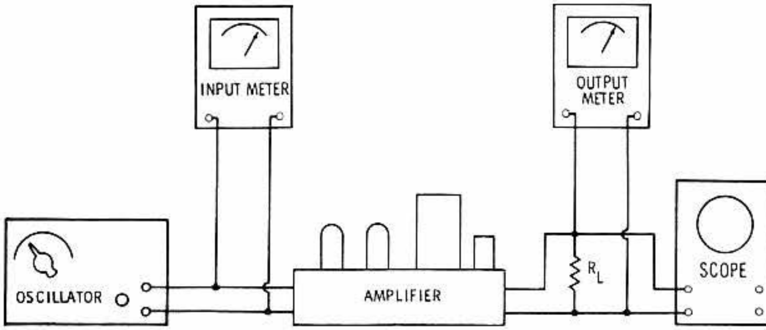
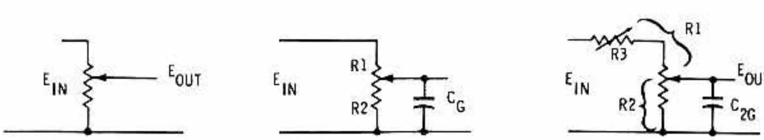


Fig. 2-4. Circuit used for measuring frequency response.

A level control can be considered as a resistive voltage divider, as shown in Fig. 1-5. A more exact representation of the level control as it is commonly used in the grid circuit of a vacuum tube is shown in Fig. 2-5B. C_G represents the total capacity between the grid and cathode. It is equal to the grid-to-cathode capacity added to the product of $(K + 1)$ and the grid-to-plate capacity (Miller effect). K is the gain of the tube. Transistor units are also subject to the Miller effect. In the common-emitter mode of operation, the total input capacity is approximately equal to the base-emitter capacity added to the product of the voltage gain $+1$ $(K + 1)$ of the circuit and the base-to-collector capacity of the transistor.

It can be shown from Fig. 2-5B that the high-frequency response is a function of the control setting. The derivation of this is delegated to Appendix B. It proves that the frequency response is a function of the relative values of R1 and R2.



(A) Representation of a level control in a circuit.

(B) Representation of a level control in a conventional stereo circuit.

(C) A balance control in a stereo circuit.

Fig. 2-5. Ideal level control.

This situation is even more serious in stereo amplifiers. A potentiometer is usually placed in series with e_{in}, used for balance between the two channels. The frequency response must roll off at the upper end of the band when this configuration exists, because R3 in Fig.

2-5C behaves as if it were part of R1. In testing this type of amplifier, it is proper to set the level controls at maximum, and the balance control for equal output from both channels. The response cannot be as flat at the upper end of the band as was the case with monophonic units. Because the rolloff is slow and usually starts at about 10 kHz, the effect will probably not be audible.

Continuing with the mechanical features of the test procedure, choose a convenient output impedance on the power amplifier and place the load resistor across it. The 8-ohm output terminals are usually used. Connect an 8-ohm, 25 watt, noninductive resistor across these terminals. The power developed across this resistor is measured in terms of voltage on a wide-range a-c voltmeter placed across this resistor. (The readings may be converted to power if desired, using the V^2/R formula, where R-8 in the example cited.) Place a scope across the load resistor. This last step does not result in actual data, but is required for monitoring the waveshape. An essentially sinusoidal output is required if the meter readings are to be significant.

The actual readings can now be made. Set the signal generator for a specific reading on the db meter at 1000 Hz. Switch to all other significant frequencies (from 10 Hz to 40 kHz or more) and read the deviation from the original db setting. It must be remembered that the measurement is for frequency response—not power response. The output must be adjusted so that the signal will not distort at any frequency under test. A 1-watt level is usually satisfactory. When the signal begins to distort, the reading is no longer valid. Start the test again at some lower output and repeat the measurements. Only then can you be certain that you are reading frequency response rather than power response.

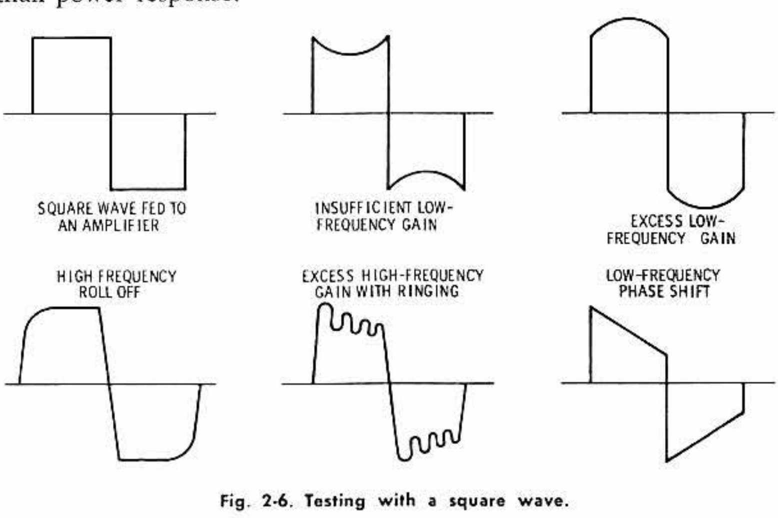


Fig. 2-6. Testing with a square wave.

The frequency response should be a smooth curve over the complete range. Any peaks are usually an indication of a tendency toward instability. Peaks (of about 2 db or more) within the audible range of 20 Hz to 20 kHz add undesired effects to the reproduced sound. The much disputed "presence peak" at about 2000 Hz is said to add to the realism—but the purist will certainly disagree.

A square-wave test can provide a rough indication of the frequency response. Fig. 2-6 illustrates how an amplifier may affect a square wave. Tilt and other variations of the waveshape are possible and may be observed, but they have more significance in describing the phase shift rather than the frequency response. The rise time of a square wave is a fairly accurate check on the upper limit of an amplifier's frequency response. A high-frequency square wave is illustrated in Fig. 2-7. This may be considered as the form assumed after having passed through an amplifier. It is actually a plot of output voltage against time.

The theoretical square wave has a zero rise-time—that is, it takes zero time for the voltage to rise from zero to maximum. However, when a square-wave signal is sent through an amplifier, a finite length of time will elapse from the instant the rise starts until the peak output voltage is reached. This is due to limited bandwidth. This passage of time will elapse from the instant the rise starts until the peak outventionally defined as the time the signal takes to rise from 10 to 90 percent of its final value. The passage of time is shown at T_2 and t_1 in Fig. 2-7.

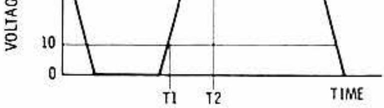


Fig. 2-7. Square wave with measurable rise time.

The relationship between the high-frequency limit and rise time can be found from the equation.

$$f = \frac{0.35}{t_r} \qquad (2\text{-}8)$$

where,

t_r is the rise time,

f is the frequency at which the high frequency has rolled off 3 db. Equation 2-8 is proven in Appendix C.

The rolloff curve with respect to the 3-db point is shown in Fig. 2-8. Note that after the response has dropped 12 db, it continues to decrease by a fixed 6 db per octave or 20 db per decade. Unfortunately, the rise-time measurement cannot be readily made on all oscilloscopes found in the average laboratory; it must be made on scopes in which the horizontal axis has been calibrated in time. Only on these more expensive types of equipment can this test be made accurately.

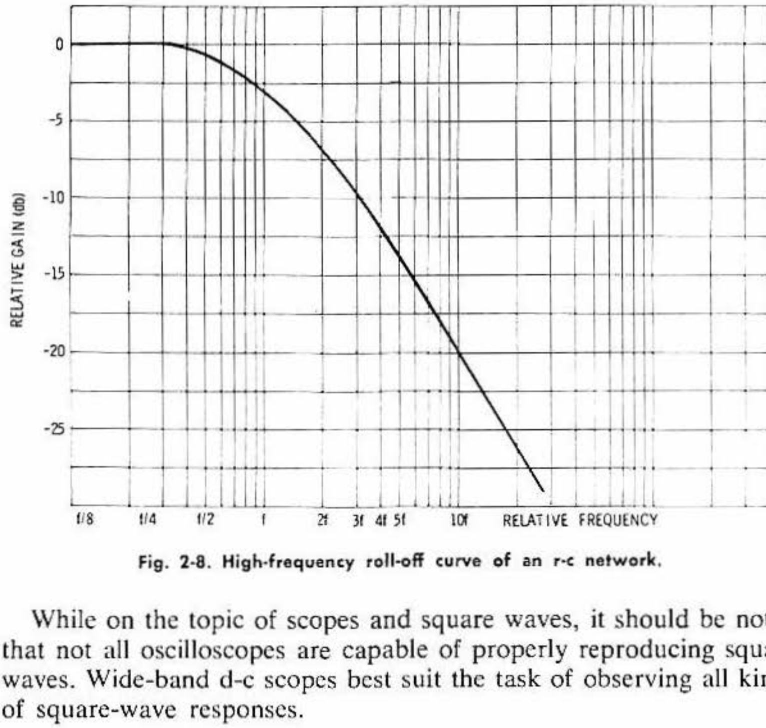


Fig. 2-8. High-frequency roll-off curve of an r-c network.

While on the topic of scopes and square waves, it should be noted that not all oscilloscopes are capable of properly reproducing square waves. Wide-band d-c scopes best suit the task of observing all kinds of square-wave responses.

MEASURING PREAMPLIFIERS

In general, the test procedure and setup for measuring a preamplifier is identical to that shown in Fig. 2-4. One important exception must be considered. An 8-ohm load has been placed at the output of the power amplifier. This is an extremely low impedance. Any normal capacitance due to instruments such as the a-c voltmeter, oscilloscope, distortion analyzer, and so on, is negligible.

The output of a preamplifier is at a relatively high impedance. It should be loaded by a 100,000-ohm resistor in parallel with a 1000-pfd capacitor. The capacitance due to the instruments as well as the connecting leads may have a considerable effect on the frequency response. For this reason, all instruments not actually involved in the test should be disconnected. The connecting leads should be made of low-impedance single-conductor shielded cable, and kept as short as practicable.

Frequency response is an extremely important characteristic of an amplifier, but it should be considered in its true perspective. Just as a wide frequency response does not necessarily indicate an excellent unit, a limited bandwidth does not necessarily indicate a poor amplifier. Either extreme can be a detriment as well as a benefit. A good design is concerned with all factors, and the best compromise is achieved only after everything involved in proper audio reproduction is considered.

CHAPTER 3

Measuring and Matching the Phono Equalization Curve

Phonograph cartridges may be divided into two major categories. One type produces an output voltage proportional to the amplitude of the signal recorded on a disc. The second, and probably the more popular with audiophiles, provides a signal dependent on the velocity of the stylus in the recorded groove. In each category there can be found good-quality and poor-quality units for both stereo and monaural.

RECORDINGS AND CARTRIDGES

Group 1 contains crystal, ceramic, and semiconductor types. Their output is proportional to the recorded amplitude in the record groove. All frequencies recorded with equal amplitudes would theoretically produce constant voltages at the terminals of the cartridge. If the recorded amplitude were doubled at any frequency, the output voltage would also double, depending on the quality of the specific cartridge and the amplifier circuitry provided for reproduction.

Velocity-sensitive phonograph cartridges in group 2 are exemplified by magnetic (variable-reluctance or moving-coil) pickups. The output voltages they produce can be understood with the help of Fig. 3-1. Assume the frequency of recorded signal A is 1000 Hz and that of recorded signal B is 2000 Hz. Both signals have been recorded with equal amplitudes in the record groove. The stylus must travel twice the distance to trace curve B as it does to trace curve A. The output from curve B will be double that of curve A, despite identical recorded amplitudes, because the stylus traveled twice the distance during the same interval. From this it can readily be inferred that

Fig. 3-1. Signal recorded in grooves.

during reproduction, the output voltage is proportional to the frequency, assuming a constant recorded amplitude.

The velocity-type cutter is generally used to record discs. The voltage applied to record signal B must be double that used for signal A, if both are to have the same groove amplitude on the recorded disc. Stated more analytically, to record a constant amplitude using a velocity-type recording head, the voltage applied to the head must vary directly with the recorded frequency.

Curve A-A in Fig. 3-2 shows the amplifier compensation required for recording a uniform amplitude with a velocity-type cutter. Curve B-B shows the equivalent curve required to reproduce the constant recorded amplitude using a constant-velocity cartridge. One is the mirror image of the other about the zero axis. Note its 6-db-per-octave characteristic. For completeness, C-C is drawn as the recording and reproducing curve using constant-amplitude cartridges.

Recordings can be made using either the constant-amplitude or constant-velocity characteristic, or a combination of both. Assume for the moment that only the constant-velocity characteristic were used

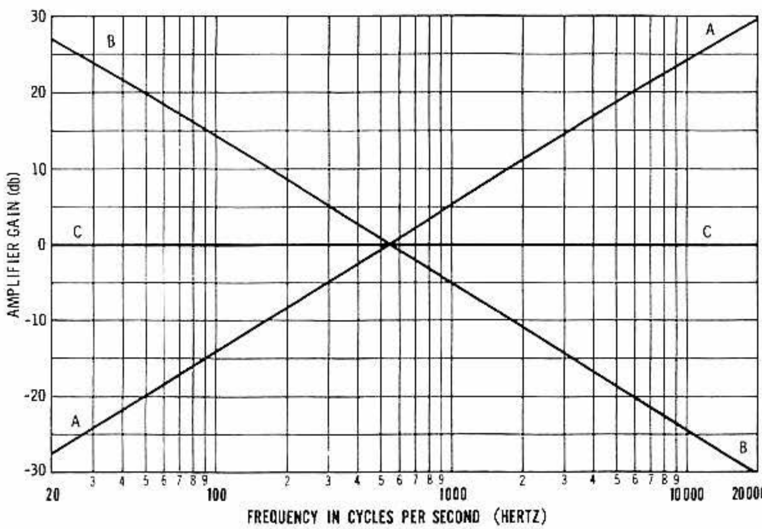


Fig. 3-2. Theoretical curves required to record and reproduce constant amplitudes.

and that voltages equal in amplitude at all frequencies were applied to a constant-velocity type of recording cartridge. The recorded groove swing would then vary inversely with the frequency as shown in Fig. 3-3. The amplitude in the record groove would be much greater at the low than at the high frequencies. Choosing two extremes of the audio band, the recorded amplitude at 30 Hz is 500 times that at 15 kHz.

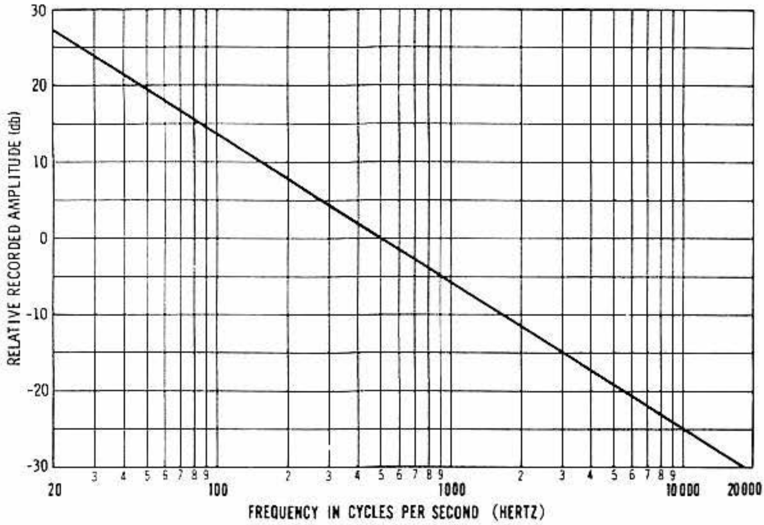


Fig. 3-3. Groove amplitude resulting from applying a constant voltage to a velocity-type cutter.

This condition is intolerable for two reasons. At the extreme low frequencies, the record groove would be exceptionally wide. At the high frequencies, the recorded amplitude would be too low to override the record surface noise. To overcome this, the RIAA (Recording Industry Association of America) adopted the recording curve shown in Fig. 3-4, which has become the standard of the industry. Assume that the record amplifier has been designed to produce a signal at the constant-velocity record head having the frequency response of the curve shown in Fig. 3-4. If a uniform signal from 20 Hz to 20 kHz is applied to the recording amplifier, the resulting groove cut on the record will have constant amplitudes up to 500 Hz. This limits the groove swings at the low frequencies. The constant-velocity mode is used from 500 Hz to 2120 Hz. In this range, the groove swings decrease with rising frequency. Above 2120 Hz, the constant-amplitude characteristics take over to allow the high frequencies to override record noise. Here, as below 500 Hz, a constant-amplitude

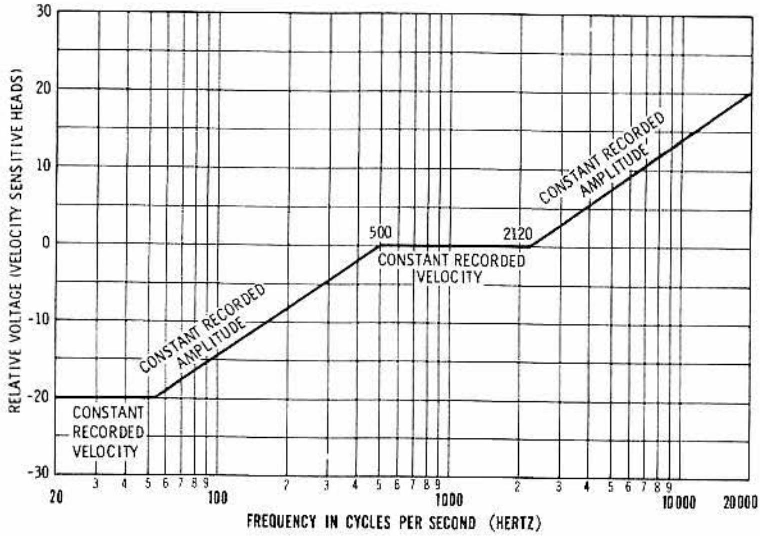


Fig. 3-4. Curve adopted by RIAA for recording. Practical curves have rounded corners.

groove is recorded for applied constant-amplitude voltages to the record head. The resultant excursions in the recorded groove are shown in Fig. 3-5.

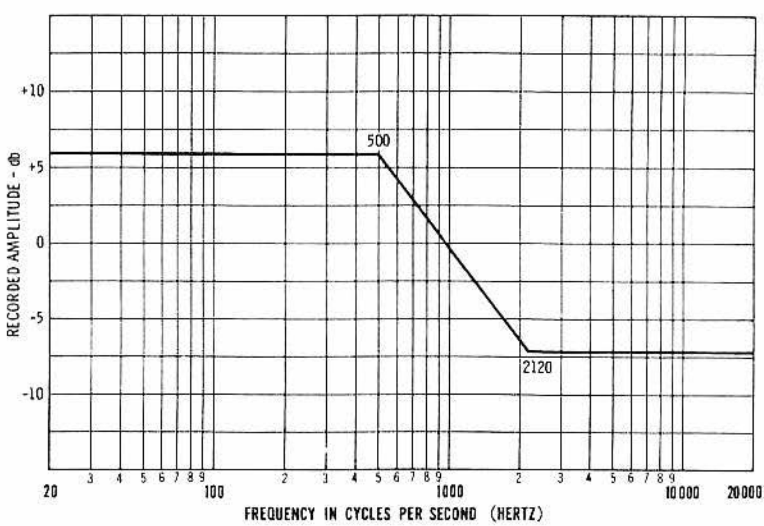


Fig. 3-5. Recorded amplitude resulting from constant-velocity cutter and RIAA recording curve.

Fig. 3-5 results from the following considerations. The recording-amplifier compensation curve required to produce uniform groove excursions at all frequencies, using a velocity-type cartridge, must have the rising characteristic of A-A in Fig. 3-2. This is the slope of the constant-amplitude sections in Fig. 3-4. A constant-velocity section is drawn between the two constant-amplitude sections. This requires no record-amplifier compensation as it follows the natural characteristics of the velocity-type record head, illustrated in Fig. 3-3. In this range,

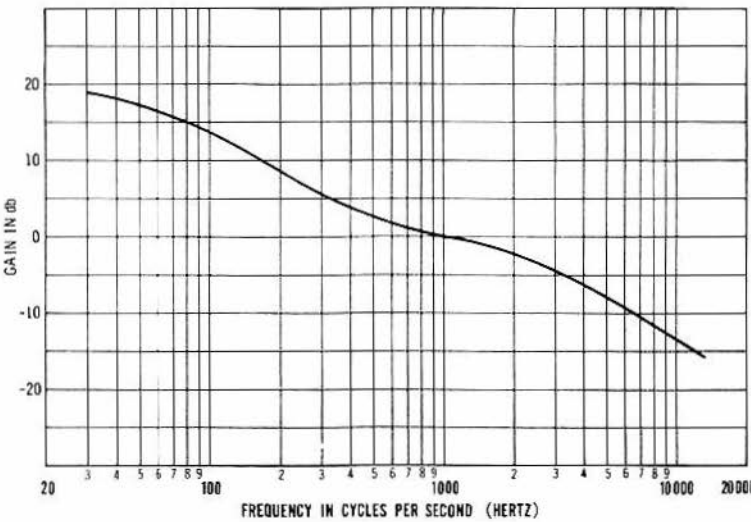


Fig. 3-6. RIAA playback curve.

recorded-groove excursions vary inversely with frequency. The net recorded amplitude of the curve in Fig. 3-4 thus becomes the curve in Fig. 3-5.

Reproducing records using a constant-velocity cartridge requires a playback amplifier with the frequency-response curve being the mirror image around the 0-db axis of the curve shown in Fig. 3-4. An exact curve, with the corners rounded off, is drawn in Fig. 3-6. Adding the curves in Figs. 3-4 and 3-6 (assuming the practical version of the curve in Fig. 3-4 has rounded corners) will result in a flat output along the 0-db axis. Obviously, to reproduce the input signal faithfully, there must be a uniform output from the playback preamplifier for a uniform input to the record amplifier.

The setup shown in Fig. 3-7 is used for measuring the response of the amplifier. A more complete test would include employing the particular cartridge to be used with the amplifier, as an integral part of the test circuit.

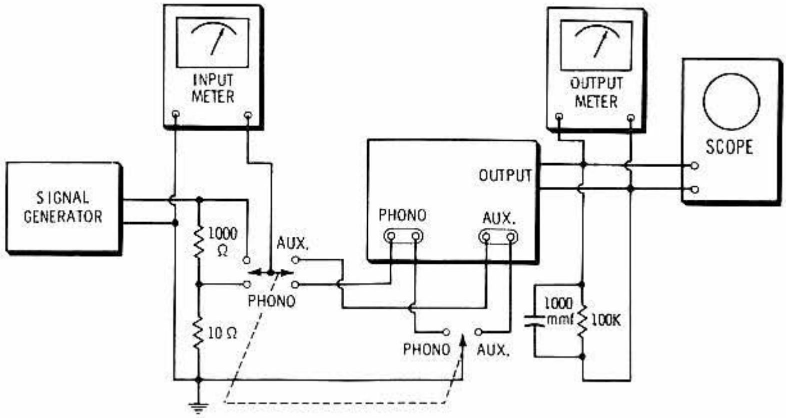


Fig. 3-7. Equipment setup for measuring the phono equalization curve.

CHECKING THE EQUALIZATION CURVE

In Fig. 3-7, a vtvm is shown monitoring the input to the amplifier. If the sensitivity of the vtvm is insufficient when the vtvm is connected as shown, it may be connected directly to the generator, before the divider network. The unit under test is assumed to be a preamplifier. Because of its relatively high output impedance, the connecting leads to the output meter and scope should be of the low-capacitance shielded type. The test instruments should present a low capacitance to the preamplifier. If an integrated preamplifier-power amplifier were under test, all that would be required is the use of a noninductive load resistor at the output, replacing the parallel combination of the 100K resistor and 1000-pf capacitor.

The *aux* input is asumed to encompass that portion of the preamplifier exhibiting a flat response (i.e., no equalization). Other inputs with this characteristic may be labeled *tuner* or *multiplex*. The *phono* channel performs two functions. First, it encompasses the amplification necessary for the low-voltage output from the constant-velocity cartridge. Next, it provides the equalization necessary for the particular cartridge.

In the test procedure, the output from the signal generator is first fed to the *aux* input. A considerably reduced signal, attenuated by the action of the 1000-ohm to 10-ohm voltage divider, is then fed to the *phono* input. If these resistive values load the audio generator excessively, the resistors in the dividing network may be increased in value proportionately. The desired signal level can be chosen using the slide switch shown in the drawing.

In most amplifiers, the *phono* and the *aux* grounds on the preamplifier section are independent to prevent hum-sensitive loops from being formed between these inputs. To avoid any upset of this condition and to maintain the independent ground returns built into the amplifier, the ground lead from the signal generator should be switched when the specific input and signal levels are chosen. This is accomplished by using the slide switch.

For the actual check, use the following procedure to avoid false measurements: Feed the signal to the *aux* input, and adjust all frequency-compensating controls for a flat output. Switch to 30 Hz and turn up the level control to the point before the preamplifier begins to distort; then, note this reading, in db, on the output meter.

Now, feed 1000 Hz to the *phono* input. Set the level control on the signal generator so that the meter at the output of the preamplifier indicates some convenient point 19 to 25 db below the reading taken above at 30 Hz. This is the reference, or 0-db, level. Measurements at all other frequencies are referred to this figure in terms of the number of db above or below this 0-db level.

Select several frequencies on the curve in Fig. 3-6. Set the generator to each of these frequencies. The output from the generator must be maintained at a constant level, as read on the input-monitoring meter. Note the indication in db above and below the 1000-Hz reference frequency. Compare these readings with the various points on the curve in Fig. 3-6. A good amplifier follows this curve with less than 2-db variation.

A more precise test includes the inductive effects of the cartridge to be used with the system. This factor becomes more critical when the cartridge is loaded by an impedance other than that prescribed by the cartridge manufacturer. When the preamplifier under test is transistorized, the inductive effect of the cartridge must be considered in the measurements.

Proper measurements will result only if the cartridge is placed in series with the *phono* output from the generator. Excessive high-frequency rolloff (possibly due to capacitance in the lead between the cartridge and the preamp) can then be adjusted reasonably well by increasing the value of the input resistor to the *phono* preamplifier. Low-frequency inaccuracies may require redesign of the equalizing network.

CONSTANT-AMPLITUDE CARTRIDGES

The output from a constant-amplitude cartridge is theoretically proportional to the recorded amplitude, as shown in Fig. 3-5. The playback curve must be the mirror image of this curve around the 0-db axis. An exact drawing of this curve, with corners rounded off

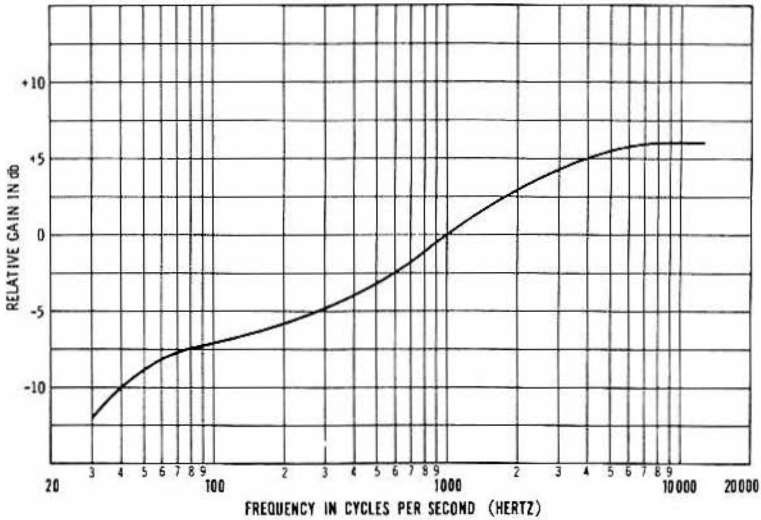


Fig. 3-8. RIAA curve for reproduction using a constant-amplitude cartridge.

to reflect more accurately the actual component characteristics, is drawn in Fig. 3-8.

The output from an amplitude-sensitive cartridge is usually reproduced through a linear amplifier. For accurate performance, the cut at the low frequencies and the boost at the high frequencies must be built into the cartridge. The high-frequency boost is achieved by resonances purposely designed into the unit. Low-frequency cut is a function of the loading at the output of the cartridge.

The equivalent circuit of this type of cartridge is a voltage source in series with a capacitance. The load resistor shown in Fig. 3-9 determines the loss at the low frequencies.

In this circuit;

$$e_{out} = e_{in}\left(\frac{R}{R + X_c}\right) \tag{3-1}$$

or,

$$\frac{e_{out}}{e_{in}} = \frac{R}{R + \frac{1}{j\omega C}} = \frac{j\omega RC}{j\omega RC + 1} \tag{3-2}$$

where,

e_{out} is voltage out,
e_{in} is voltage in,
R is equivalent resistance,
C is equivalent capacitance.

The result is the curve shown in Fig. 3-10. The numerator indicates that at 0 Hz the output is zero, and rises from there. The denominator indicates the frequency at which the output has dropped 3 db from the maximum flat output. The 3-db point is reached when, in the denominator, $j\omega RC = j$.

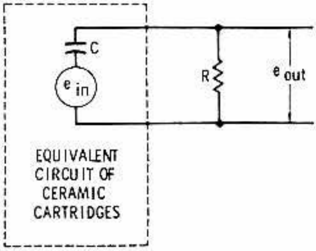


Fig. 3-9. Equivalent circuit of a ceramic cartridge feeding a resistive load.

The capacitance of the cartridge is usually assumed to be 500 pf. The frequency where the output has fallen 3 db is, from equation 3-2:

$$f = 1/2\,\pi\,RC \qquad (3\text{-}3)$$

At this frequency the $j\omega RC$ term in the denominator of equation 3-2 is equal to j.

Referring to the curve in Fig. 3-8, you will note that the response is down about 12 db at 30 Hz. Figuring back, it would be down 6 db

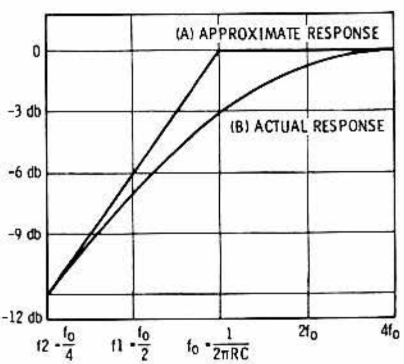


Fig. 3-10. Response curve of circuit in Fig. 3-9.

at twice this frequency, or 60 Hz; and would be at the breaking point, which is already 3 db down, at 120 Hz. Substituting this information into equation 3-3, we see that the required load resistor, R, is equal to:

$$R = \frac{1}{2\pi fC} = \frac{1}{6.28 \times 120 \times 5 \times 10^{-10}}$$

$$= 2.7 \text{ megohm} \qquad (3\text{-}4)$$

Unfortunately, few amplifiers have load impedances of that magnitude at the auxiliary inputs. A cathode follower is required at the input to provide this impedance. If the input impedance is 2.7 megohms, response at the *aux* input should be measured as described in Chapter 2. The output must be flat to enable the cartridge and preamplifier combination to provide the curve shown in Fig. 3-8.

It is frequently desirable to send the signal through the phono preamplifier when these cartridges are used. This serves the dual function of modifying the extreme requirements placed on R in equation 3-3 while permitting exact equalization for the different types of older records, when this is required and provided for in the amplifier. In the past, it was the practice to provide these different equalizations in the phonograph preamplifier for velocity-type cartridges only. These equalization curves are not shown here. The modern amplifier seldom includes these, for all record manufacturers have settled on the RIAA curve as standard. Older units may still incorporate the various equalizations.

A manufacturer of one of the popular types of amplitude-sensitive cartridges recommends that the circuit shown in Fig. 3-11 be used between the ceramic cartridge and the velocity-equalized phono preamplifier. This network is used to modify the curve originally used with velocity-type cartridges to a curve suitable for use with ceramic amplitude-type cartridges.

An analysis of the circuit in Fig. 3-11 can become quite involved, but is nonetheless worth-while. The general sinusoidal solution can be applied to most constant-amplitude cartridges. The mathematics is carried out in Appendix D.

The response curve of this network, using a ceramic cartridge, is shown in Fig. 3-12A. The RIAA playback curve, originally drawn in Fig. 3-6, is repeated in Fig. 3-12A with the corners squared off. This is an integral characteristic of the phonography preamplifier in question. The sum of the two curves is the resultant curve necessary to produce the output from a ceramic cartridge properly.

Two things should be noted. First, at these frequencies the built-in rolloff characteristic in the preamplifier just about cancels the high-frequency boost from the network in Fig. 3-11. The resultant re-

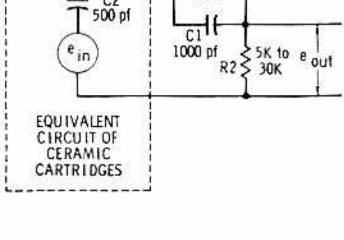
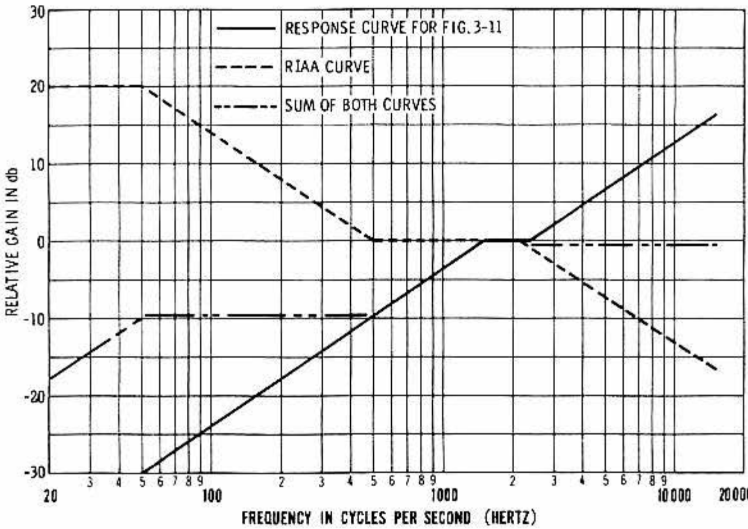
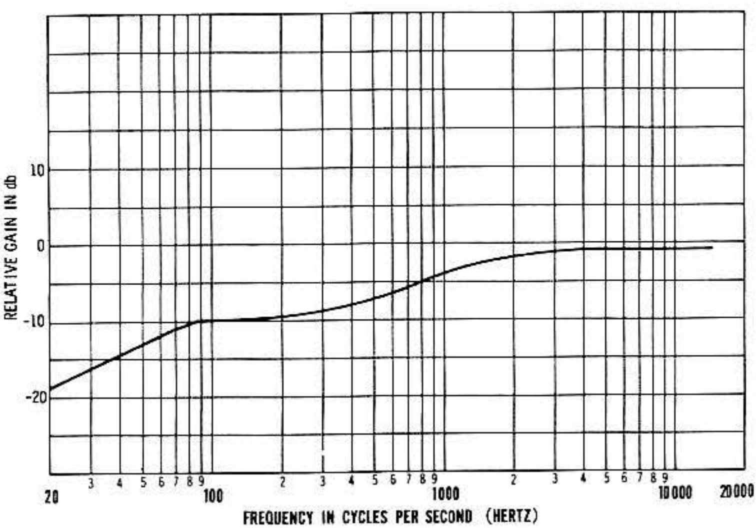


Fig. 3-11. Network used between ceramic cartridge and phono preamp with constant-velocity equalization.

sponse at these frequencies is flat. At the lower frequencies, the curve conforms with that required to reproduce the output from a ceramic cartridge properly at the low end of the band. The flat response at



(A) Produced with a ceramic cartridge.



(B) Resulting curve from circuit in Fig. 3-11 in conjunction with a preamp equalized for constant-velocity RIAA characteristic.

Fig. 3-12. Recording curves.

high frequencies, as shown in Fig. 3-12, does not mean that the boost shown in Fig. 3-8 is disregarded. The rise in response is still present because of the built-in resonances in the cartridge, as discussed previously. R_2 should be chosen so that a ceramic cartridge, when played through a velocity-equalized preamplifier, will provide sufficient output without overloading the preamplifier stages.

Checking the preamplifier response, when this network is used, is straightforward. First, check the preamplifier stage as previously described. Next, insert the network in Fig. 3-11 between the oscillator and the preamplifier. A 500-pf capacitor, representing the equivalent capacitance of the cartridge, should be inserted in series with the lead from the signal generator. The resultant output should follow the combined curve in Fig. 3-12, but with the corners rounded off as shown in Fig. 3-12B.

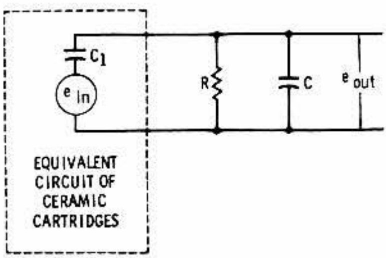


Fig. 3-13. Circuit used with a ceramic cartridge when feeding a relatively low-impedance amplifier input.

CERAMIC CARTRIDGES FEEDING MEDIUM IMPEDANCES

Any arrangement which results in the curve shown in Fig. 3-12B would result in proper reproduction from a disc recorded in accordance with the RIAA characteristic and played with a constant-amplitude cartridge. Compensation is required only at the lower frequencies. The required high-frequency boost is controlled by resonances built into the cartridge.

Let us assume that you wish to use the high-level *aux* input for the ceramic cartridge, and that the input impedance at the amplifier is less than the 2.7 megohms recommended for R in Fig. 3-9 and determined by equation 3-4. In this instance, the circuit in Fig. 3-13 is used. Here, R is the resistance at the *aux* input of the amplifier.

As you may recall, this section of the amplifier provides an output with a uniform frequency response when a constant voltage is applied at the input. Also, C_1 is the capacitance in the equivalent circuit of the cartridge. Capacitor C must be chosen to shunt the resistive input impedance of amplifier R. The equation describing this network is:

$$\frac{e_{out}}{e_{in}} = \frac{R/(1 + j\omega RC)}{R/(1 + j\omega RC) + 1/j\omega C_1} \qquad (3\text{-}5)$$

where, $R/(1 + j\omega RC)$ is the impedance of the parallel r-c combination and $1/j\omega C_1$ is the impedance of C_1. Simplifying equation 3-5 gives:

$$\frac{e_{out}}{e_{in}} = \frac{j\omega RC_1}{j\,(\omega RC_1 + \omega RC)} \qquad (3\text{-}6)$$

The frequency when the output has rolled off 3 db is:

$$j\,(\omega RC_1 + \omega RC) = j$$

$$\text{or } \omega = 1/(RC_1 + RC)$$

$$\text{and } f = \frac{1}{2\pi R(C_1 + C)} \qquad (3\text{-}7)$$

The $j\omega RC_1$ term in the numerator indicates that the output is zero at d-c with a steady rise of 6 db per octave. The curve fitting equation 3-5 is shown as Fig. 3-14A. The output drops approximately 6 db per octave beginning at the frequency of $f = 1/2\pi R(C_1 + C)$, and is equal to zero when $f = 0$. It should be noted that (A) of Fig. 3-14 is an approximation of the actual curve. The output is actually 3 db down from its maximum when $f_0 = 1/2\pi R(C_1 + C)$; it is down 7 db when $f_1 = f_0/2$; it is down 12 db at $f_2 = f_1/2$; and it drops 6 db per octave from there on, as shown in (B) of Fig. 3-14. If the curve with the 6-db-per-octave slope were extended back to the 0-db reference axis, it would intersect it at f_0. Thus, curve (A) of Fig. 3-14 is used as an approximation for this curve, plotting f_0 on the 0-db axis rather than 3 db below it. This approximation has been, and will continue to be, applied through the book for convenience and ease of illustration.

As an example in the use of equation 3-5, assume R to equal 5×10^5 and C_1 to equal 5×10^{-10}. A good approximation to the re-

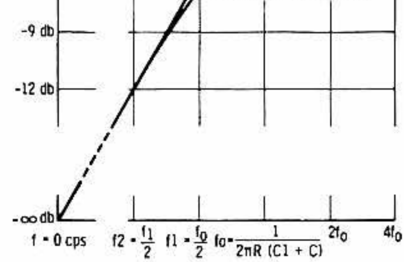


Fig. 3-14. Response curve of circuit in Fig. 3-13.

quired curve in Fig. 3-8 can be found if we let f equal 120 Hz. (This was discussed above in referring to Fig. 3-9 and equation 3-4.) Substituting these constants into equation 3-7 would make C approximately equal to 2200 pf. Whatever the values of C_1 and R may be, the capacitor required to shunt R for a reasonably flat response can be readily calculated from equation 3-7. It should be remembered that these results fit the curve to a fair approximation only. In all practical situations, reproduction will be satisfactory.

The capacitor can be built into the amplifier, shunting the input resistor, or it can be an integral part of the cartridge. If the frequency response of the amplifier is checked, a capacitor across the input should not alter the response of the amplifier. This is especially true if the output impedance from the signal generator used in making the test is low. Should a high-impedance signal source be used, some rolloff at the high frequencies may be noted. This rolloff will be a function of the impedance of the generator.

CHAPTER 4

Measuring and Matching Tape Playback and Microphone Preamplifiers

The frequency-response characteristics of the tape playback head are similar to those of the constant-velocity phonograph cartridge. If a tape were magnetized equally at all frequencies and played on a tape machine, the output from the playback head would rise linearly with frequency at the rate of 6 db per octave. This rise would continue to the frequency limits of the head if several factors did not intervene to upset this convenient state of affairs.

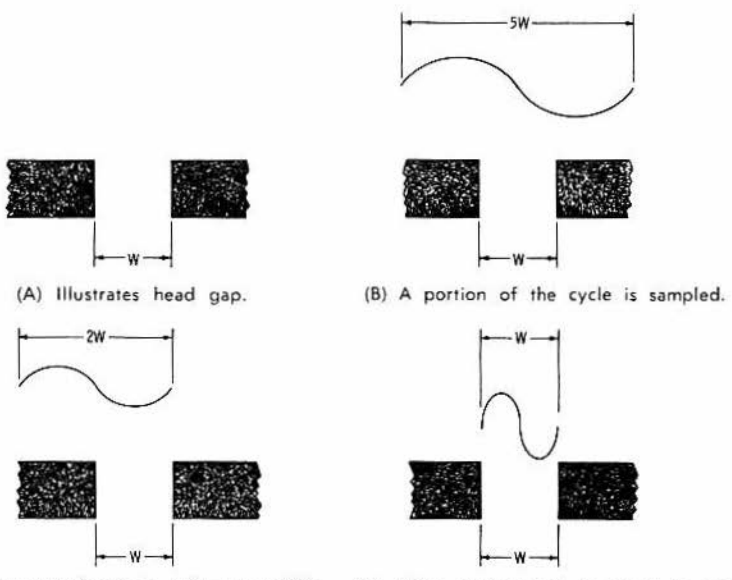
In the recording process, the degree of tape magnetization theoretically is proportional to the current passing through the recording head. At high frequencies this characteristic is altered because of the proximity of the tiny magnets which make up the recorded signal on the tape. The opposite poles adjacent to each other "cancel" some of the magnets. Another loss is due to insufficient depth in the recording layer. The frequencies at the treble end of the spectrum are recorded at the surface of the magnetic material, while the lower frequencies penetrate the recording layer to a greater degree.

It can also be demonstrated that the recording bias current has a definite relationship to the recorded signal appearing on the tape. Excessive bias current tends to decrease the high frequencies more than it does the low end of the band. On the other hand, insufficient bias will cause excessive distortion. A compromise between the two extremes must be chosen to make a good recording.

Deviation from a standard relationship (for the particular head) between recorded flux and audio-signal power fed to the record head can be related to power losses in the head. This is particularly true at high frequencies. The standard ratio assumes that equal magnetic fields are set up for equal head currents at all frequencies. This is not the case. There is more power dissipated in the head at high frequencies than at low frequencies. As a result, less high-frequency power remains to magnetize the tape.

Although high-frequency losses are of primary significance in the recording process, losses at both extremes of the audio band are important in playback. If the opposing magnetic poles which result when a recording is made on tape lie near the playback head gap, the output will be high. At very low frequencies, because of the large wavelengths the poles may be far from the head gap and the output will tend to decrease.

The width of the gap in the playback head is significant for high-frequency response. This can be explained with the help of Fig. 4-1. Assume that tape passes over a gap of the width at w. If the recorded wavelength is greater than w, let us say 5w, the gap sees only a portion of the recorded waveshape at any one instant of time as the tape moves over the head. Each portion sets up a different current through the head coil until the entire sine wave is traced. As the wavelength is decreased, a greater portion of the cycle appears across the gap at any one instant. When the wavelength recorded on the tape is twice the gap width, the current through the head represents the average



(A) Illustrates head gap.

(B) A portion of the cycle is sampled.

(C) When wavelength is twice gap width, half of the cycle is sampled at any one moment.

(D) When wavelength is equal to gap width, output is zero.

Fig. 4-1. Effect of gap width on current through playback head.

current of a half cycle or I_{peak}/π (The derivation of this appears in Appendix E). When the gap width is equal to a complete cycle, the average output current in the playhead is zero. The output increases again, with frequency, after the null. At the null frequency, the gap width equals the recorded wavelength.

This leads to two commonly assumed factors. First, the usable minimum wavelength is double the gap width of the playback head, for only then is there sufficient compensatable output from the head. Second, for the same gap width, there is less usable upper frequency at slow speeds than at high speeds, because a specific recorded wavelength on the tape is reached at a lower frequency when recording at slow speeds than at high speeds.

Finally, the usable bandwidth can be calculated from the gap width. The null in output is at the frequency when the recorded wavelength is equal to the gap width. The recorded wavelength can be derived from the equation

$$\lambda = vt = \frac{v}{f} \qquad (4\text{-}1)$$

where,

λ (lambda) is the wavelength,
v is the velocity of the tape,
t is the period or the time duration of one cycle,
f is the frequency in cycles per second, or Hertz.

The null frequency is then:

$$f = \frac{v}{\lambda} \qquad (4\text{-}2)$$

As an example, use a velocity of 3.75 ips and a wave-length of 0.0001 inch per cycle. The wavelength is numerically equal to the gap width. Substituting these into equation 2 yields:

$$f = \frac{3.75 \text{ inches/second}}{0.0001 \text{ inches/cycle}} = 37{,}500 \frac{\text{cycles}}{\text{second}} = 37{,}500 \text{ Hz}$$

The usable upper limit of the reproducible band is 37,500/2, or about 18,000 Hz. This limit is seldom realized because the effective gap is usually greater than the mechanical gap. The imperfections at the edge of the gap across which the tape rides are a major factor in producing this discrepancy. A typical playback curve resulting from a constant-current recording is shown in Fig. 4-2. The rolloff frequency is a function of the gap width, tape speed, and tape-resolution characteristics.

To produce a flat playback response from a constant-current recording, the playback preamplifiers are equalized in accordance with the curves shown in Fig. 4-3, accepted as a standard by the industry.

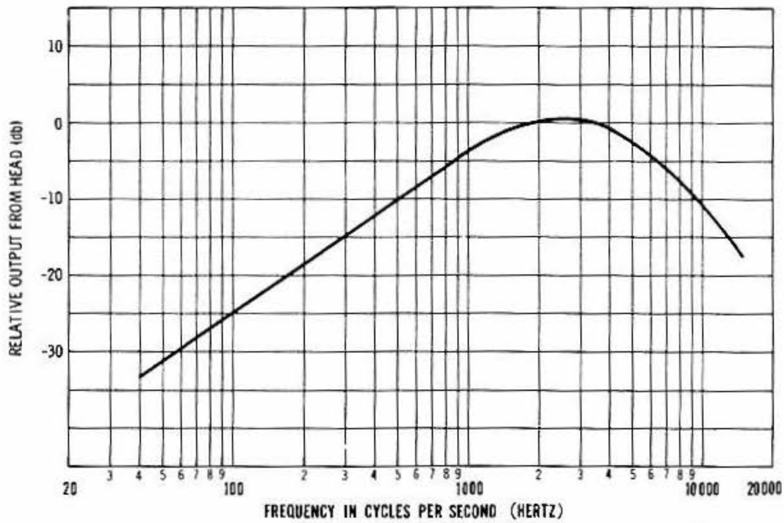


Fig. 4-2. Output from playback head reproducing a constant-current recording at 7½ ips. Gap width is 0.0002 inch, although effective gap is greater.

In order to determine the resultant frequency response due to the preamplifier and the playback-head combination, the curve in Fig. 4-2 (head characteristics) must be added to one of the equalization

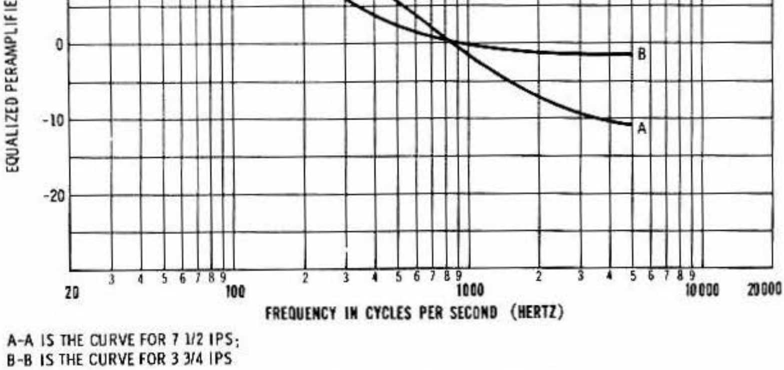
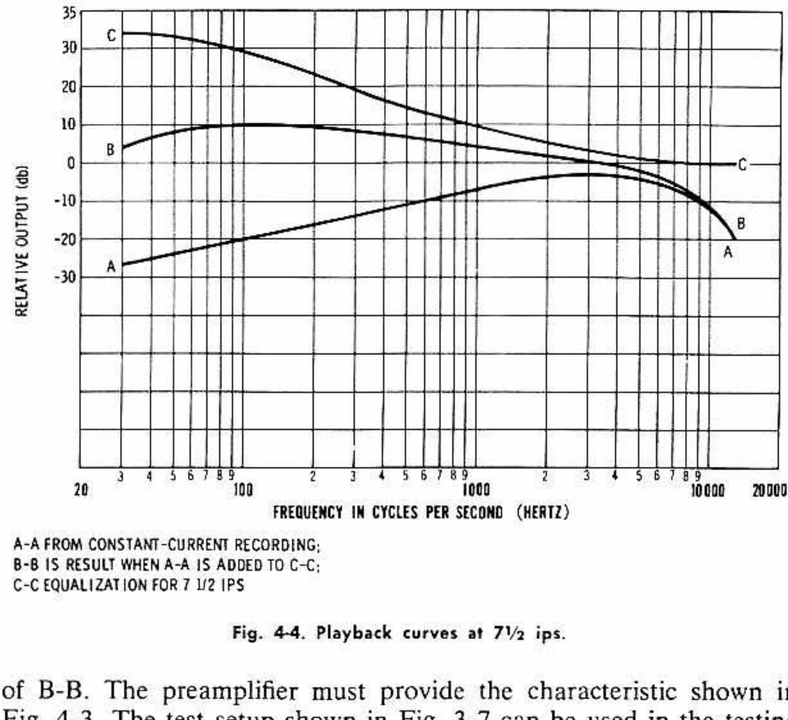


Fig. 4-3. Standard preamplifier equalization curve.

curves in Fig. 4-3. An example of this procedure at the 7½-ips speed is shown in Fig. 4-4.

Fig. 4-4 indicates that there is insufficient boost at the high and low frequencies of the audible band. This is compensated for in the recording process and thus is of no interest here. Record preamplifiers are equalized to a response that is the approximate mirror image



A-A FROM CONSTANT-CURRENT RECORDING;
B-B IS RESULT WHEN A-A IS ADDED TO C-C;
C-C EQUALIZATION FOR 7 1/2 IPS

Fig. 4-4. Playback curves at 7½ ips.

of B-B. The preamplifier must provide the characteristic shown in Fig. 4-3. The test setup shown in Fig. 3-7 can be used in the testing procedure. For a discussion of the test circuit and the correct test procedure, see Chapter 3.

MICROPHONE CHARACTERISTICS AND REPRODUCTION

Piezoelectric microphones used in high-fidelity applications usually fall into two groups. The first group utilizes crystal or ceramic elements which operate on the piezoelectric principle. They are designed so that the resonant frequency is above the audible range. The output voltage will then be substantially constant for a uniformly applied signal at all frequencies. Equalizing networks similar to the type used with constant-amplitude phonograph cartridges can be used to flatten the response as required over the audio spectrum.

The equivalent circuit of the crystal and ceramic microphone is a voltage source in series with a capacitor, which should work into an essentially resistive load at the microphone preamplifier. All this is shown in Fig. 4-5.

The equation for the circuit in Fig. 4-5 is:

$$\frac{e_{out}}{e_{in}} = \frac{R}{R + \frac{1}{j\omega C}} = \frac{j\omega RC}{j\omega RC + 1} \qquad (4\text{-}3)$$

The numerator in the equation indicates that the output for a zero frequency (d-c) input is zero. The 3-db rolloff point is determined by the denominator when:

$$j\omega RC = j \text{ or } f_1 = \frac{1}{2\pi RC} \qquad (4\text{-}4)$$

The curve described by this equation is shown in Fig. 4-6.

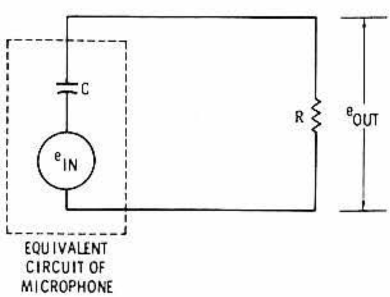


Fig. 4-5. Equivalent circuit of a ceramic microphone working into a resistive load.

At the frequency f_1, the output from the microphone is down 3 db. At $2f_1$, the output is down 1 db, while at $4f_1$ the output may be considered as not having dropped off at all. The proper procedure for deriving R in equation 4-4 is first to select the lowest frequency which must be reproduced without any rolloff, consistent with the quality of the microphone used. This frequency is $4f_1$. The frequency f_1 is one-quarter of the lowest frequency requiring a nominally flat amplification. Substitute this value for f_1 as well as the capacitance of microphone C into equation 4-4, and solve for resistor R. (The capacitance of a crystal microphone varies from 500 to 15,000 pf.) The exact value of the capacitance should be obtained from the manufacturer of the particular microphone in question. If the actual value cannot be obtained from the manufacturer, assume it to be 500 pf.

In the test procedure, use the circuit shown in Fig. 3-7. Insert a capacitor, equal in value to the equivalent microphone capacitance, in series with the lead connecting the signal generator to the preamplifier.

The second popular type of microphone is the moving-coil, or dynamic, variety. Here, a diaphragm is set in motion by the air pressure emanating from an audible signal. A coil attached to this diaphragm moves in a strong magnetic field, causing current to be generated in its winding. The output from that microphone normally would rise with frequency for a constant signal pressure. Careful and proper choice of the mechanical resonant frequency can provide an output uniform with frequency over most of the audible spectrum.

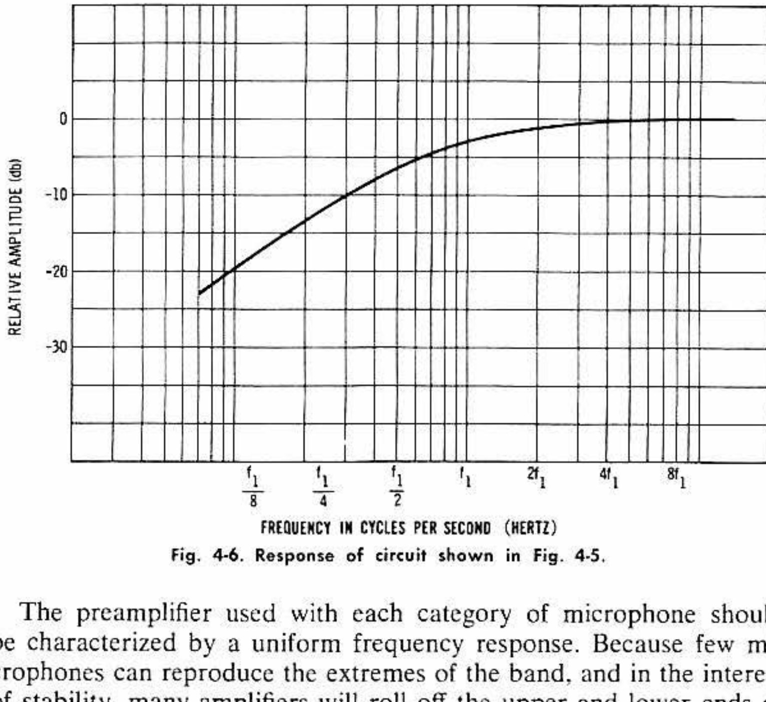


Fig. 4-6. Response of circuit shown in Fig. 4-5.

The preamplifier used with each category of microphone should be characterized by a uniform frequency response. Because few microphones can reproduce the extremes of the band, and in the interest of stability, many amplifiers will roll off the upper and lower ends of the spectrum. It is not uncommon to find that 50 Hz and 10 kHz in one amplifier, or 40 Hz and 15 kHz in a second amplifier, are 3 db down from the maximum output level.

The procedure for testing microphone preamplifiers varies only slightly from that performed for tape-head or phonograph-cartridge units. In the *aux* position, the amplifier is adjusted to perform at its most linear frequency capabilities. The switch is then thrown to the lower-output *phono* position. Adjust the level control on the signal generator so that the amplifier under test will deliver about 5 db below the maximum undistorted output at 1000 Hz. Proceed to check the relative gain at all other frequencies, using this point as the reference level.

PRECISE TESTING

Just as in the case of the constant-velocity phonograph cartridge, the tape-head and dynamic microphone used should be included as an integral component in the test setup. This would take into account the effect of the inductance of these components as well as the effect of the input impedance of the preamplifier. However, several precautions must be taken to make valid readings. Refer to Fig. 3-7.

1. The resistor connected to ground in the divider network at the signal generator must be small in value. It must be negligible compared to the impedance of the transducers under test.
2. The resistors in the divider network must be noninductive.
3. The leads from the generator should be of the low-capacitance shielded type.
4. The transducer must be connected in series with the "hot" lead from the generator.
5. The transducer must be shielded so that it does not pick up stray electrical or acoustical signals.
6. The signals fed to the preamplifier through the cartridge, tape head, or microphone must be on the order of magnitude of the signal normally expected out of the particular transducer involved.

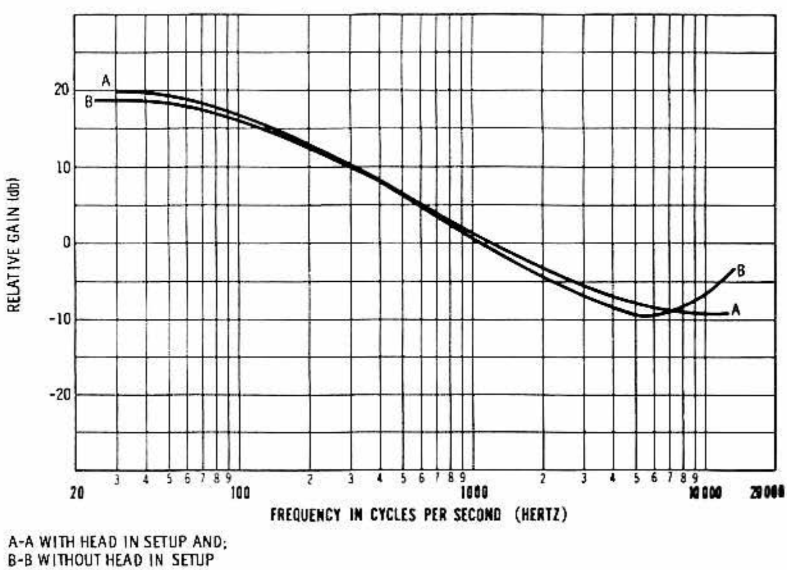


Fig. 4-7. Response curve illustrating the effect of stray capacitance in test equipment and shielded cables.

Isolation between the signal source and the transducer may be difficult to accomplish. Phono cartridges and tape heads should be placed in magnetically shielded boxes, and microphones should be housed in acoustically as well as magnetically shielded containers. Their leads should be shielded from stray-field pickup.

One further precaution: the divider network should be at the generator. Use three feet of shielded lead to connect the generator to the transducer. Do not form a coil with the shielded lead. Stretch it out over the workbench. This will maintain the signal in the shielded lead at a low level, minimizing the probability of stray fields being set up and induced into the transducer. Placing the transducer three feet away from the signal generator minimizes direct induction from the generator into the transducer.

The importance of the transducer in the test circuit cannot be overemphasized. In Fig. 4-7, note the different curves resulting from the tests made on a transistorized playback preamplifier used in a commercial tape deck. Note the difference in the curves with and without the playback head in the test setup. Curve A-A adheres closely to the standard when the head is used in the more accurate method of testing just discussed.

CHAPTER 5

Checking Frequency-Compensating Circuits

High-fidelity amplifiers are usually equipped with ample frequency-discriminating functions. These include some type of variable bass- or treble-compensating network as well as loudness controls, scratch, and rumble filters. Each designer of audio equipment has his own prejudices as to the relative need for each of these functions. Some engineers feel that any deviation from a perfectly flat frequency response is a deviation from "true" high fidelity. Others have made a life study of frequency-compensation requirements and have designed complex networks to suit their research conclusions. In this chapter, four groups of these functions will be considered: the tone control, scratch filter, rumble filter, and loudness control.

THE TONE CONTROL

Every high-fidelity amplifier has at least one tone control, although most have two. The primary function is to compensate somewhat for room acoustics. Many audiophiles use these controls to satisfy various secondary functions—to compensate for a poor phonograph cartridge, for example, or to get a booming bass, produce a piercing "high-fidelity" treble, reduce turntable rumble, record scratch, or tape hiss, and so on.

A complete set of tone controls should provide, in varying degrees, five groups of frequency characteristics. The four most obvious are bass and treble, boost and cut. The fifth, and most important, is a flat frequency response. While most controls will give adequate boost and cut, many will not have one definite position where the frequency response is held to within +1 db over the entire audible spectrum.

The tone controls can be tested with the setup in Fig. 3-7. This circuit is also used to measure the equalization of the preamplifier sections. A few words on application should be noted here to avoid fallacious readings. All level controls on the amplifier under test must be set at maximum, to prevent the frequency-response curve from being influenced by the effect of any stray capacitances and by the setting of these controls.

Next, the output from the signal generator must be low enough so that the amplifier will not be overloaded at any test frequency. As an example, an amplifier providing 5 volts at 20 Hz should never be required to deliver more than this at any setting of the tone control. If 20 db of boost is expected at 20 Hz in the maximum bass-boost position, set the output from the generator so that the amplifier can provide this boost without distortion. If the maximum undistorted output voltage from the amplifier is 5 volts at 20 Hz, the reference level should be at most 20 db below this point, or 0.5 volt.

Finally, monitor the output from the signal generator, using an a-c voltmeter. Monitoring is important because the output from the generator may not be constant over the entire frequency range. The output from the amplifier should also be observed on a scope, to be certain that the output meter is indicating a sinusoidal voltage and not hum or a nonsinusoidal signal.

If the unit under test is a preamplifier, use the circuit shown in Fig. 3-7. If the unit is an integrated preamplifier-power amplifier combination and the output is to be measured after the power amplifier, replace the 1000-pf capacitor and 100kΩ resistor with the proper matching load resistor.

Once these precautions are borne in mind, the various functions of the tone controls can now be checked. Determine the maximum essentially sinusoidal output from the amplifier at 20 Hz and 20 kHz. Choose a convenient voltage (or level on the db scale) 20 to 25 db below the lower of these two readings. This will be the reference level. Adjust the tone controls for the most linear output over the entire audio range at this reference level, and draw this curve.

Note the amplifier settings for flat frequency response. Now set the bass control for maximum boost. Readjust the signal generator so that the output meter will once again read the reference level at 1000 Hz. Draw the frequency-response curve, using this 1000-Hz level as the 0-db reference point. The curve should be checked from 20 Hz to 20 kHz, to determine the effect of the control on all portions of the audio range.

Take readings from different settings of the bass control, midway between the flat response and maximum bass boost. Do the same for the maximum bass attenuation setting, as well as at some point midway between the flat response and maximum cut. Repeat all four measurements for the various settings of the treble control. Each time, check the response from 20 Hz to 20 kHz relative to the same 1000-Hz, 0-db reference level.

Two of the many possible sets of curves resulting from these tests are shown in Fig. 5-1. The curves shown in Fig. 5-1A are usually provided by a losser-type tone-control circuit, while those in Fig. 5-1B are characteristic of the Baxendall-type feedback circuit.

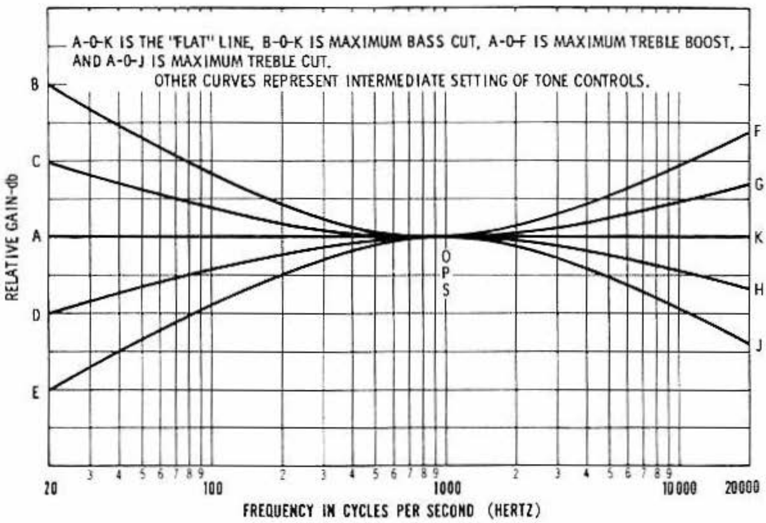
In the Baxendall circuit, the curves for the extreme setting of the controls pivot around some midfrequency, such as 1000 Hz ("O" on the curve). At intermediate settings of the bass control, the curves start to rise and fall at some frequency lower than 1000 Hz ("P" on the curve), while the upper and middle frequencies are not affected. Similarly, the curves for intermediate settings of the treble control start to rise and fall above 1000 Hz ("S" on the curve). Only the extremes of the audio band are affected by this type of tone-control network.

This is not the case with controls using the losser type of network. Here, the curves pivot around a midfrequency for all settings of the controls. Effectively, points "O," "P," and "S" merge to one point at about 1000 Hz. The entire upper half of the band is affected when the treble control is adjusted, and the entire lower half is affected when the bass control is adjusted.

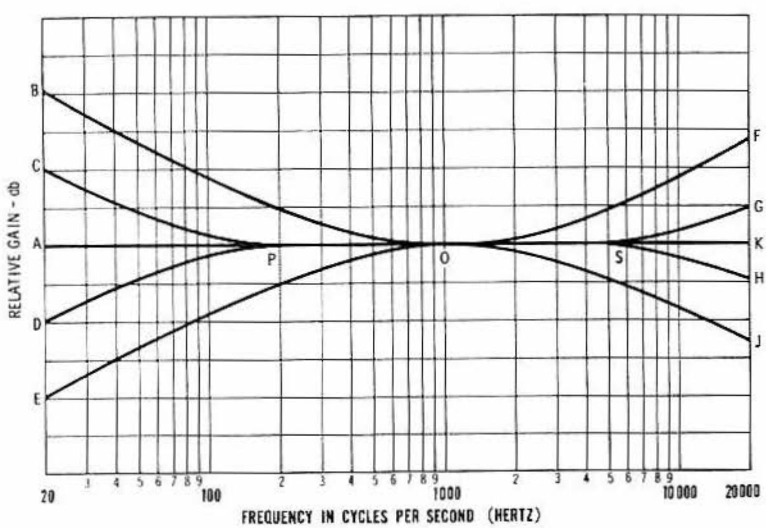
The Baxendall arrangement is most desirable in high-fidelity applications because it is more selective than the losser arrangement. In amplifiers where the frequency range is limited, the losser type of circuit is more desirable. In such amplifiers, extremes of the audio range do not exist. The Baxendall circuit produces no noticeable effect for most of the range, while the frequency variations caused by the losser circuit are usually quite obvious.

The curve for flat response, "A O K," has been assumed perfectly flat. This is, of course, an ideal situation seldom obtained, especially with circuits of the losser type. It has also been assumed that each control affects only a portion of the frequency range. Theoretically, the bass control should affect only the low frequencies, while the treble control should affect only the highs. Actually, this is only partially true. The suggested test for each setting covered the entire audible spectrum. This was done to check the effect of the controls over the entire audio range, to determine the effects on the sections they should not control as well as the amount of control they do exert over the assigned portion of the spectrum. The amount the bass control affects the treble and the amount the treble control affects the bass are measures of quality. Naturally, the least effect is the most desirable.

Tone controls should not affect the average volume level. The 1000-Hz reference level should remain unchanged, whatever the set-



(A) A losser circuit.



(B) A Baxendall circuit.

Fig. 5-1. Typical tone-control curves.

ting of the controls may be. Any change from the predetermined reference level at 1000 Hz is an indication of the quality of the particular circuit.

As a final thought, no distortion should be produced at any setting of the tone controls. The Baxendall type of tone control is reasonably distortion-free. In a well-designed circuit, the losser type can also provide clean output. Some "economy" arrangements, such as placing the tone controls in a feedback loop around the output transformer, give satisfactory performance at flat response settings, but result in entirely too much distortion in extreme boost positions. This can readily be verified in amplifiers using this circuit.

THE SCRATCH FILTER

Any filter placed in a circuit limits the full frequency range. Scratch filters are no different. A good filter is designed to minimize this factor. Scratch noise from phonograph records and hiss from tape produce audible components, primarily in the upper-frequency range. The convenience of this coincidence allows one filter to be designed to eliminate or minimize both types of interference by simply limiting some of the higher frequencies. A good filter will minimize only the highest frequencies. A poor filter will also attenuate the midfrequencies.

The exact frequency at which the attenuation should be maximum to best eliminate noise is a variable often determined by the whim of the particular design engineer. For good noise suppression, this frequency must be down far enough to encompass the lowest noise frequencies, and it must be high enough not to eliminate any more of the high frequencies than are absolutely necessary. In this discussion (for convenience and reasonable logic), it will be assumed that it is desirable to be 6 db down at 6300 Hz and have at least a 6-db-per-octave rolloff above that.

Although it may be analytically difficult to calculate R and C from the 6-db point, it can be easily determined when the 3-db attenuation frequency is considered. From the curve in Fig. 5-2, the frequency at which the output has dropped 3 db is 3500 Hz. Thus, the product of R and C required to produce this curve for the network in Fig. 5-2 is:

$$RC = \frac{1}{2\pi f} = \frac{1}{2\pi(3500)} \qquad (5\text{-}1)$$

All of R and a portion of C are usually integral parts of the circuit (inherent tube capacitance and circuit resistance) in which the filter is to be included. The amount of C that must be added to the existing circuit to provide the 6-db cut at the predetermined frequency must be determined experimentally.

A close look at the curve in Fig. 5-2 shows two major disadvantages of the filter. First, the high-frequency rolloff starts at 1000 Hz—a fre-

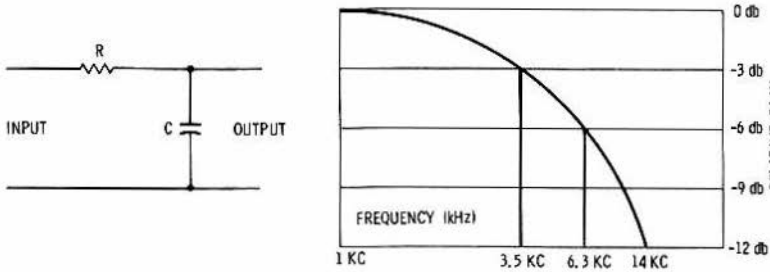
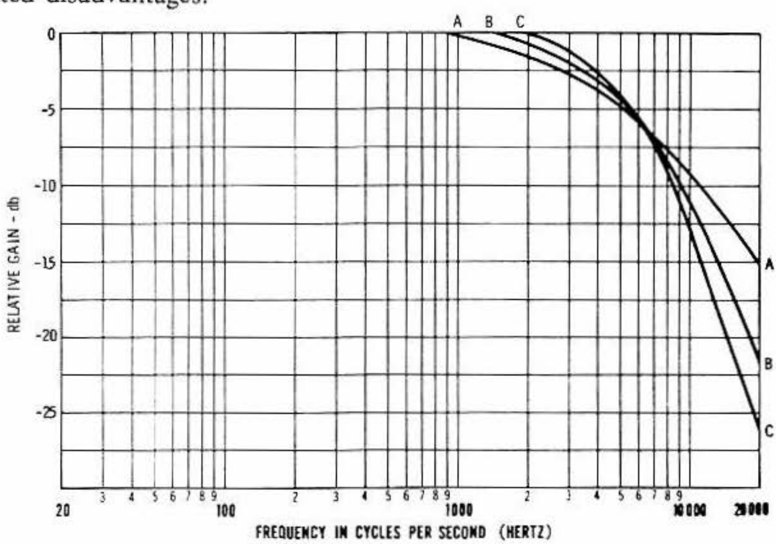


Fig. 5-2. A schematic and response curve of a typical scratch-filter circuit.

quency which we do not want affected to any noticeable degree. Second, at just about the highest frequency that must be attenuated by the filter, 14 kHz, the gain is only down about 12 db. This rolloff is frequently inadequate. The information is redrawn as curve A in Fig. 5-3.

More networks of the type in Fig. 5-2A, with identical time constants R C, can be added into the path of the signal to further attenuate the high frequencies. Curve C in Fig. 5-4 shows the result of three such networks in one amplifier. The rolloff starts at 2000 Hz, an improvement of one octave over curve A, and the gain is down 20 db at 15 kHz, an improvement of 7.5 db. Although adequate, the three tandem networks add attenuation into the circuit, with all the associated disadvantages.



(A) One filter. (B) Two filters. (C) Three filters.

Fig. 5-3. The effect on an amplifier using the scratch filter shown in Fig. 5-2. All curves are 6 db down at 6300 hertz.

A good compromise uses two such networks. Comparison of curves B and C in Fig. 5-4 indicates just how minor this compromise is. Scratch filters (and the rumble filters discussed later) have frequently been included in feedback networks to provide an extremely sharp rolloff. Stereo amplifiers seldom include these networks in feedback loops because of their high cost as well as dubious value.

The equipment in Fig. 3-7 may be used to test the filter network. First, adjust all controls on the amplifier for maximum flat response. Next, switch the filter into the circuit. Set the output from the signal generator so that the amplifier will reproduce 1000 Hz at a level sev-

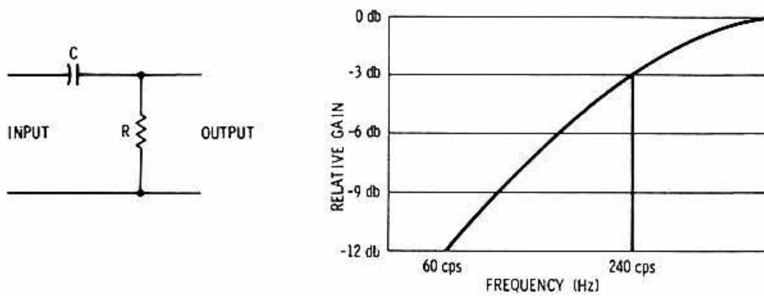


Fig. 5-4. A rumble-filter circuit and typical response curve.

eral db below its maximum output capabilities. Check the response from 20 Hz to 20 kHz to determine how the rolloff is affected by the filter, at what frequency the attenuation is 6 db, and how the filter reshapes that portion of the audio spectrum where it actually should have no effect.

THE RUMBLE FILTER

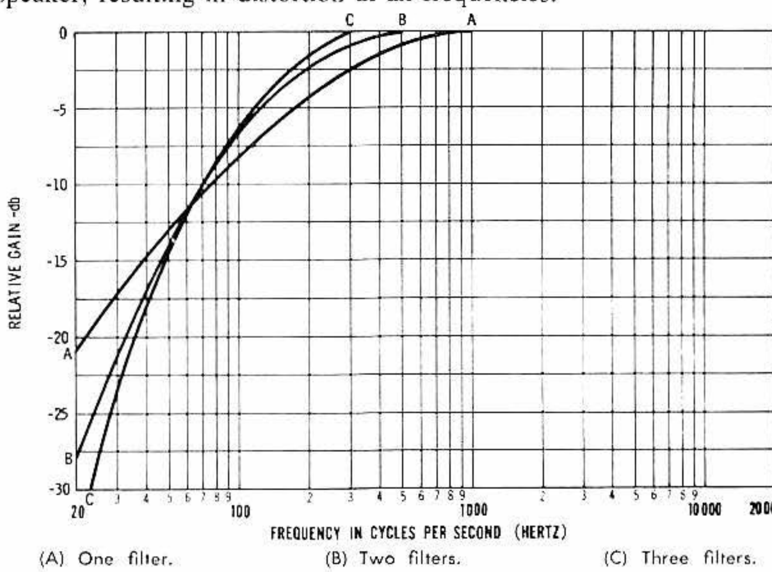
The rumble-filter network attenuates the low frequencies caused by rumble in the turntable. For a turntable with a two-pole motor, attenuation must be relatively high beginning at 60 Hz. If the turntable uses a four-pole motor, attenuation should be concentrated from 30 Hz down. In both cases, attenuation should be at least 10 db at the frequency indicated, and more at lower frequencies. In this discussion, we will consider 60 Hz as the objectionable frequency, with 12 db assumed to be the required attenuation at this frequency.

The rudimentary rumble filter is shown in Fig. 5-4, along with the response curve. The analysis is similar to the one for the scratch filter: For an attenuation of 12 db at 60 Hz, 240 Hz must be attenuated 3 db, and the product of R and C becomes:

$$RC = 1/(2\pi f) = 1/(2\pi)(240).$$

It is obvious that one such section is inadequate; the attenuation is considerable at a relatively high frequency (1000 Hz), and drops slowly (6 db per octave) below 60 Hz. This is redrawn as curve A in Fig. 5-5. Curve C shows a more idealized case using three networks in the circuit, while curve B, using two such networks, is a reasonable compromise.

Once again, the test equipment shown in Fig. 3-7 may be used for testing the response of a compensating circuit—this time, the rumble filter. The procedure and conclusions indicated under the scratch-filter discussion apply here as well. Although 60 Hz is the highest frequency considered in rumble-filter design, lower frequencies may also be generated and must be eliminated. While elimination of rumble should not be the concern of an amplifier, it is important that rumble be nullified. Excessive rumble can overload an amplifier or speaker, resulting in distortion at all frequencies.



(A) One filter. (B) Two filters. (C) Three filters.

Fig. 5-5. The effect on an amplifier using the rumble filter shown in Fig. 5-4. All curves are 12 db down at 60 hertz.

LOUDNESS CONTOUR

Tones of identical sound level, but differing in frequency, do not sound equally loud to the human ear. Experiments have been performed to relate the apparent loudness to the actual relative sound intensity at all frequencies of the audio spectrum. The most commonly accepted results were compiled in the 1930's by H. Fletcher and W. A. Munson. Fig. 5-6 shows the result of their research.

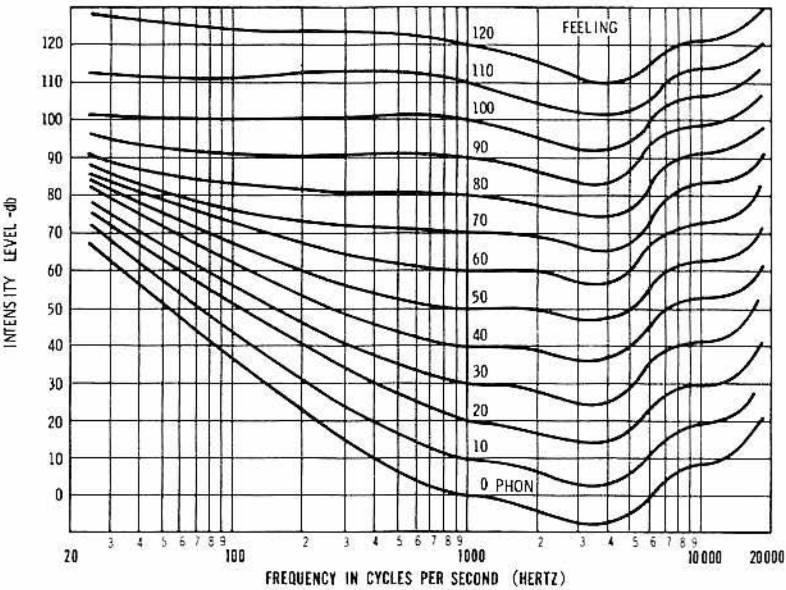


Fig. 5-6. Loudness-level contour curves (Fletcher and Munson).

In common with most audio measurements, the curves are based on a 1000-Hz reference frequency. With this in mind, let us study the curve marked "70."

Notice that 1000 Hz is at a level marked 70 db. This 70 db is the zero reference level for this particular curve. At the higher and lower frequencies, the curve goes up. This means that at these frequencies, the sound pressure must rise if tones at these frequencies are to sound as loud as the 1000-Hz note. The actual intensity must increase 10 db at both 60 Hz and 15 kHz if these frequencies are to seem equal in level to the 1000-Hz signal. It must also fall 5 db at 4000 Hz.

There are similar curves from "0" to "120." Each curve represents the equal loudness curve for different sound pressures, with 1000 Hz as the reference frequency. The equal-loudness contours vary with the actual intensity of the sound. At low listening levels (curves marked "0" to "30"), there must be a greater relative sound pressure at the low and high frequencies, if all frequencies are to seem equally loud, than at loud listening levels (curves marked "70" to "20"). The loudness control is based on this fact.

Each curve depicting the equal loudness contour is marked in units called "phons." The "phon" refers to an equal apparent loudness level. Even though the *actual* sound pressure varies several db over the frequency range, the rating in "phons" is identical when the *apparent* loudness is identical. The "phon" refers to the loudness contour curve, while the db rating refers to the relative *actual* sound pressure.

The phon is identical to the presure level in db at 1000 Hz if the 0-db level is considered to be an intensity of 10^{-16} watts/cm^2. At frequencies other than 1000 Hz, the phon varies from the intensity rating in db. It is generally accepted, for design purposes, that music is played and recorded at the 70-phon level and played back in the home at the 40-phon level. If music is to be reproduced exactly as originally performed, the apparent loudness contour must follow the 70-phon curve. An amplifier with a flat frequency response will reproduce sound accurately only at the 70-phon level. But music is reproduced in the home at the 40-phon level. The loudness control is used to reshape the 40-phon curve so that the relative intensity level will follow the contour of the 70-phon curve.

In the reshaping process, 1000 Hz is still the reference frequency. At 60 Hz, the 70-phon curve requires an increase of 10 db in actual intensity, if the output at 60 Hz is to be at the exact audible level as the note at 1000 Hz. Similarly, on the 40-phon curve, the increase in actual intensity required at this frequency is 30 db.

Putting this another way, if you were listening to music in the concert hall, your hearing would follow the 70-phon curve. The 60-Hz notes would seem 10 db lower than the 1000-Hz notes of identical amplitude. In the home, your hearing will follow the 40-phon curve because of the reduced volume in reproduction. Now, the 60-Hz note will seem 30 db lower than the 1000-Hz note of like intensity. To make the reproduced sound the equivalent of the original, your amplifier will have to compensate for the difference in the curves at the two listening levels. Circuitry can be used to increase the 60-Hz tone by 20 db (30 db − 10 db) over the signal at 1000 Hz. Now, when listening with compensation at the 40-phon level, the 60-Hz signal will be 10 db lower than the 1000-Hz signal. This is exactly the relationship that exists at the 70-phon level in the concert hall.

The difference between the two curves must similarly be reproduced at all frequencies: At 200 Hz the difference is 14 − 4 db, or 10 db; at 10 kHz, the difference is 12 − 12 db, or 0 db; at 3000 Hz, the difference is $(-3.5) - (-3)$ db, or -0.5 db. An exact loudness-contour compensation curve for making the 40-phon level sound like the 70-phon level is shown in Fig. 5-7. The compensation required when using an amplifier at the 60-phon level (for people who like their hi-fi very loud) and at the 20-phon level (for those who use their equipment at very low levels) is also shown.

It is quite simple, with proper circuitry, to provide these equal-loudness contours in the amplifier. At least one contour, at about the 40-db level, is frequently found in audio amplifiers. But how valid are any of these compensations?

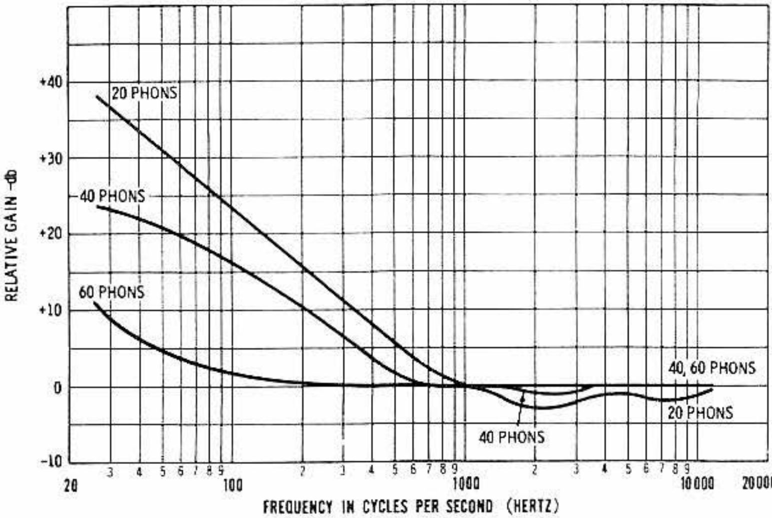


Fig. 5-7. Compensation for Fletcher-Munson curves; 70 phons is standard playback level.

To determine this, let us look at the curves by Robinson and Dadson, shown in Fig. 5-8. These were determined much later than the Fletcher-Munson curves. Fig. 5-8 also shows the effect of the listener's age, where the ability to hear high frequencies falls off with increased age.

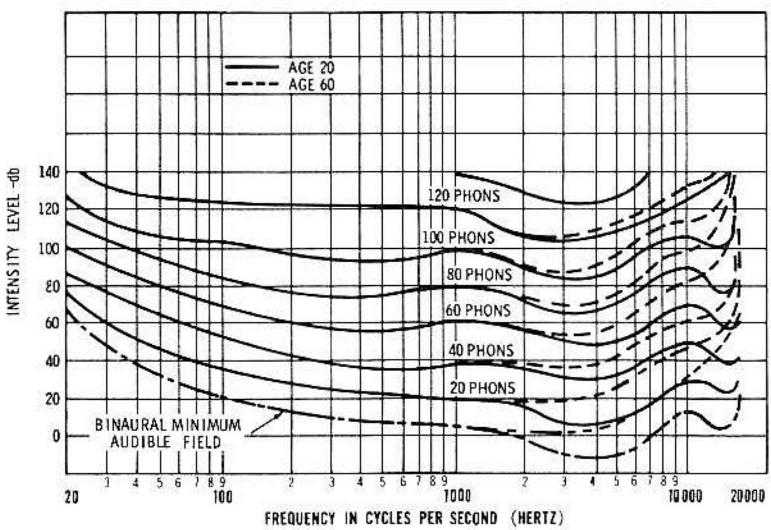


Fig. 5-8. Equal-loudness contours (Robinson and Dadson).

The compensation required for the Robinson-Dadson curves is shown in Fig. 5-9 as interpolated from Fig. 5-8. Again they are referenced to a 70-phon performance level. Note that with the Robinson-Dadson curve, less loudness compensation is required than with the Fletcher-Munson curve. The former also requires treble boost. Which set of curves is correct? Which set should be considered the standard when designing for the loudness contour? Nobody knows.

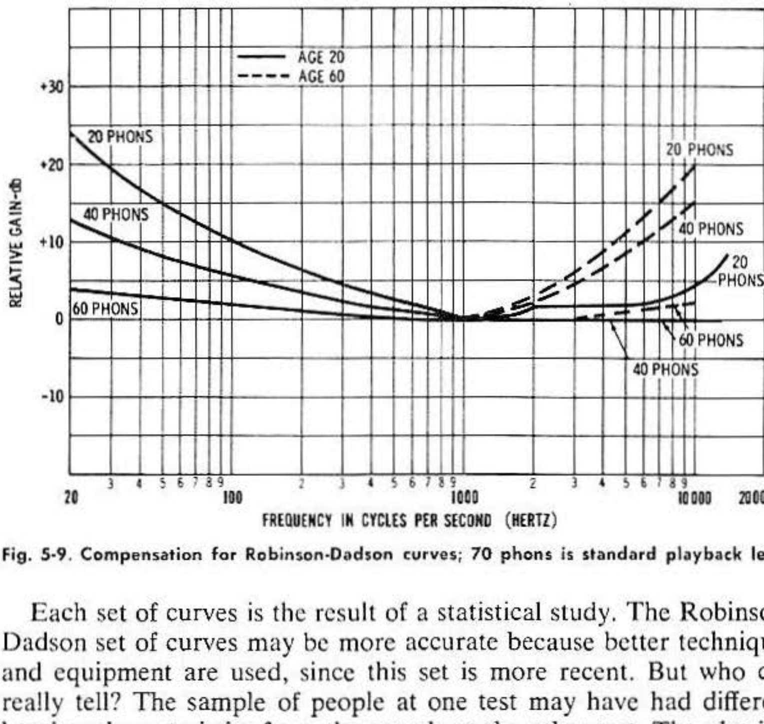


Fig. 5-9. Compensation for Robinson-Dadson curves; 70 phons is standard playback level.

Each set of curves is the result of a statistical study. The Robinson-Dadson set of curves may be more accurate because better techniques and equipment are used, since this set is more recent. But who can really tell? The sample of people at one test may have had different hearing characteristics from the sample at the other test. The physical and mental health of the two groups of people may have been significantly different. Both factors would affect results.

Other considerations also must be taken into account. For one, the response to the pure (sinusoidal) tones used in either test is probably unlike the response to complex musical tones. Equal-loudness contours are also affected by the relative direction of the sound source from the individual. These factors are pointed out and both sets of data are presented here to emphasize the impossibility of setting any meaningful standard contour curve for loudness compensation at this time. It is a highly subjective variable. Best personal compensation can be attained when each individual sets the tone controls to best suit his own hearing and particular room acoustics.

It should furthermore be remembered that 70 db is the average loudness level near the orchestra. At a distance the level changes, and so does the equal-loudness contour curve. The curves for those sitting at the rear of the concert hall differ from the curves for those in the front-row seats. Just what the listening level and associated contour curve are is thus determined by the distance from the orchestra. The curve one prefers is a subjective matter and should not be dicated by the whim of an amplifier designer. Each curve at each level is correct.

No matter what type of loudness compensation is provided for on the amplifier, it may be checked with reference to a 1000-Hz, 0-db level, using the test equipment in Fig. 3-7. All controls on the amplifier must be set for a flat response. The loudness control can then be switched into the circuit, and the effect on the frequency response noted. No stage in the amplifier should be overloaded at any time during the test. The entire audio range should be scanned, and the curve plotted. Just which set of data your curve should be compared with is your choice.

CHAPTER 6

Harmonic Distortion—Tests and Measurements

A discussion of harmonic distortion should begin with a thorough study of the Fourier analysis of nonsinusoidal waves. Since this topic has been most adequately covered in many excellent engineering texts, it will not be repeated here. Much of the information included below will be taken from the analysis, however, and will be stated rather than derived.

A REVIEW OF TRIGONOMETRY

In Fig. 6-1A, R is one of the radii of a circle. As the tip of the radius (a point on the circumference) rotates from o to a, the radius must rotate through an angle θ. Angle θ can be measured in degrees. As drawn in Fig. 6-1A, θ is 45 degrees, and in Fig. 6-1B, θ is 90 degrees. All the values of θ are noted on the drawing, until it rotates through the complete 360 degrees. If it rotates twice through 360 degrees, θ will equal 720 degrees.

The angle can also be designated in radians. There is a direct relationship between degrees and radians. In Fig. 6-1F, 360 degrees equals 2 radians. It follows that $2 \times 360° = 2 \times 2\pi$ radians, or $720° = 4$ radians. Several other intermediate relationships are shown in Fig. 6-1.

If the radius rotates at an angular velocity of ω radians per second, and if t is the time required by the radius to rotate through angle θ, the relationship is:

$$\theta = \omega t \qquad (6\text{-}1)$$

where,

θ is the angle,
ω is the radians per second,
t is time.

This is the basic equation that distance is equal to velocity multiplied by time; "ω" is a basic notation found in harmonic analysis. It is angular velocity, in terms of the number of radians the radius turns each second. A radius rotating through 360 degrees has completed one cycle. A velocity of one cycle per second is the equivalent of 2π radians each second. Two cycles completed each second is

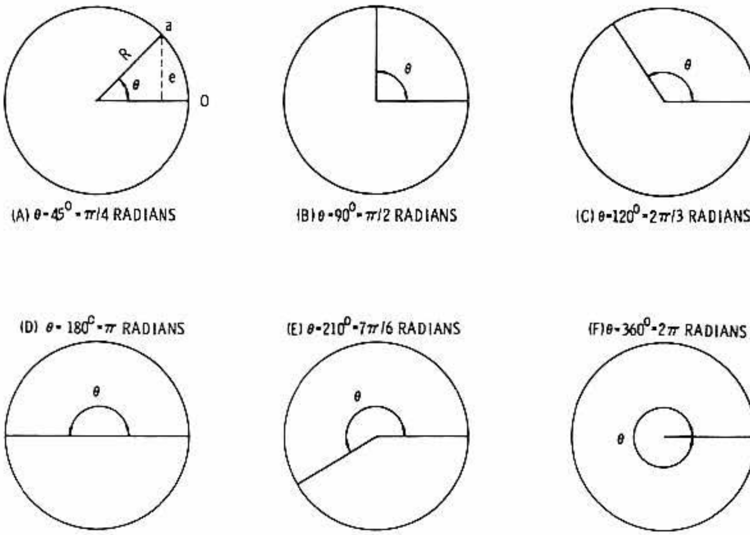


Fig. 6-1. The various angles of theta (θ).

the equivalent of covering $2 \times 2\pi$ radians per second. Thus, ω can be stated mathematically in two ways, with identical meaning:

$$\omega = \frac{2\pi}{T} = 2\pi f \qquad (6\text{-}2)$$

where,

T is the time for one cycle (period),
f is the frequency in cycles each second.

Returning to Fig. 6-1A, it is desirable to find the sine of angle θ for all values of θ. By definition:

$$\sin \theta = e/R \qquad (6\text{-}3)$$

the opposite side over the hypotenuse. It should be noted that the length of the radius remains unchanged for all values of θ. For the sake of simplicity, R is made equal to 1. Then:

$$\sin \theta = e \qquad (6\text{-}4)$$

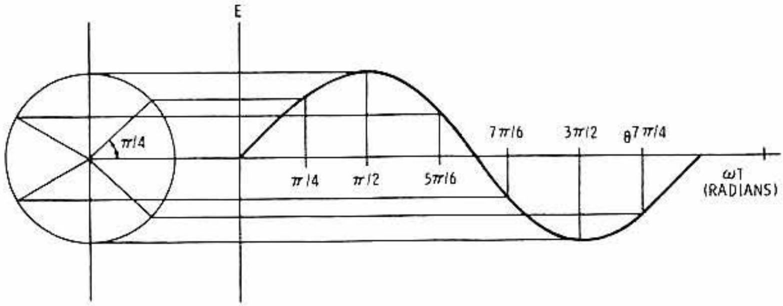


Fig. 6-2. Graphical portrayal of the generation of a sine wave.

By the simple construction in Fig. 6-2, the sin-θ function which is equal to sin t—for $\theta = \omega t$ from equation 6-1—can be plotted on the axis, showing how e varies with ωt. Each cycle must go through 2π radians.

The plot can be changed from being a function of ωt to a function of time. This is shown in Fig. 6-3A. When the wave goes through a complete cycle in 1 second, its frequency is 1 cycle per second. If it does this twice in one second, its frequency is 2 cycles per second, as shown in Fig. 6-3B. Each complete cycle represents an angle of 2 radians. The radius has traversed these 2π radians twice in Fig. 6-3B. The equation for the sine wave in Fig. 6-3A can be written as:

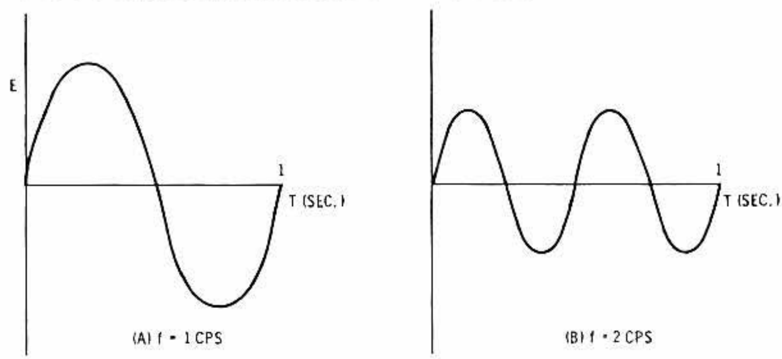
$$e_1 = E_1 \sin \omega t \qquad (6\text{-}5)$$

and for Fig. 6-3B as:

$$e_2 = E_2 \sin 2\omega t \qquad (6\text{-}6)$$

where,

E is the peak value of the sine wave,
e is the value it assumes each instant of time.

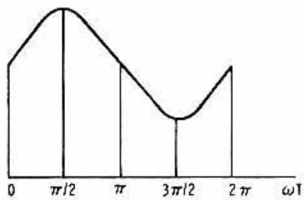


(A) One cycle.

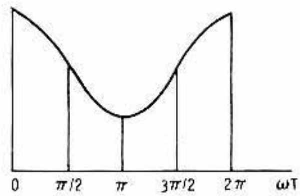
(B) Two cycles.

Fig. 6-3. A typical sine wave.

The exact frequency is immaterial, but the relationship between the two is represented by a factor of 2, indicating that one is double the frequency of the other. The cos-θ curve is identical in shape to the sin-θ curve. It just starts 90 degrees, or $\pi/2$ radians, later. The two are compared in Fig. 6-4.



(A) Sine-wave cycle.



(B) Cosine-wave cycle.

Fig. 6-4. Comparison between one sine-wave cycle and one cosine-wave cycle.

DEFINING HARMONIC DISTORTION

From the Fourier analysis, it can be determined that a distorted wave is composed of the original frequency (the fundamental), as well as frequencies which are two, three, four, and so on, times the original frequency (harmonics). Stated analytically, it can be written:

$$e = E_1 \cos \omega t + E_2 \cos 2\omega t + E_3 \cos 3\omega t + \ldots + B_1 \sin \omega t + B_2 \sin 2\omega t + B_3 \sin 3\omega t + \ldots \qquad (6\text{-}7)$$

where,

$\cos \omega t$ or $\sin \omega t$ is the fundamental,
$\cos 2\omega t$ and $\sin 2\omega t$ represent the second harmonic,
$\cos 3\omega t$ and $\sin 3\omega t$ represent the third harmonic,
E_1, E_2, E_3, B_1 B_2, B_3 are the voltage peaks of the harmonics.

The remaining terms of the equation, representing the fourth, fifth and sixth harmonics, up to the infinite harmonic, are not shown, but should be understood to exist. Harmonics greater than the third are usually too small to be of any significance, and consequently will not be considered here.

Not all the harmonics in equation 6-7 are present in a distorted wave. By careful choice of the 2π axis, several of the terms in the equation can be discarded as contributing nothing to the wave. Several examples are shown in Fig. 6-5.

Assume the 2π axis can be chosen as shown in Fig. 6-5A, so that points equidistant from and on either side of the 2π axis are identical (with respect to the 2π axis) in amplitude, and in the same (positive or negative) direction. That is, the amplitudes at t and $-t$ are identi-

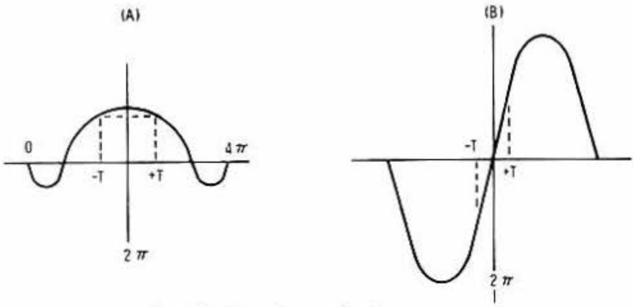


Fig. 6-5. Two harmonic sine waves.

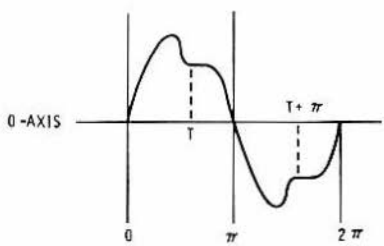
cal. Then, only cosine terms will be present in the equation. The sine components in equation 6-7 can be discarded.

Assume that the curve takes the form shown in Fig. 6-5B. Here the amplitude at t and −t are identical, but one is the negative of the other. In this arrangement, only the sine terms in equation 6-7 apply. Some functions can be either odd or even (see Fig. 6-5), depending on the choice of the 2π axis. The result of the analysis is the same. The choice is determined by convenience. The only difference is the phase and thus does not affect the harmonic content.

Either choice can exhibit both even and odd harmonics. From Fig. 6-6, we can see one more criterion to simplify analysis. If the amplitude at any time (t) is identical, but opposite in sign, to the amplitude at the time $(t + \pi)$, only odd harmonics are present in the curve. The even-harmonic terms can thus be eliminated from equation 6-7. When this is combined with the considerations in Fig. 6-5, most of the terms in equation 6-7 can be eliminated. Do not confuse the m with the even and odd functions defined in Fig. 6-5. In Fig. 6-6 we refer to the order of harmonics only, while in Fig. 6-5 we identify cosine functions as even and sine functions as odd. This has nothing to do with the odd or even harmonics discussed in Fig. 6-6.

An easy way to identify the presence of odd harmonics is to move the section of the curve between π and 2π under the section from 0 to π. If one is the mirror image of the other around the 0 axis, then

Fig. 6-6. Odd-harmonic sine wave.



only odd harmonics are present. A curve which does not exhibit mirror symmetry around the t axis is composed primarily of even harmonics. Through the remainder of this discussion, it will be assumed that only cosine terms are present in the distorted signal. All higher harmonics are supposed to be so small as to render them negligible.

The percent of distortion is defined as:

Percent HD =

$$\frac{\text{(sum of the amplitudes of the harmonics squared) } 1/2}{\text{amplitude of the fundamental}} \times 100$$

$$= \frac{(E_2^2 + E_3^2)^{1/2}}{E_1} \times 100 \qquad (6\text{-}8)$$

where,

E_1 is the magnitude of the fundamental,
E_2 is the magnitude of the second harmonic,
E_3 is the magnitude of the third harmonic.

This exact formula can be used if all components are known from measurements on a wave analyzer. In most cases, equation 6-9 is used:

$$\text{Percent harmonic distortion} = \left(\frac{E_2^2 + E_3^2}{E_1^2 + E_2^2 + E_3^2}\right)^{1/2} \times 100 \qquad (6\text{-}9)$$

This is effectively identical to equation 6-8 when the harmonics are small—not more than 10 percent of the fundamental. The conventional harmonic-distortion meter measurements comply with equation 6-9 and are thus inaccurate for large percentages of distortion.

OBSERVING DISTORTION OF AN OSCILLOSCOPE

Distortion magnitudes of 5 percent (or in some cases, 10 percent) or less can seldom be seen on the scope. Inserting the differentiating network between the signal and the scope, as shown in Fig. 6-7, will frequently make it possible to observe small percentages of distortion.

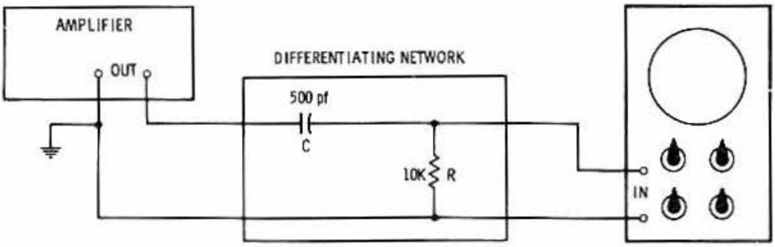


Fig. 6-7. Differentiating network used for checking distortion.

This can be explained by noting that the differentiating network discriminates against the low frequencies more than it does against the high frequencies. If proper values of resistance and capacitance are chosen, the amount of low-frequency fundamentals passed on to the scope will be reduced, while the high-frequency harmonics will be passed on freely. As a result, the curve observed on the scope will have a greater percentage of harmonic components than the original signal had. If distortion due to harmonic components is present in the original signal, the effect of the network on the distorted shape of the signal will be emphasized. An analytic derivation is given in Appendix F.

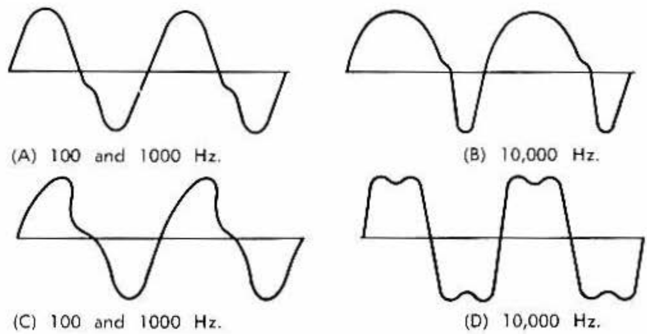


Fig. 6-8. Second- and third-harmonic waveforms.

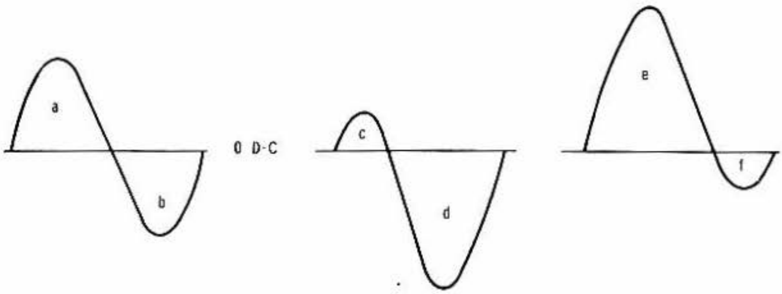
A signal with second-harmonic components was applied to the network in Fig. 6-7. The input to the network looked perfect. The output at 100 Hz and 1 kHz looked like curve A in Fig. 6-8, and at 10 kHz it looked like curve B. When a second group of apparently perfect signals was fed to the network, third harmonics appeared at the output, as shown in C and D of Fig. 6-8.

Other qualitative methods are possible to indicate small amounts of distortion. For example, an LC network, resonant at about 50 kHz, can be put in series with the load. An undistorted signal will pass through without alteration. When distortion is present, it will trigger the network into oscillation. This condition will turn up as pockets of oscillation in an otherwise perfect-looking sine wave.

From all the possible qualitative methods, it appears that the differentiating network in the output circuit is about the most effective in causing small amounts of distortion to be perceptible on an oscilloscope screen.

ESTIMATING HARMONICS FROM OSCILLOSCOPES

Several methods have been developed for estimating the harmonic components of signals appearing on the oscilloscope screen. Analyses of these signals have appeared in numerous books. The complete analysis, using these methods, is most useful for complex waves. In the description presented here, only second- and third-harmonic components are considered to be of any significance. The analysis will be further simplified by the assumption that in all practical situations, the 2π axis can be chosen as shown in Fig. 6-5A. It is thus assumed that only cosine functions and second- and third-harmonic terms will be present in the signal. One simplifying factor has been assumed up to now and will continue to be used. This requires some explanation.



(A) Zero d-c. (B) Negative d-c component. (C) Positive d-c component.

Fig. 6-9. Comparison of d-c levels in harmonic distortion.

Equation 6-7 assumes a zero d-c component. This means that the area between the signal and the zero axis above the axis is equal to the area between the signal and the zero axis below the axis. In the curve shown in Fig. 6-9A, the area *a* above the axis is equal to the area *b* below the axis. Because area *a* minus area *b* is equal to zero, the d-c component must be zero.

Looking at the zero axis as shown in Fig. 6-9B, you can see that a negative d-c component will exist, since the area *d* under the zero axis is greater than that of *c* above the axis. A positive d-c component exists in Fig. 6-9C, since *e* is greater than *f*. Actually, all three curves are identical. Only the placement of the zero-axis line determines whether d-c components will be present in the mathematical analysis. It simplifies matters considerably if the axis is chosen so that no d-c will be present. The analysis can then be completely accomplished with equation 6-7.

It should also be pointed out that curves do not have to be sinusoidal to have a zero d-c component. They can take any shape, as shown in Fig. 6-10. Considering all these factors, the analysis below is based on the Fisher-Hinen method. See appendix G for the derivation of methods for determining the second- and third-harmonic distortion terms from the oscillograms.

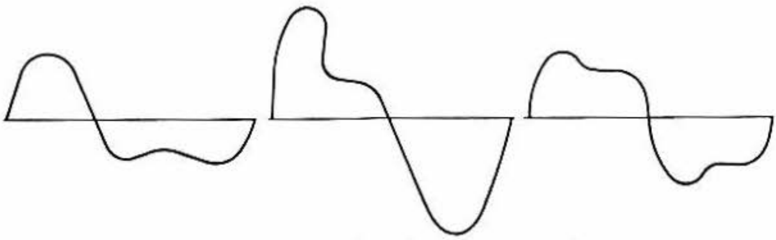


Fig. 6-10. Examples of curves adjusted to have a zero d-c component.

A curve containing second-harmonic distortion is the sum of the fundamental component and the second harmonic. This is illustrated in Fig. 6-11. The practical laboratory procedure for determining the amount of second-harmonic distortion can be stated simply in several steps. Use a scope with d-c amplifiers and a repetitive sweep. Refer to Fig. 6-11 for the position of the ordinates.

1. Determine that the curve on the oscilloscope screen has primarily even harmonics, as discussed above.
2. Set up the scope so that two cycles of measurable amplitude are on the screen.
3. Turn off the amplifier under test. Set the vertical-position control on the scope so that the swept line is coincident with the center of the scope screen. This sets the center of the scope screen at the 0 d-c level. Do not touch the vertical-position control after this.
4. Turn the amplifier on. Adjust the horizontal-gain control and position control so that one cycle at the fundamental frequency

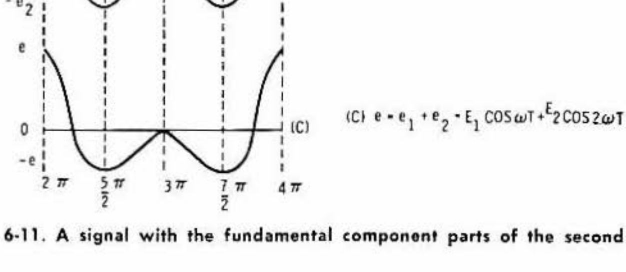


Fig. 6-11. A signal with the fundamental component parts of the second harmonic.

occupies a measurable portion of the screen. The cycle should be arranged so that it appears as an even function (see Fig. 6-5A).

5. Measure the amplitude at the beginning of this signal. Refer to this as $e_{2\pi}$.
6. Measure the amplitude at the middle of the curve. This is $e_{3\pi}$.

Find the second harmonic from the following equation:

$$E_2 = \frac{e_{2\pi} + e_{3\pi}}{2} \qquad (6\text{-}10)$$

and the fundamental from this equation:

$$E_1 = \frac{e_{2\pi} - e_{3\pi}}{2} \qquad (6\text{-}11)$$

Using equations 6-10 and 6-11, determine the percentage of the second-harmonic distortion from:

$$\text{percent second harmonic} = \frac{E_2}{E_1} \times 100 \qquad (6\text{-}12)$$

The practical step-by-step procedure for determining the percentage of third harmonic distortion is identical to that applying to the second-harmonic case, with the exception of the ordinates under consideration. Before the step-by-step procedure is applied, it must be determined whether the signal contains second or third harmonics. Then proceed accordingly with the test. Refer to Fig. 6-12 to determine the exact location of $e_{2\pi}$, $e_{8\pi/3}$, and $e_{10\pi/3}$, the three ordinates. Use equations 6-11, 6-12, and 6-13 in appendix G to calculate the distortion.

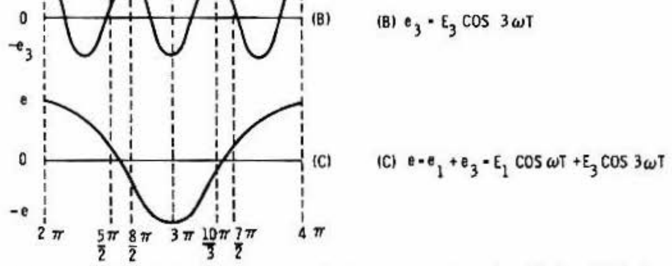


Fig. 6-12. A signal with the fundamental component parts of the third harmonic.

EXACT MEASUREMENT TECHNIQUES

So far, two methods for determining harmonic distortion were discussed. The first, using a differentiating network, merely indicates whether any distortion is present in the signal. The second is a method of estimating the quantity of distortion.

Instruments for measuring harmonic distortion were discussed in Chapter 1. One instrument, the wave analyzer, measures the amplitude of each harmonic component and the fundamental. The data from these measurements are substituted into equation 6-8 to calculate the percentage of harmonic distortion.

Some engineers prefer to use a weighted value for percentage of distortion. It is weighted as a function of the number of the harmonic. The weighted distortion factor is:

$$\frac{1}{2}\left(\frac{\sqrt{(2E_2)^2 + (3E_3)^3 + (4E_4)^4 + (5E_5)^5 + \dots}}{E_1}\right) \times 100 \quad (6\text{-}13)$$

Whichever system provides the most valid number for the percentage of distortion, it is certain that stated specifications will reject the use of equation 6-13 because the distortion figure is larger than that provided by equation 6-8.

A second instrument, the harmonic-distortion meter, first measures the size of the composite signal—the fundamental plus the harmonics. The fundamental is then filtered out, and the remaining harmonics measured. The ratio of the two measurements multiplied by 100 provides the solution to equation 6-9 for percentage of harmonic distortion. This percentage can be read directly on the meter, for the two measurements are only relative. The reading is accurate at low percentages of distortion.

Exact readings must be made under exacting conditions. Several factors must be taken into account so that the measurements will be valid.

1. The response of the amplifier under test should be relatively flat. Deviations will tend to weigh the reading in one direction or the other. Switch out all filters and adjust tone controls to the best position for a reasonably flat response.
2. Use only high-level inputs. Phono and tape-head inputs do not have a flat response and will give weighted results.
3. Feed the amplifier from a low-distortion signal generator, or use filters at the output of the generator.
4. The line voltage is a factor in the distortion reading. Adjust the line voltage to a standard value with a variac, and monitor it with a meter throughout the test. Use the line voltage at which

the amplifier is rated. If the range of voltage is not given, assume 117 volts to be the standard.

5. Make tests after the unit has been on for one hour. The power transformer heats up with time, causing the d-c and filament voltages to drop. One hour can be considered as the average time an amplifier may be used. This can be used as a reasonable standard.
6. Measurements should be made at all audio frequencies.
7. Measurements should be made at the 1-watt level as well as at the rated power of the amplifier.
8. Observe the output from the distortion meter on an oscilloscope. This is important in determining just what portion of the measurement is due to distortion and what portion is due to hum and noise. The percent of the measurement which represents distortion should be calculated from the thickness of the signal as it appears on the scope.

The dividing line between high fidelity and ordinary reproduction has been considered 0.5-percent harmonic distortion. Amplifiers with lower percentages of distortion do sound better and are thus most desirable.

CHAPTER 7

Intermodulation Distortion

The nonlinear characteristics of audio amplifiers are measured in various ways. The most common is to check the harmonic components. Unfortunately, correlation between harmonic content and listening quality has been poor. Intermodulation-distortion measurements are used with the hope of better matching measurements with the subjective tests.

NONLINEAR CHARACTERISTICS

Like harmonic distortion, intermodulation distortion can be caused by the nonlinear characteristics of the amplifying device. This nonlinearity is shown by the curves which describe the operation of these devices. If a curve for the 12AT7 tube, for example, were plotted, assuming a load resistor of 30,000 ohms in the plate, the resultant nonlinearity would be obvious, as shown in Fig. 7-1. It can be seen that a change of −2 volts from the operating point, −3 volts, to −5 volts in grid potential causes a change of 2.4 ma in the plate current, while a change from −3 volts to −1 volt causes a plate-current change of 2.7 ma. If the curve were linear, a grid-voltage change of 2 volts in either direction would indicate a plate-current change of 2.4 or 2.7

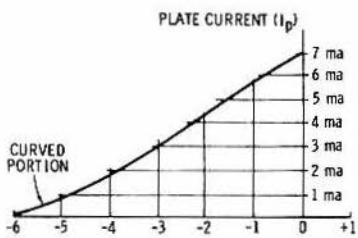


Fig. 7-1. Typical grid-voltage and plate-current characteristic curve.

ma at all times. If a 6-volt peak-to-peak sine-wave signal were applied to the grid circuit of this tube, one half would be amplified more than the other half, as shown in Fig. 7-2. This distortion of the original shape of the signal is due to the nonlinear characteristic of the tube.

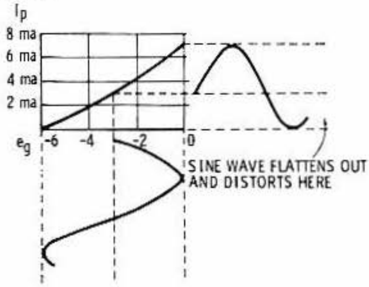


Fig. 7-2. A sine wave applied to a nonlinear portion of the tube characteristic.

MODULATION

To the technician, modulation is not a new concept. The radio station sends out a modulation signal, as illustrated in Fig. 7-3. When analyzed mathematically, an amplitude-modulated signal can be seen to consist of a high-frequency carrier, such as 1 MHz, with an audio signal, such as 400 Hz, changing the strength or amplitude of this carrier. The result is a 1 MHz wave varying 400 times a second in amplitude. When the variation of the 400-Hz modulating signal is great in amplitude, the peaks and crests of the 1 MHz carrier are greater; when the 400 Hz is low in amplitude, the carrier varies to a smaller degree.

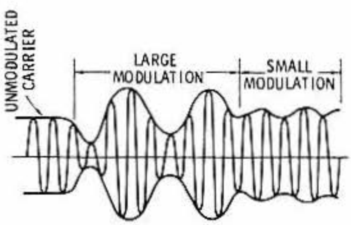


Fig. 7-3. Waveform of a modulated signal.

Also, this variation creates new frequencies. Not only are the 1 MHz and the 400-Hz signals being transmitted, but there are also sum and difference frequencies present. Thus, because of this modulation, the following four frequencies (in hertz) are present: 1,000,000 Hz, 400 Hz, 1,000,400 Hz, 999,600 Hz. The latter two frequencies are known as the sidebands.

The same principle of modulation with sidebands is used in every superheterodyne radio receiver. The 1000 kHz arriving from the radio station is mixed with 1455 kHz generated by the local oscillator in the radio. The result is the creation of the sum freqency, 2455 kHz, and the difference frequency, 455 kHz. Only the 455-kHz sideband is amplified by the i-f amplifier; the 2455-kHz sideband is discarded. This mixing of the two signals by the first detector is accomplished because of the nonlinear action of this first tube. If this tube were perfectly linear as far as its input-voltage—output-current characteristics were concerned, there would be no mixing and no 455-kHz sideband. (There are linear methods of modulation, but these are seldom applied to a-m radios).

GENERATING INTERMODULATION DISTORTION

When this theory of modulation is extended to audio equipment, the mechanics of intermodulation distortion becomes obvious. In music, there is always more than one frequency present. Assume, in the simplest case, that there are only two frequencies available—100 Hz and 5000 Hz. If the amplifier were perfectly linear, there would be only two frequencies coming out of the unit—100 and 5000 Hz—neither one of which would be distorted or mixed in any fashion. However, if the amplifier were not perfectly linear—as is usually the case—the 100 Hz and the 5000 Hz would mix, modulate each other, and there would be the addition of the sum and difference frequencies, 5100 Hz and 4900 Hz. A mathematical proof of this is given in Appendix H. The amount of these sum and difference frequencies present would constitute the intermodulation distortion.

However, this distortion goes one step further. Since the amplifier is nonlinear, there is also harmonic distortion present. Thus, not only are there 100 Hz and its harmonics such as 200, 300, and so on, but there are 5000 Hz and its harmonics such as 10,000, 15,000, and so on. There are also the sum and difference frequencies of these harmonics present to add to the intermodulation distortion.

The higher harmonics usually have small amplitudes and may be considered negligible. This by itself would narrow down the intermodulation-distortion components considerably. The components can be further narrowed down when it is observed that higher-order sidebands and harmonics are outside the audio range of 20 Hz to 20 kHz. For this discussion, it will be satisfactory to use the two fundamentals and their sidebands as the sole factors contributing to intermodulation distortion.

This might give a clue as to why intermodulation tests correlate better with listening tests than do harmonic-distortion measurements. Harmonics are present in music fed to an amplifier. The added harmonics produced by an amplifier will therefore be masked somewhat by the music. They may not be noticeable to the listener, and hence, will not be considered too objectionable. But, if intermodulation distortion is significant, sum $(\omega_1 + \omega_2)$ and difference $(\omega_1 - \omega_2)$ frequencies of two musical notes will be created in the amplifier. These frequencies will tend to be more noticeable, even at low levels of distortion.

Whatever the true explanation for the correlation may be, one thing is certain: the correlation is there. The exact mechanics for the creation of a variation in the amplitude of the high-frequency wave due to the modulation by a lower frequency can be seen in Fig. 7-4. Here, the high frequency is superimposed on the lower frequency in the grid input circuit. Because of the nonlinearity of the relationship between the grid input voltage to plate output current, there is a variation of the high-frequency amplitude in the plate circuit of the tube. The resultant amplitude of the high-frequency components is shown below the output wave as a modulated signal.

The test for intermodulation distortion is straightforward. Two frequencies are fed simultaneously to an amplifier. If the amplifier is linear, only these two frequencies will appear at the output. If there is nonlinear distortion, other frequencies will be present at the output, along with the two signals which comprise the input. The presence of newly created frequencies in a nonlinear system can readily be deter-

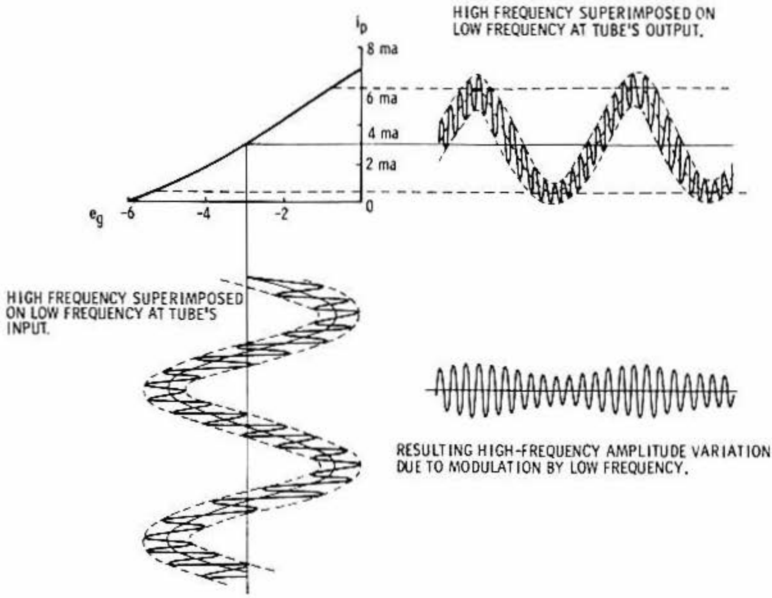


Fig. 7-4. Results of feeding two signals of different amplitudes to the curved portion.

mined from the mathematical analysis in appendix H. A large variety of tests are possible for measuring intermodulation. Two are used in various phases of the audio industry. One of them has been adopted by most hi-fi manufacturers.

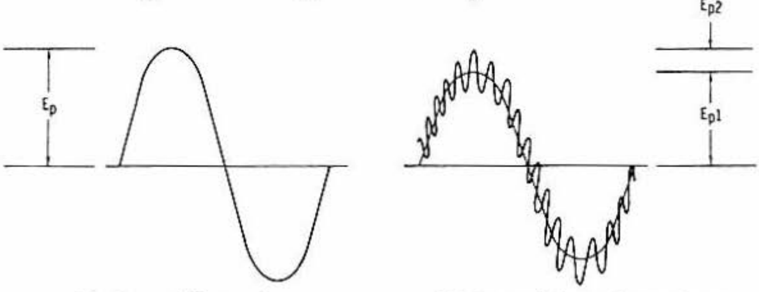
The first, recommended by the Society of Motion Picture and Television Engineers (known as the SMPTE method), is used as a standard by much of the hi-fi industry. This method specifies that one low frequency (between 0 and 400 Hz) and one high frequency (between 1000 Hz and 12 kHz) be fed simultaneously to the amplifier under test. Stringent test methods require that the higher frequency be half the upper-frequency limit of the amplifier and that the lower frequency be the low-frequency limit of the amplifier.

A second test would be performed using the same upper frequency, with 100 Hz as the lower frequency. However, the industry has more or less settled on one test only, usually using 60 and 7000 Hz. These are referred to as ω_2 and ω_1 respectively.

The ratio of the amplitudes of the lower to the upper frequency is 4:1, or a difference of 12 db. The output from the generator appears somewhat as shown in Fig. 7-5B. One pair of sidebands generated using this type of test is at frequencies $(\omega_1 + \omega_2)$ and $(\omega_1 - \omega_2)$. Call the amplitudes of these sidebands A_A and A_B, respectively. Another pair of possible sidebands is $(\omega_1 + 2\omega_2)$ and $(\omega_1 - 2\omega_2)$. Refer to the amplitudes of these frequencies as A_C and A_D, respectively. There are additional sidebands, but they are not required in this analysis.

If the relative amplitudes of individual sidebands can be measured on a harmonic analyzer, the percentage of intermodulation distortion can be determined from equation 7-1. E_2 is the amplitude of the higher frequency fed to the unit under test.

$$\left(\frac{(A_A + A_B)^2 + (A_C + A_D)^2}{E_2^2}\right)^{1/2} \times 100\% \qquad (7\text{-}1)$$



(A) Sinusoidal signal.

(B) Sinusoidal signal mixed at a ratio of 4:1.

Fig. 7-5. Intermodulation sine waves.

The SMPTE method indicates the effect which low-frequency nonlinearity has on a high frequency. The second, or CCIF (International Telephonic Consultive Committee) method of measuring intermodulation distortion, is used to check the other end of the band.

In this latter method, two high frequencies of identical amplitude are fed to an amplifier. If distortion is present, sum and difference frequencies will be formed. The difference frequency is considered more indicative of distastful distortion, so that the percentage of distortion, using this method, can be determined from:

$$\frac{\text{Amplitude of difference frequency}}{\text{sum of amplitudes of two test signals}} \times 100\%$$

Once again, a harmonic analyzer can be used to determine the relative amplitudes of the various components. If the two high frequencies are distorted, the generated low frequency will indicate the amount of distortion present. All portions of the high-frequency end of the spectrum can be checked simply by shifting (in frequency, not amplitude) the two test frequencies while maintaining a constant difference frequency. An increase in amplitude of this difference frequency indicates just where the amplifier under test fails.

Each method provides significant results. It is impossible to pinpoint one method as more indicative of quality than the other. The SMPTE method will be discussed in detail only because it has become more widely used and not because of any apparent superiority.

SMPTE METHOD

A test setup used for measuring intermodulation distortion is shown in Fig. 7-6. The output from the amplifier is connected across an accurate load resistor, R_L. If the output is sinusoidal, the power, W, delivered by the amplifier is determined by measuring the rms voltage, E, with the output meter. The power may then be calculated from:

$$W = E^2/R_L \qquad (7\text{-}2)$$

where,

W is the power,
R_L is the value of the load resistor,
E is the rms voltage.

The peak voltage is E_p. The rms value of this, E, is $E_p/\sqrt{2}$. The rms voltage across the resistive load, R_L, is used to determine the amount of power dissipated in the load. Making use of equation 7-2, and letting $E = E_p/\sqrt{2}$ and $E^2 = E_p{}^2/2$, the power due to the sine wave in Fig. 7-5A is:

$$W = (E_p{}^2/2)\ (1/R_L) \qquad (7\text{-}3)$$

In order to measure intermodulation distortion, a signal similar to that in Fig. 7-5B must be fed to the amplifier. The peak voltage at the output of the amplifier must be identical to that of E_p in Fig. 7-5A, if the intermodulation test is to be made at the equivalent level as the harmonic-distortion test described in Chapter 6. This is necessary to maintain the identical voltage swing in the amplifier in both cases.

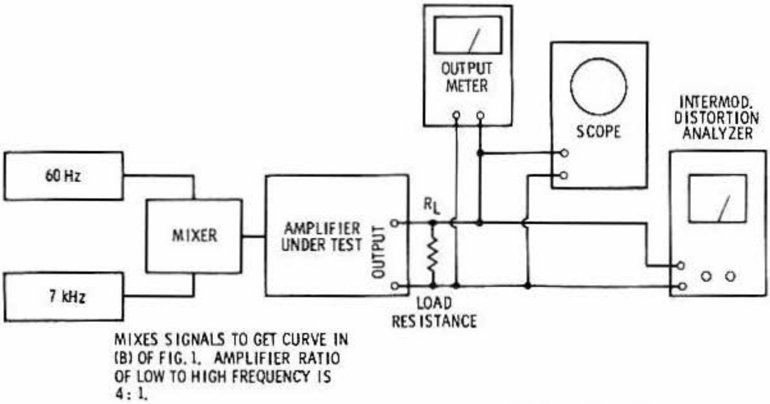


Fig. 7-6. Test setup for measuring intermodulation distortion.

To produce this condition:

$$E_p = E_{p1} + E_{p2} \qquad (7\text{-}3)$$

where,

E_p is the peak amplitude of the sine-wave signal,
E_{p1} is the peak amplitude of the low-frequency signal,
E_{p2} is the peak amplitude of the high-frequency signal.
E_{p1} and E_{p2} are both used in the intermodulation tests.

When a peak-to-peak type of output meter is used, the required output level can be found quite easily. The peak-to-peak levels are identical in both the sinusoidal and modulated cases.

For example, assume you wish to measure the intermodulation distortion at an output level of 25 watts, and the load resistor, R_L, is 16 ohms. If a pure sine wave were fed to the amplifier, the rms voltage across the 16-ohm load resistor would have to be, from equation 7-2:

$$E^2_{rms} = R_L W = 16\ (25) = (400\ \text{volts})^2$$

and,

$$E_{rms} = 20\ \text{volts}$$

The peak voltage is $E_p = \sqrt{2}\,E_{rms} = \sqrt{2}\,(20) = 28.2$. A peak-to-peak reading meter will read double this, to 56.4 volts. To measure intermodulation at this level, the peak-to-peak signal must also be

56.4 volts, and can be read directly on the peak-to-peak scale of the output meter.

Although the peak-to-peak output voltages in the sinusoidal- and modulated-signal cases are identical, the power delivered to the load is different in both instances. Because of this, a ratio must be established between the measured power on a meter and the actual power. A rigorous deviation of this ratio is given in Appendix I.

Assume a signal of 1 volt is superimposed on a signal of 4 volts. The maximum voltage applied will then be 5 volts. (See Fig. 7-5B.) Power is proportional to the square of the voltage for $P = E^2/R$.

Both frequencies deliver their individual amounts of power to the amplifier output. The true power output is actually the sum of the powers delivered by each frequency component. In this case, the output power is proportional to $(4 \text{ volts})^2 + (1 \text{ volt})^2$.

The power output for the maximum signal voltage (5 volts) is proportional to 5^2. The ratio of the maximum signal power to the actual power is $5^2/(4^2 + 1^2) = 25/17$, or 1.47. Thus, to find the true power, the actual power indicated on the meter when making the intermodulation test is multiplied by 1.47. Amplifiers are rated at this power, commonly called "equivalent sine-wave power." This refers to the power in a sine-wave signal whose peak voltage equals the peak voltage of the intermodulation signal.

The procedure for setting the output level from an amplifier can be accomplished as follows. The numbers stated above will be used in this example.

1. Assume you wish to measure the distortion at the equivalent of 25 watts of sinusoidal output. The rms power of the intermodulation signal is 17/25 of 25 watts, or 17 watts.
2. If the load resistor, R_L, is 16 ohms, the voltage across it for 17 watts of output is $V_{rms} = \sqrt{WR} = \sqrt{(17)}\ (16) = 16.45$ volts.
3. Adjust the input so that the rms-measuring output meter will show 16.45 volts. This is 25 watts of equivalent sine-wave power. Measure the intermodulation distortion at this level.

Little error would be introduced if an average measuring meter were to replace the rms measuring meter. In either case, the rms scale should be used for the reading. Use the intermodulation analyzer in accordance with the instructions supplied by the manufacturer of your particular instrument. On most instruments, the input to the amplifier under test and the analyzer are mounted on one common chassis. There is a common ground between the input and output. This can lead to complications.

On some stereo amplifiers, the 4-ohm tap on the output transformer is connected to ground. The ground at the input and the common ground at the output of these amplifiers must be isolated from each other to permit the intermodulation test. An isolation transformer must be used at the output of the amplifier, as shown in Fig. 7-7. The transformer should have at least double the power capabilities of the amplifier under test and should have a 1:1 turns ratio.

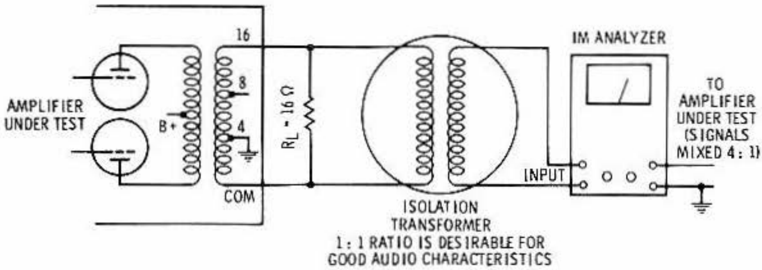


Fig. 7-7. Isolation transformer used when common tap of output transformer is not grounded.

TYPICAL IM ANALYZER

Fig. 7-8 shows a theoretical schematic of an intermodulation analyzer and the method of operations. In Fig. 7-9 two signals having an amplitude ratio of 4:1 are combined and fed into the audio amplifier.

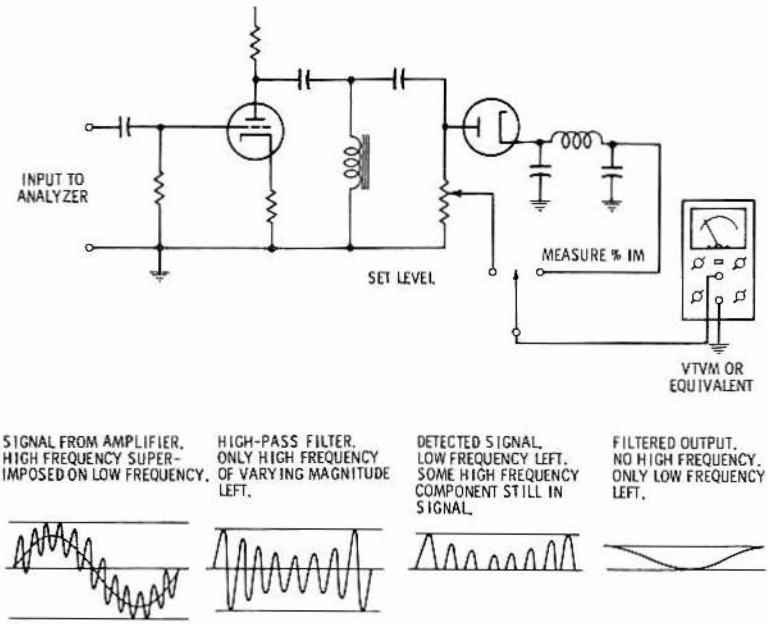


Fig. 7-8. A simplified schematic of an intermodulation analyzer.

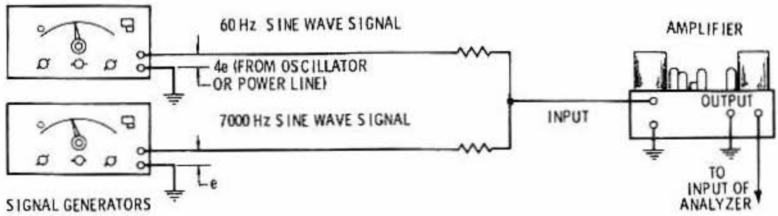


Fig. 7-9. Arrangement of generators and amplifier for making intermodulation test.

Coming out of the audio amplifier are the two signals modulated, with the amplitude of the low-frequency signal varying in accordance with the high-frequency signal. The high-pass filter in the analyzer eliminates the low-frequency component and passes only the high-frequency component, which has a low-frequency amplitude variation caused by the distortion in the amplifier. This modulated signal is used to set the reference-voltage level for a vacuum-tube voltmeter. The modulated signal then passes through a detector similar to that found in a radio receiver. The resultant signal is the low-frequency component which originally modulated the high frequency. The low-pass filter bypasses any of the high frequency left after detection, with the result that only the modulating low-frequency component is left. This component is a measure of the actual amount of intermodulation distortion created by this amplifier. Feeding this signal to the vtvm and comparing it with the original amplitude of the 7000-Hz modulated signal indicates the percentage of intermodulation distortion.

To specify intermodulation distortion by itself is not enough. The method used for testing is significant. Measuring intermodulation distortion using the SMPTE method, it will be found that distortion below 2 percent cannot readily be detected by the ear. Values measured by the CCIF method cannot be related directly to those obtained by the SMPTE method.

PREAMPLIFIERS

The circuit shown in Fig. 7-6 can be used to measure distortion in preamplifiers as well as power amplifiers, although, when testing preamplifiers, the load resistor, R_L, must be omitted from the setup. It should be replaced by a 100kΩ resistor shunted by a 1000-pf capacitor.

The 17/25 fraction is no longer valid, for it is not power that is of prime concern here, but solely voltage. The usual convention is to rate the intermodulation of a preamplifier at levels determined by an output reading on an average or rms measuring instrument. The 4:1 ratio of signal amplitudes fed to the power amplifier can also be applied to the low-gain section of the preamplifier because the gain is relatively uniform over the entire audio spectrum. The percentage of distortion can be read as in the power-amplifier case.

But how should equalized stages be measured? No convention has been established to measure distortion originating in this section of the amplifier. But the following considerations should be noted.

If the ordinary 4:1 mixed signals were fed to an equalized preamplifier, the high-frequency sideband components would usually be attenuated and the low frequencies emphasized. This type of signal mixture at the input to equalized playback stage is not usual.

Let us assume a complete recording system is under test. Two signals at 4:1 ratio would be fed to the recording preamplifier and recorded on a disc or tape. When these two signals are played back, their ratio at the output of a properly equalized preamplifier would still be 4:1; but there would also be the additional components produced because of intermodulation distortion. The measured intermodulation would be the total distortion from all factors—the record preamplifier, the recording medium (disc or tape), and the playback preamplifier.

We are interested only in the distortion due to the playback preamplifier itself. To be representative, the output from the preamplifier should consist primarily of the signals at the 4:1 ratio. The signals from the distortion analyzer should be mixed in the proper proportion and fed to the preamplifier so that this ratio will be maintained at the output. Intermodulation components due to the nonlinearity of the preamplifier can then be measured in the conventional manner. The following procedure may be used to perform this test.

1. Feed the lower frequency into the preamplifier. Adjust the level control on the generator to read 1 volt on the output meter.
2. Remove the lower frequency and feed the upper frequency to the preamplifier. This time, adjust the level control on the second generator for an output reading of 0.25 volt.
3. Mix the two signals and feed them together to the preamplifier. Adjust the combined signal-level controls so that the meter at the output indicates the voltages at which the distortion measurement is required.
4. Read the percentage of distortion, as usual.

Although many people measure distortion by simply feeding two signals with a 4:1 ratio to the input of the preamplifier, the above detailed procedure will probably provide a more valid reading.

CHAPTER 8

Power

Force, energy, and power are three common words used in everyday speech. Many times these words are misused by writers and physicists, but the student must always remember that these three terms have distinctly different meanings. Power means the rate at which work is done, the amount of work done, or energy transferred per unit of time. Energy is the capacity for performing work, and force is any physical action capable of moving a body or modifying its motion.

Work has been defined by the product of force × distance, (fd), so that the above definition for power may be expressed by the equation:

$$P = fd/t = f\ (d/t) = fv$$

where

fd is force × distance,
P is power,
t is the time for each cycle,
v is average velocity.

ELECTRIC POWER

No single component of audio equipment has been presented more loosely nor created more confusion than has its power. Four categories of power ratings are frequently found in the various specification sheets. They are continuous sine-wave power, peak-power, Institute of High Fidelity (IHF) music power, and now, dynamic output. The fourth is only a fancy term for music power. In each instance, the required measurement can be made with the setup shown in Fig. 8-1.

The input to the amplifier is fed from a low-distortion source of sinusoidal signal (or square-wave or modulated signal, where required). A resistive load, R, is substituted for the speaker system. The output voltage across the load is measured with an a-c meter. The distortion level is determined from the reading on the distortion meters and observed on the scope. The scope may be calibrated and used to measure voltages.

The power delivered by the amplifier to the resistive load can be calculated from the voltage, E, read on the a-c vtvm. Assuming that the output meter reads rms voltages, the familiar equation:

$$P = E^2/R \qquad (8\text{-}1)$$

will be used. The procedure is self-evident when continuous sine-wave power is to be determined.

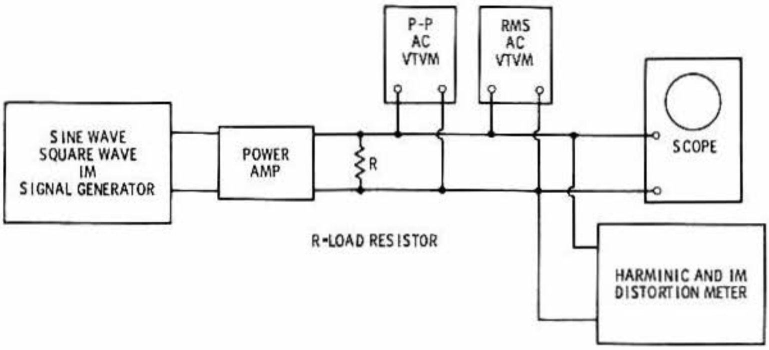


Fig. 8-1. Setup for measuring power input.

SINE-WAVE POWER TO PEAK POWER

The voltage that appears at the output of an amplifier (across the load resistor) takes the same form (exclusive of distortion) as the voltage input to the amplifier. The usual input-signal voltage is in the form of a sine-wave (Fig. 8-2). It is obvious that the amplitude or voltage of this waveform varies through a complete cycle.

Just what voltage represents an average value for the sine wave, to be substituted into equation 8-1 to calculate the average power out-

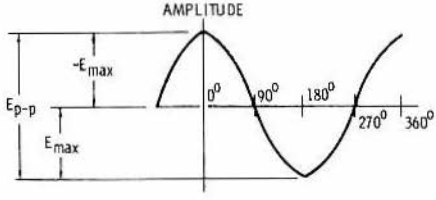


Fig. 8-2. Shape and dimensions of an a-c cosine wave.

put, can be derived mathematically. It is first necessary to set down the equation of the sine wave or cosine wave—depending on the placement of the zero axis or starting point of the wave.

$$e = E_{max} \cos 2\pi ft \qquad (8\text{-}2)$$

where,

e is the actual voltage at any instant of time,
E_{max} is the peak or crest voltage of the cosine wave,
f is the frequency of the cosine-wave,
t is the time for each cycle.

After some minor mathematical manipulations (see Appendix J) we conclude with the well-known expression for the effective or rms voltage:

$$E_{rms} = E_{max}/\sqrt{2} \text{ or } E_{max} = \sqrt{2}\, E_{rms} \qquad (8\text{-}3)$$

Where the above is substituted into equation 8-1, the power in the sine wave becomes:

$$P_{av} = E_{rms}{}^2/R \qquad (8\text{-}4)$$

The sine wave goes through a peak. If this peak were extended over a complete cycle, it would result in a power:

$$P_{peak} = E_{max}{}^2/R = (\sqrt{2}\, E_{rms})^2/R$$

from equation 8-3. Peak power is then equal to:

$$2(E_{rsm})^2/R = 2P_{av} \qquad (8\text{-}5)$$

Thus, if the voltage at the peak of the sine wave were extended for a complete cycle, placing emphasis on *extended for a complete cycle*, the peak power output would be equal to twice the average sine-wave power output.

Equation 8-5 would tend to indicate that all requirements for determining peak power can be met by a simple measurement and calculation: Determine the continuous sine-wave power from measurements and multiply that figure by two. This is exactly the procedure followed by many specifications writers.

SQUARE-WAVE POWER

Amplifiers are not meant to deliver sine-wave power only. Audio amplifiers are required to reproduce the more complicated wave shapes created by speech and music. The ultimate combination of an infinite number of sine waves superimposed on each other is the square wave shown in Fig. 8-3.

The square wave does not vary throughout the complete cycle as extensively as does the sine wave. During each half of the cycle, the square wave behaves as if it were d-c. Since the voltage, E_{max}, remains constant through 180 degrees, the average voltage is obviously E_{max}. Since the same voltage exists in the negative half of the cycle, the effective or rms voltage of a square wave through a complete cycle is E_{max}.

Measuring the square-wave power delivered by an amplifier is analogous to that of the sine-wave example (Fig. 8-1). An audio-frequency square wave is fed into the input of the amplifier. The output is observed on an oscilloscope. The gain is turned up until the maximum output is reached while the rectangular shape is retained. A peak-reading voltmeter is connected across the output load resistor. The power is calculated from:

$$P_{peak} = E_{max}^2/R \qquad (8\text{-}6)$$

where,

E_{max} is the peak voltage.

Most meters are designed to indicate peak-to-peak rather than just one peak. In this case, $E_{p\text{-}p}$ (peak-to-peak voltage) is $2E_{max}$, or $E_{max} = E_{p\text{-}p}/2$. Then the power is shown by the equation:

$$P_{peak} = \left(\frac{E_{p\text{-}p}}{2}\right)^2 \frac{1}{R} = \frac{E_{p\text{-}p}^2}{4R} \qquad (8\text{-}7)$$

The voltmeter and resistance readings are then substituted into equation 8-7. Results can differ considerably from those determined using equation 8-5. Which is valid? The answer will follow the description of the IHF power rating.

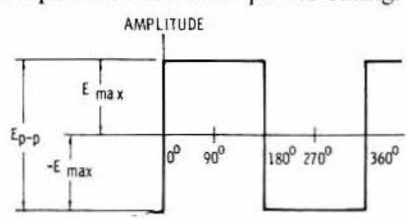


Fig. 8-3. Shape and dimensions of a square wave.

IHF MUSIC POWER AND DYNAMIC OUTPUT

IHF music power is identical to sine-wave power, except that it assumes that the power-supply voltages do not change with signal level. In this measurement, all d-c supply voltages are maintained constant by using regulated external power supplies. Using these supply-voltage conditions, the sine-wave power is measured at a predetermined distortion level, as described above. The IHF power is considerably above the figure determined in the original sine-wave power test. Many manufacturers use this rating.

Peak IHF music power is calculated from the IHF measurement by multiplying the IHF figure by two. This follows the logic of the peak continuous sine-wave power just discussed in equation 8-5. In any of the above measurements, special precautions must be observed if a transistor power amplifier is under test. While tubes can take voltage and power dissipation overload for a short period of time without being destroyed, this is not true with transistors. An instantaneous overload can destroy the power transistor.

Indeed, caution must be observed that the transistor is not overloaded at any time. Double check the transistor characteristics before applying a sine wave which will drive the amplifier to full output. Be even more cautious with the square-wave signal. In either case, do not feed the signal through the amplifier for a longer period of time than is absolutely required to make the test. The transistor may overheat and result in thermal runaway.

Transistor amplifiers are usually transformerless. It has become a practice throughout much of the industry to test for rms power using an 8-ohm load at the output, and for music power using a 4-ohm load. This is due to the characteristic of output transformerless circuits in which more power can be delivered to lower values of load resistors.

Do not use a lower value of impedance or resistance at the output of a transistor amplifier than is recommended by the manufacturer. At high output-signal conditions, the output transistors can be overloaded and destroyed. Ideally, these precautions should not be necessary. But at the present state of the art, they are vital.

SENSE OR NONSENSE

Each type of power rating has its own merits and drawbacks. The discussion below is opinion *only*, although it is based on fact. Opposing opinions can be stated with equal conviction and justice.

Continuous sine-wave power measurements make a lot of sense if the amplifier is to be used to amplify continuous sine-wave signals. But high-fidelity amplifiers are used for instantaneous sounds, such as music and speech.

The d-c supply voltages in class-AB amplifiers change from the quiescent values when the amplifier is delivering its maximum continuous output power. However, with music or speech as a signal source, the voltages do not vary so much, because of a rapidly changing signal level. Power-supply time constants are too high to follow rapid signal variations. It appears to be more realistic to use the IHF music rating rather than the continuous sine-wave output as the standard. It should also be remembered that the "low loading" power amplifiers were based on this premise. No one, as yet, has actually disproved this.

However, music can incorporate sustained notes which will cause a variation in the supply voltages. In that case, the continuous sine-wave power is the most valid measurement. In reality, both figures should be stated. An extreme difference between the two indicates a poor amplifier due to inadequate power-supply regulation. Low-frequency instability, such as motorboating, should be checked in any amplifier where the difference between these two figures seems excessive.

A big difference also indicates that the d-c supply voltages will probably vary considerably with the various instantaneous music levels. At which supply voltages does one make the harmonic-distortion test? In this instance, distortion data would appear to present a distorted view of an amplifier's capabilities.

As for peak power, the value is arrived by multiplying the continuous or IHF power by the figure 2. This number looks impressive on specification sheets, but has little significance unless the data supplied by the manufacturer are divided by a factor of 2.

Measurements made using the square-wave test described are more significant. It not only covers power-supply weaknesses, but encompasses circuit and output-transformer inadequacies or capabilities as well. The above measurements are usually made at 400 Hz or 1000 Hz. High power output at low distortion levels is important at all segments of the audio spectrum between 50 Hz and 10 kHz. Some experts would extend the importance even beyond these limits. However, speaker and ear limitations seem to make this range sufficient for most reproduction requirements. The measurements should be conducted to supply data at these frequencies. A curve should be plotted showing this data.

As a last factor, intermodulation distortion must be considered. Should an amplifier be rated at a predetermined intermodulation-distortion level or at a predetermined harmonic-distortion level? Probably both ratings are important. An intermodulation distortion versus power curve should be supplied with each amplifier. The significance of the intermodulation curves should not be ignored, especially in transistor amplifiers. One common characteristic is that the intermodulation rises at low power levels. This characteristic is much less desirable than high intermodulation at rated power. A typical curve is shown in Fig. 8-4.

Ratings for stereophonic amplifiers are unusually unrealistic. The current practice is to determine the rating of one amplifier (such as the left channel), and multiply this figure by 2 to encompass the two power amplifiers (left plus right channel). Although reasonable for

IHF ratings, this procedure does not include the power-regulation factors important in the continuous power test. To compound matters, the result is frequently multiplied by a second factor of 2 to present an astronomical continuous peak-power rating.

By means of this number, it can be shown that virtually no low-power amplifiers exist (although in reality they are the most prevalent). Manipulation of numbers can easily raise a dual 21-watt stereo amplifier to a rating of 80 watts. This can be done as follows:

A 12-watt amplifier can easily have an 18-watt music power rating. Two such amplifiers on one chassis means that 36 watts of IHF power is available. Multiply this by 2 and you now have a 72-watt peak IHF power amplifier. How come 80 watts? After all this, is it so terrible to cheat a tiny bit for a measly 8 watts? And if your line voltage is high, you're not really cheating.

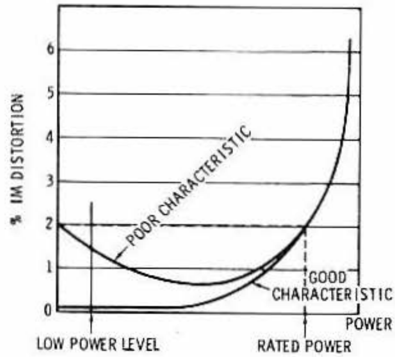


Fig. 8-4. Typical intermodulation-distortion curve.

SOME METHOD OUT OF CHAOS

An accurate method of stating power can be derived from two sets of power curves showing the rms power output over the entire range from 20 Hz to 20 kHz.

First, plot the power output for a constant voltage input throughout the complete audible range. Do this at the rated power output, 3 db below rated output, 6 db below rated output, and at the ½-watt output level. This shows the complete power and frequency response of the amplifier.

In a second series of curves, plot the power output that can be obtained at several harmonic-distortion levels between 0.1 percent and 2 percent. These two sets of curves, when studied carefully, can reveal much more about an amplifier's power output than any meaningless astronomical figures. The curves should become part of the standard specifications supplied by manufacturers.

CHAPTER 9

Sensitivity, Overload, Hum, and Noise

Each of the four factors listed in the title seems to be an independent item. Careful consideration will show their close relationship and interdependence.

SENSITIVITY

Sensitivity of an amplifier determines how much signal (voltage or power) must be fed to it so that it can deliver its rated output. As an example, consider a signal fed to the tuner input on a 20-watt amplifier. If it is necessary to feed an input signal of 0.3 volt in order for this amplifier to deliver 20 watts, then the amplifier has a sensitivity of 0.3 volt for 20 watts of output. A setup for measuring this sensitivity is shown in Fig. 9-1. The procedure is as follows:

First, feed a 1000-Hz signal from an audio signal generator to the amplifier. Set all controls on the amplifier for the most linear response at the output while the volume control is at maximum. Adjust the output from the signal generator so that 20 watts is developed across the load resistor, R_L, at the output of the amplifier. The output power can obviously be determined by dividing the load resistance into the square of the voltage read on the a-c output meter. Now read the voltage on the input meter. This is the sensitivity of the amplifier for 20 watts of output power. This same procedure can be repeated to check the magnetic phonograph, tape head, microphone, etc., sensitivities of the amplifier.

The sensitivity is specified at the rated power output of the amplifier. This does not permit you to compare directly the sensitivities of two units with different power ratings, and it does not give you one output level on which to base all measurements. When the specific sensitivity of an amplifier is stated, it is necessary to indicate the reference used. As yet, no standard output for use as a reference has been established.

In order for two amplifiers with different output capacities to be compared, their inputs must be measured when they are each delivering identical outputs. This information can be calculated from the given data. If the smaller amplifier delivers an output power P_1 for an input of v_1 volts, it is only necessary to find the input v_x volts required from the second (and larger) amplifier to deliver the same P_1 watts of power.

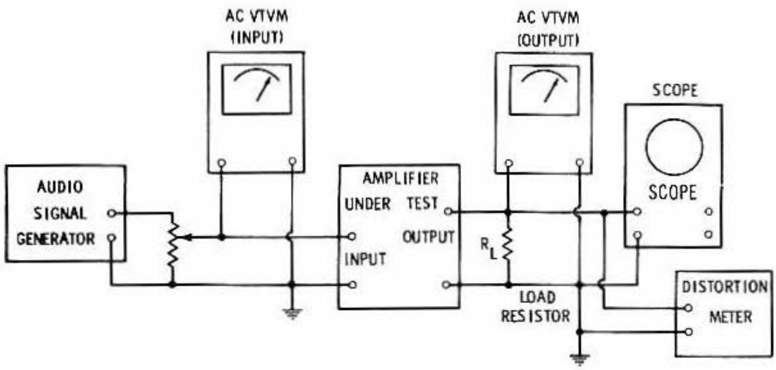


Fig. 9-1. Measuring the sensitivity of an audio power amplifier.

Assume that the larger amplifier is rated at v_2 volts of input for P_2 watts of output. The desired information is the required voltage input, v_x, for an output of P_1 watts from the larger amplifier. It may be calculated from the formula:

$$v_x = v_2 \left(\frac{P_1}{P_2}\right)^{1/2} \qquad (9\text{-}1)$$

The derivation of this can be found in appendix K. The comparison between v_x and v_1 will let you determine which amplifier has the higher sensitivity. The lower number indicates the amplifier which has more gain and higher sensitivity.

As as example, consider two hypothetical amplifiers. One requires an input of 0.2 volt for an output of 10 watts. The second requires an input of 0.6 volt for an output of 40 watts. Which is more sensitive?

From the equation 9-1:

$$v_x = v_2 \left(\frac{P_1}{P_2}\right)^{1/2} = .6\left(\frac{10}{40}\right)^{1/2} = .3 \text{ volt}$$

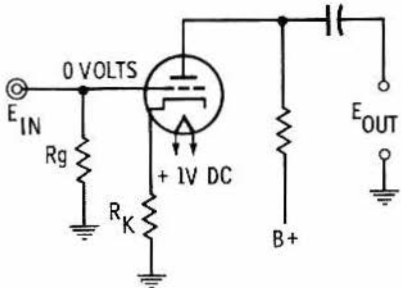
The 40-watt amplifier requires 0.3 volts for 10 watts of output, while the 10-watt unit requires only 0.2 volt. The 10-watt unit is therefore more sensitive than the 40-watt amplifier.

OVERLOAD—MAXIMUM INPUT SIGNAL

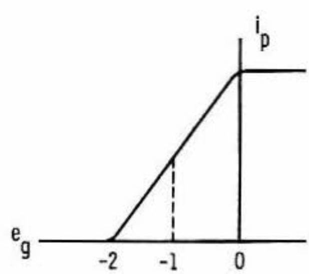
This factor is seldom given in a set of specifications. Yet, it is as important as any other factor limiting an amplifier's ability to perform. Every vacuum tube and transistor has its limits of operation. Consider the circuit in Fig. 9-2 A, and the typical characteristic curve in Fig. 9-2B. The grid is biased to −1 volt because of the cathode resistor and plate current. The grid voltage can swing only as far negative as −2 volts before plate current will cease to flow and the tube is in the cutoff region. It can also swing in the positive direction to "0" before a second limiting region, saturated plate current, is reached. This swing can be caused by an input signal, E_{in}, at the grid, superimposed on the d-c bias voltage. If the voltage swing is outside the 0 to −2-volt region, the tube will distort the signal. The grid circuit will be overloaded. This overload voltage is a limit on the amplifier's ability to accept an input signal.

The placement of the volume control determines the overload characteristic of the overall amplifier. If the volume control is placed before the first amplifier stage, the amount of input signal reaching the grid circuit at the amplifier is limited by the setting of the control. Very large signals can be fed to the input, but only a small signal will be present at the grid. The amount of signal the circuit will accept is practically limitless. A strong signal cannot overload the circuit.

Now suppose the volume control follows the vacuum-tube circuit. There is nothing to limit the size of the signal between the grid and cathode. A signal greater than the +1-volt peak at the input will overload the amplifier tube. The rms voltage swing this amplifier will accept before overload is $1/(2)^{1/2}$.

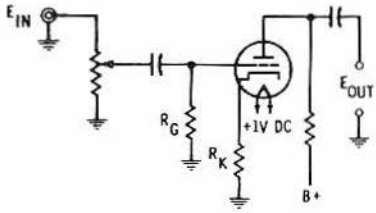


(A) Tube circuit.

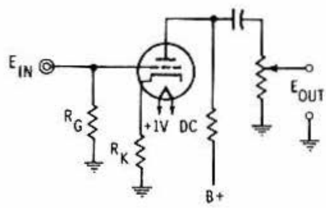


(B) Vacuum-tube characteristics.

Fig. 9-2. A typical vacuum-tube circuit and tube characteristics.



(A) Before amplifier tube.



(B) After amplifier tube.

Fig. 9-3. Two possible locations for a volume control in a circuit.

To check the overload ability of an amplifier, use the circuit in Fig. 9-1. First, with the volume control fully up, measure the distortion at the 1-watt or 1-volt output level. (Control the amount of signal fed to the amplifier at the signal generator.) Then turn the volume control down so that the output is reduced 1 or 2 db. Increase the output signal from the signal generator until you once again have 1 watt or 1 volt at the output. Measure the distortion again. If it has increased, you have passed the capability of the amplifier to accept signal. If distortion remains constant, repeat the procedure until the distortion does begin to rise. At this voltage, you have reached the amplifier's limits to accept input signals.

Should the volume control follow the amplifying device as in Fig. 9-3B, these limits can be easily reached. If it precedes the tube or transistor, the limit of your signal generator to produce signal will be reached before the amplifier's limits are reached. With an ideal volume control (zero resistance between the wiper on the control and ground), the limit will never be reached for the circuit in Fig. 9-3A.

The importance of overload cannot be overemphasized. Because of several considerations, magnetic phonograph, tape head, and microphone preamplifiers must be placed before the volume control. Unless the preamplifiers can accept the peak signals the transducers can produce, the integrated preamplifier–power-amplifier combination will deliver a distorted output, even though the power-amplifier section has not reached its own limits in delivering power.

Although the overload problem can be severe in vacuum-tube amplifiers, its serious consequences are magnified in transistor circuitry. Because of low collector supply voltages (when compared to plate supply voltages), the undistorted signal-swing capability is small. The tendency to overload is apparent much sooner than in the vacuum-tube example. In a good design, only a very small signal from a magnetic phonograph cartridge, tape head, or microphone preamplifier will precede the volume control. While a cathode follower can precede the level control in vacuum-tube units, the emitter-follower must be fully checked before being placed in this position.

HUM AND NOISE

Most types of undesirable audio interference can be classified under the category of either hum or noise. Power-line frequencies or their harmonics may appear at the output of an amplifier.

Because of its nature, hum is more prevalent in vacuum-tube than in transistorized preamplifiers. Assuming a well-filtered power supply, hum in vacuum-tube units is usually induced into the signal circuit through direct radiation from the power transformer, from filament leads running near grid circuits, and from heater-cathode leakage. Under similar conditions, with an extremely well-filtered power supply, hum can be induced into the signal circuitry of transistor units only by direct radiation from the power transformer. This assumes that normal precautions have been taken to dress all current-carrying leads sensibly.

Transistor as well as vacuum-tube circuits are subject to noise. Noise is a more random interference than is hum. It does not follow a set, repetitive pattern. It appears as hiss at the speakers.

Either type of circuit produces noise because the components in the circuit generate noise due to random electron motion. The noise voltage at the terminals of a resistor is proportional to the square root of the temperature, resistance, and bandwidth under consideration.

In triode tubes, the types primarily used in audio amplifiers, two types of noise are most prevalent. The first, known as shot noise, is caused by the random rate of emission of electrons from the cathode. The noise here is proportional to the square root of the d-c plate current and the amplifier's bandwidth.

The second major cause of tube noise can be traced to ions formed in a tube by the collision of the traveling electrons and any residual gas left after the "vacuum" has been formed. These ions cause grid current, with the associated shot noise, and also cause variations in space charge.

Transistor noises can be traced to three primary factors. First, there is the famous 1/f semiconductor noise at low frequencies. This states that noise generated by the transistor material is inversely proportional to the particular section of the frequency spectrum under consideration. Thus, the components of noise at 60 cycles due to the semiconductor material is double the noise components at 120 cycles due to the material. Some experts feel that the importance of this has been exaggerated out of proportion.

The second cause of noise is the random motion of conduction particles from the emitter to the collector and back. This is the shot effect in transistors. There is also the thermal noise due to base resistance. It has been shown experimentally and theoretically that noise is directly dependent on the size of the resistance of the signal source.

Transistors used for audio equipment are usually pretested for noise and given a noise-figure rating. The noise figure, F, is:

$$F = \frac{\text{Signal-to-noise ratio at the transistor input}}{\text{Signal-to-noise ratio at the transistor output}} \qquad (9\text{-}2)$$

$$= \frac{1}{\text{Gain}} \times \frac{\text{Noise at the transistor output}}{\text{Noise at the transistor input}} \qquad (9\text{-}3)$$

This ratio is usually checked for a 1-cycle bandwidth at 1000 Hz and is considered valid because it is within the 1/f region. It is known as the spot noise. A more reasonable noise figure would consider the entire band used in the amplifier. However, experience has shown that the spot-noise figure correlates well with the results obtained experimentally.

Feedback is usually prescribed as a cure-all. It reduces distortion from an amplifier. It will reduce the noise (theoretically, at least) by the same amount that it reduces the signal. Thus, the all-important signal-to-noise ratio will remain unchanged. If the feedback is taken over several stages, the ratio may be increased because noise generated in later stages will be fed back to the earlier tubes or transistors in the circuit, where it can be further amplified.

Noise should be reduced if feedback is taken around a noisy stage. However, the noise is of such random nature that the feedback network may not respond fast enough to the noise impulses. If it does reduce noise, the reduction will probably not be as great as calculated from the feedback factor, $1 - A\beta$:
where,

A is the gain of the amplifier without feedback,
β is the ratio of the feedback signal to the output signal.

Hum and noise were considered here under one heading because practical measurements usually consider both at one time rather than as two individual elements. The following steps can be used to check the total hum and noise output from an amplifier. Use the setup shown in Fig. 9-1.

1. Remove all input signals from the amplifier.
2. Put a resistor load across each input. These resistors should match the output impedance of the components feeding the amplifier. The impedance of a magnetic cartridge or tape head is several thousand ohms, the impedance of a ceramic cartridge and crystal microphone is several hundred thousand ohms, and the output impedance from the cathode follower in a tuner is usually somewhat above 500 ohms. The magnetic cartridge or tape-head inputs can be properly loaded with a 3300-ohm re-

sistor, the ceramic or crystal-transducer inputs with a 470,000-ohm resistor, and the tuner input with a 1000-ohm resistor. To make the test on a tuner input more stringent, a 10,000-ohm resistor can be used. This will take into account the units that do not use cathode or emitter followers at the outputs.

As with the vacuum-tube units, the inputs to transistor amplifiers should be loaded when hum and noise tests are made. Although the output impedance of a transistor tuner using an emitter follower is lower than the output impedance of a vaccum-tube unit, the recommended input load resistors are the same as those used when testing vacuum-tube amplifiers. This will take into account the many vacuum-tube tuners used with transistor amplifiers. The only exception to the rule is the microphone input.

Transistorized equipment will seldom accommodate crystal or ceramic microphones. Lower-impedance dynamic microphones are more likely to be used. If a low-impedance dynamic microphone is to be specified for use with an amplifier, load the input with a 470-ohm resistor; if a high-impedance type is specified, use a 27,000-ohm resistor at the input. In either case, the resistors can be chosen arbitrarily. The values itemized above will simulate the worst possible input conditions.

3. Set all controls so that the output from the amplifier is flat. Turn the level control to minimum, and note the output voltage. Refer to this as V_R, the residual hum and noise voltage at the output.
4. Turn the level control up for maximum output. Set the input selector to one of the inputs—let us say the tuner—and note the output voltage. This is V_T, the noise and hum voltage from the tuner channel in the amplifier.
5. Set the input selector switch to each of the other inputs on the amplifier, and note each voltage output. Assume for this discussion that the phono is the only input left. The hum and noise output voltage here is V_p.
6. Calculate the output voltage at the rated power level from the formula:

$$\text{output voltage } V_o = (\text{power} \times \text{load resistance})^{1/2} \qquad (9\text{-}4)$$

(NOTE: In a separate preamplifier, V_o is the rated preamplifier output voltage. It is the voltage developed across a 1,000,000-ohm load resistance in parallel with a 1000-pf capacitor. This load should be at the output of a preamplifier when all tests are made.

7. Calculate the ratio of the output voltage to the residual hum and noise, using the formula:

$$\text{residual hum} = (20 \log V_o/V_R) \text{ db} \qquad (9\text{-}5)$$

8. Calculate the noise and hum ratio in the tuner channel, using the formula:

$$\text{tuner hum} = (20 \log V_o/V_T)\text{db} \qquad (9\text{-}6)$$

9. Calculate the phono noise and hum ratio from the formula:

$$\text{phono hum} = (20 \log V_o/V_P)\text{db} \qquad (9\text{-}7)$$

To avoid the calculations necessary in steps 7 through 9, the curve in Fig. 9-4 may be used. Just determine the ratio of V_o to the hum and noise voltage measured in steps 3, 4, or 5. Refer to any of the hum and noise voltages as V_H. Read the relative hum in db from the curve.

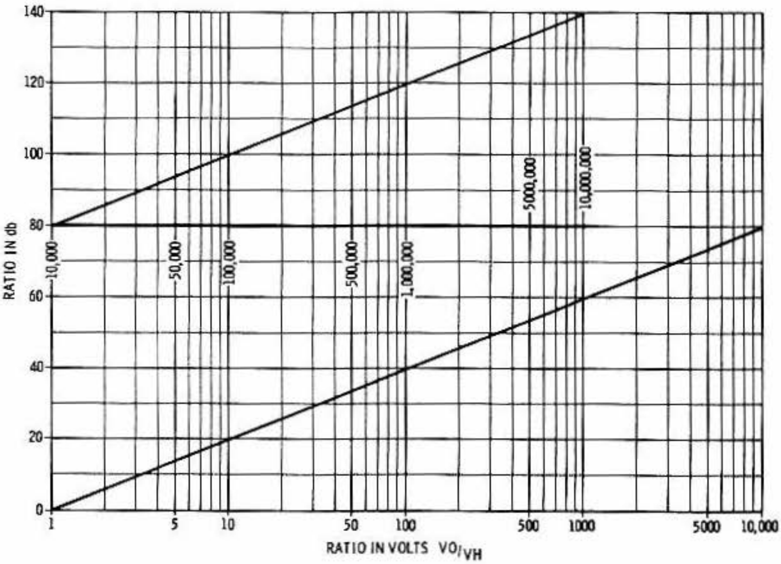


Fig. 9-4. Graph showing voltage ratio in terms of db.

Noise measurements discussed here have considered the overall bandwidth of the amplifier. Other types of measurements take the characteristics of the human ear into account, in accordance with the equal-loudness curves. The noise is measured after the frequency response (after the output of the amplifier) is shaped to match some low-level equal-loudness curve. These frequency characteristics are built into some meters, in accordance with ASA standards. However, it should be noted that these measurements include both the amplifier noise and supposed human-ear response to noise. To be fully accurate, the test should also consider the speaker and input-source characteristics. Because all extraneous factors cannot be accounted for in the measurement, it is probably best to measure the amplifier's unweighted noise characteristic.

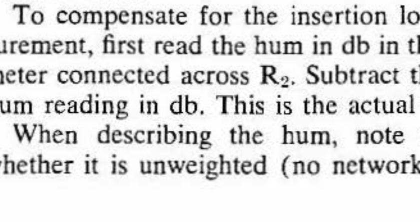
Weighted measurements can be established using one of the limited bandwidths set as a standard by the industry. Filters can be used to set bandwidth limits from 20 Hz to 10 kHz, or 160 Hz to 10 kHz, or 280 Hz to 10 kHz. The gain at these frequencies is 3 db down from the maximum output. Outside these limits, the response keeps dropping at the rate of 6 db per octave. The frequency limits the manufacturer uses should be stated in the specification so that bench tests can be made to these standards. Single-section r-c filters at the amplifier output can be used to set the required bandwidth. A typical network is shown in Fig. 9-5.

In the equation, $1/2\pi R_1C_1$, there is a gain loss of 3 db in the upper frequency. The values of the capacitors can only be approximated from the formula. The response should be checked with a single generator to correct for any inaccuracies in the formula and component values. Typical values for $R_1 - C_1$ and $R_2 - C_2$ are shown in the figure. In any weighted test, R_1 and C_1 remain constant. Only R_2 and C_2 change, depending on the desired frequency where the low end of the band should start to roll off.

The insertion loss is the amount the signal at midfrequencies is reduced by the network. It may be determined as follows: Feed a 1000-Hz signal to the amplifier, and read the output across R_L in db. Next, read the output across R_2 in db. The difference in db readings is the insertion loss.

To compensate for the insertion loss when making the hum measurement, first read the hum in db in the normal fashion on the output meter connected across R_2. Subtract the insertion loss in db from the hum reading in db. This is the actual weighted hum of the amplifier.

When describing the hum, note the frequency limits. Indicate whether it is unweighted (no networks and no insertion loss to take



TYPICAL COMPONENT VALUES

I. HIGH-FREQUENCY ROLL-OFF AT 10 kHz
$R_1 = 4700\ \Omega$
$C_1 = 3300$ pf

II. LOW FREQUENCY ROLL-OFF
A. at 20 Hz
$R_2 = 100K$
$C_2 = .08\ \mu f$

B. at 160 Hz
$R_2 = 100K$
$C_2 = .01\ \mu f$

C. at 280 Hz
$R_2 = 100K$
$C_2 = .006\ \mu f$

Fig. 9-5. Circuit for measuring weighted noise of amplifier by varying bandwidth.

into account), or whether it is weighted for frequency limits of 20 Hz to 10 kHz, 160 Hz to 10 kHz, or 280 Hz to 10 kHz.

Until now, only hum and noise created in amplifiers have been discussed. Hum and noise can also be due to external causes, such as proximity to power lines, motors, arcing switches, etc. Line filters and shields are usually used to minimize some of the effects from these disturbing factors. Complete elimination can only be accomplished by removing the sources of interference.

INTERRELATIONSHIP

The measurements discussed here and in previous chapters are all related to each other. When measuring hum, some reference level must be used. What should the distortion be at the reference level? At what frequency should this measurement of distortion be made? Should it be intermodulation or harmonic distortion which determines the level? Which intermodulation method of measurement should be used?

Although each was treated individually, no one measurement is independent of the other. Considerations in design show how closely the four topics discussed here are interrelated. Start with hum and noise. No attempt was made to separate the two for measurement, even though there are methods of accomplishing this. The separation of these two factors is essential only in design. Even here, only the relative order of magnitude is most important. This required information can usually be obtained by observations on an oscillograph screen.

Hum and overload are very closely related in design considerations. The engineer who decided to place the level control after the amplifying device in Fig. 9-3B, rather than before it as in Fig. 9-3A, was not all wrong. Residual hum is certainly better in the circuit shown in Fig. 9-3B than in the circuit in Fig. 9-3A, because the tube is out of the circuit in Fig. 9-3B when the level control is set at minimum for the measurement. The designer had to weigh the overload problem against the hum to decide that the distortion created by overload was not as obvious or as objectionable as the interference created by the hum and noise. This philosophy satisfies many people—especially those with efficient or peaked speaker systems in which use of the circuit in Fig. 9-3A would have caused an audible hiss at all settings of the level control. It would not satisfy a more critical listener, who would rather suffer with more noise than be subjected to additional overload distortion. Ideal amplifiers will be designed only when compromises of this type will become unnecessary.

Sensitivity is also related to hum and noise. It is not unusual to find that the higher-gain amplifiers produce more hum. Sensitivity is frequently minimized to reduce hum. Considering tuners and magnetic phonograph cartridges in use today, it is reasonable to say that satisfactory amplifiers can be driven to full output with a phono input of 10 mv and a tuner input of 0.5 volt.

Let us say you wish to compare the relative hum levels of two amplifiers with identical power ratings. Amplifier A has a sensitivity of 5 mv for full output, and amplifier B has an equivalent sensitivity of 10 mv. Either one will accept the cartridges currently in use.

The hypothetical amplifiers may also have hum specifications reading something like this:

> Amplifier A: Phono hum is 45 db below rated output.
> Amplifier B: Phono hum is 49 db below rated output.

Using this information, which amplifier has less hum for identical sensitivities? Or stated otherwise, if identical cartridges are used with both amplifiers, which amplifier will have the better signal-to-noise ratio?

To determine this information, first get the ratio of the higher to the lower of the two sensitivities. The ratio of the sensitivity of amplifier B to that of amplifier A is 10 mv to 5 mv, or 2. This figure, 2, in Fig. 9-4 indicates a difference of 6 db. Thus, 6 db must be subtracted from the hum figure of the less sensitive amplifier (B) to compare it directly with the signal-to-noise ratio of the more sensitive amplifier (A). Thus, if a 5-mv signal were fed to amplifier B, the hum would be 49 db − 6 db, or 43 db below the output. Consequently, amplifier A would have less hum than amplifier B if both inputs were fed signals from the same source. The former would have the better signal-to-noise ratio.

The above compares hum and noise for amplifiers with different rated output powers. The hum and noise figure is closely related to output power when the rating of hum and noise is stated as a specific number of db below the rated power. Assume that a 10-watt amplifier has a hum level of 40 db below the rated power. Furthermore, assume that a 50-watt amplifier has a hum level of 50 db below the rated power. The problem is to find which amplifier is delivering more hum signal to the speaker. The actual hum voltage at the output of both amplifiers must be calculated and compared with each other.

First, let us calculate the hum voltage that would be present at the output of the 10-watt unit. The amplifier impresses 12.6 volts across the 16-ohm load (from equation 1) for 10 watts of output. From Fig. 9-14, the hum voltage for 40 db below rated output is:

$$V_o/V_H = 100 = 12.6/V_H$$

where,

V_o is the rated preamplifier output voltage,
V_H is any hum or noise voltage.

$$V_o/V_H = 100 \text{ for a 40-db hum-level ratio}$$

$$V_H \text{ for the 10-watt amplifier is } 12.6 \times 10^{-2} \text{ volts}$$

The voltage at the 16-ohm output, due to hum in the 50-watt amplifier, can be calculated in the identical manner. The amplifier impresses 28.2 volts across a 16-ohm load for 50 watts of output, and an output of 50 db below 50 watts is a voltage ratio of $V_o/V_H = 316 = 28.2/V_H$ (see Fig. 9-4). Then V_H, the hum voltage for the 50-watt amplifier is 8.9×10^{-2} volts. Thus, the 50-watt amplifier will deliver less hum voltage than the 10-watt unit.

The difference in hum in db between the two units can be found from the chart in Fig. 9-4. The voltage ratio is $12.6 \times 10^{-2}/8.9 \times 10^{-2}$, or 1.42. From Fig. 9-4, the ratio represents about 3 db. Thus, the 10-watt amplifier has 3 db more hum than the 50-watt unit.

This information can be obtained from measurements using the setup in Fig. 9-1 and the following procedure.

1. Set the level control on the amplifier to maximum, and feed a signal into the 10-watt amplifier. Note the reading on the meter.
2. Remove the signal and substitute an equivalent input load resistor for the signal. Note this hum-voltage reading on the meter.
3. Calculate the ratio of the output voltage in step 1 to the hum voltage in step 2, and determine the ratio in db from Fig. 9-4.
4. Substitute the 50-watt amplifier into the setup. Feed the identical signal as in step 1 to this larger amplifier. Now adjust the level controls on this amplifier for an output identical to that in step 1.
5. Remove the signal from this amplifier without upsetting the level-control setting. Substitute the resistor at the input as in step 2. Note the reading on the output meter.
6. Calculate the ratio of the output voltage in step 4 to the hum voltage in step 5. Determine the ratio in db from Fig. 9-4.
7. Compare the db ratio found in step 3 with that found in step 6 to determine which amplifier has the better hum characteristic.

The relationship between hum and power rating exists. The two methods outlined can be used to find the relative merit of the amplifiers. The intricate calculations or measurements are required because no industry-wide standards have been established.

Standards should be established for specifying hum and noise at standard levels below specified power and for specific voltage inputs with standard input and output loads. Until these measures are taken, calculations similar to those illustrated here will be required to enable one to compare the qualities of two units. This is even more important in output-transformerless transistorized power amplifiers, where the hum ratios will vary with the output load.

CHAPTER 10

Sundry Tests and Measurements

Any factor or number of factors may affect the listening quality of an amplifier. Distortion, frequency response, and the other items discussed in previous chapters certainly are significant characteristics. Other attributes such as damping factor, phase shift, parasitic oscillations, etc., also affect the particular sound qualities a specific amplifier is capable of producing. Just to what degree any characteristic is important in the final audible output is academic.

Seven characteristic tests and measurements that can be performed on an audio amplifier are discussed and described in this chapter. These, in conjunction with the measurements discussed in previous chapters, are a compilation of the most common and probably the most important tests to be performed on an amplifier in determining its quality.

PHASE-SHIFT MEASUREMENT

Any amplifier stage will normally shift the phase 180 degrees between the input and output. The relationship in the voltage amplifier is shown in Fig. 10-1. There would be no problem if all frequencies in the audible range were shifted this 180 degrees. The relative phases of all amplified components would then be the same at the output as they are at the input to the amplifying device. This is known as *linear phase shift*.

Because of the capacitive and inductive characteristic of the components used in an amplifier, not all frequencies are shifted by the same amount. This lack of uniform phase shift at all frequencies is known as *delay* or *phase-shift distortion*.

Phase-shift distortion can be observed on a complex wave. Suppose there is a signal composed of a fundamental with third-harmonic con-

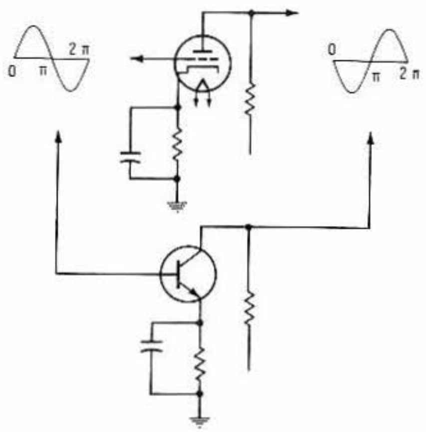
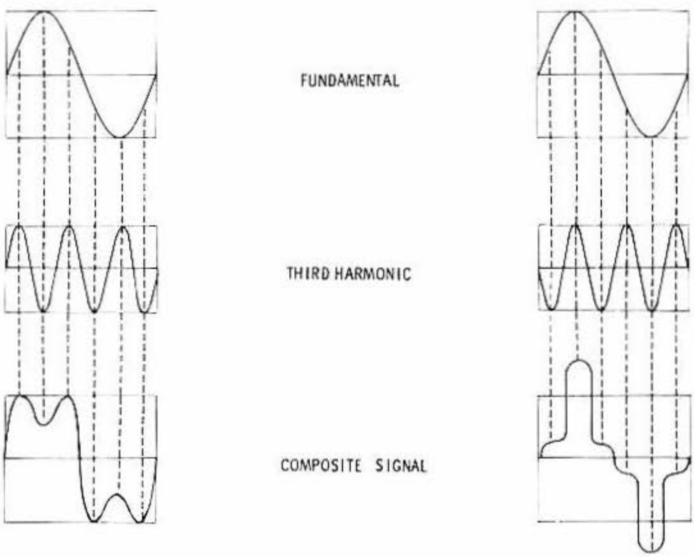


Fig. 10-1. Phase relation between input and output of a single audio-amplifier stage.

tent, as shown in Fig. 10-2A. The shape of the composite signal would be shown. If this signal were fed through an amplifier where the phase of the third harmonic was shifted 180 degrees with respect to that of the fundamental, the shape of the composite signal would be as shown in Fig. 10-2B. The composite waveshape would have



(A) Signal composed of a fundamental with third harmonic.

(B) Signal shifted 180 degrees with respect to the fundamental.

Fig. 10-2. Equal signal with equal amounts of third-harmonic content.

remained unchanged had both frequencies been shifted in equal multiples of 180 degrees.

Does the composite signal in Fig. 10-2B sound the same as the one in Fig. 10-2A? The fundamental and third harmonic are of the same amplitude, but are only shifted in phase with respect to each other. Some people say they hear the difference. Could the audible difference actually be due to a harmonic-distortion difference because the greater amplitude of the signal in Fig. 10-2B drives the amplifier harder than the composite signal in Fig. 10-2A? Or do they really hear the phase distortion?

Delay distortion is not of much importance in an audio type amplifier, since delay distortion is not perceptible to the ear. The exact phase relations between the various components of a complex sound are unimportant unless the phase shift is great enough to produce an appreciable time-delay distortion, such as 0.05 second. Phase shift is unimportant unless it is great enough to produce a time delay greater than 8 milliseconds at high frequencies and more than 15 milliseconds at frequencies below 100 cps.

As the years progressed, more and more emphasis was placed on the significance of phase distortion. It would seem that there are definite audible differences. The significance of phase shift increases in stereo listening because an important portion of the stereo effect is attributed to the relative phase of the signals from the two channels.

The circuit in Fig. 10-3 may be used to measure the phase shift of an audio amplifier at various frequencies. Several precautions must be taken when making this measurement. The phase shifts of the vertical and horizontal scope amplifiers must be identical. Otherwise, phase-correction networks must be added externally. The amplification setting on the scope must be the same for the "x" and "y" axis, so that there will be an equal number of phase-shifting stages or networks in both deflection circuits. To equalize the signal levels fed to the two scope amplifiers, two attenuator networks (R1-R2 and R3-R4) have been added to the circuit.

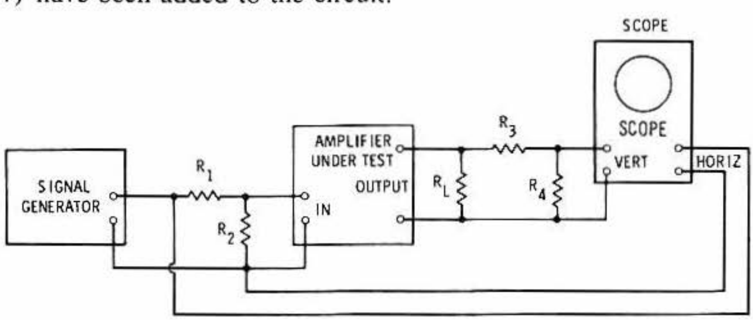


Fig. 10-3. Setup for measuring phase shift on a scope using Lissajous patterns.

The resistors in the attenuator networks must be *carefully chosen* so as not to introduce significant error. They should all be as small as possible, while R3 + R4 should be not less than 100 times the resistance of load resistor R_L. All resistors should be of the noninductive type. The resistors, along with the leads, should present a minimum capacity to the circuit.

In determining the phase shift using this setup, the signal fed to the horizontal "x" axis is considered at zero phase. The phase of the signal driving the vertical "y" axis is measured with respect to the phase of the "x"-axis signal. The formula for determining the phase shift from the scope pattern is derived in Appendix L.

The relative phase angle between the two signals—one applied to the vertical plate and the other to the horizontal plate—can readily be found from the pattern on the scope.

First, fit the pattern into a square box on the screen of the scope tube, as shown in Fig. 10-4. Next measure the lengths marked "x" and "r" from the center of the pattern. Substitute these values into the formula:

$$\sin \theta = \frac{x}{r} \qquad (10\text{-}1)$$

or,

$$\theta = \arcsin \frac{x}{r} \qquad (10\text{-}2)$$

where,

θ is the phase difference between the two signals.

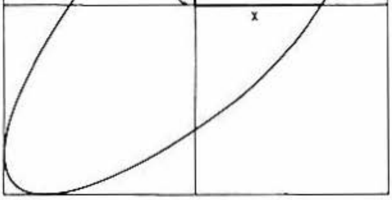


Fig. 10-4. Phase shift described by an ellipse in a square on the scope screen.

INSTABILITY

Whereas phase shift can be measured accurately using the method described in this chapter or in the one on special instruments, there is no mathematical number describing a quantity of instability. Although it is not readily measurable, we can go one step further—no form of oscillation should exist in an audio amplifier.

Parasitics can be defined as undesirable types of oscillation due to stray tank circuits in an amplifier. The frequency of oscillation is usually above the audible range. Nevertheless, there are many good reasons for eliminating this frequency, along with all other types of supersonic and subsonic oscillatory frequencies. The frequency of oscillation:

1. May be strong enough to blow a tweeter speaker.
2. Absorbs power that may be put to useful purposes.
3. May cause excess voltage to appear across a component.
4. Will cause distortion, which can appear as a buzz or unclean sound in the speaker.

The circuit in Fig. 9-1 may be used to check for parasitics in an amplifier. Turn all controls on the amplifier up to maximum. Use a 30-Hz tone. Slowly vary the strength of the signal fed to the amplifier. Observe the scope to note that no pockets of oscillation (Fig. 10-5) should appear on the signal. Repeat this with load R_L removed and also with R_L reduced to about 1 ohm. (Note: Do not reduce R_L below its rated value when testing a transistor amplifier, so that the output transistors will not be overloaded under signal conditions.)

Repeat the above for various settings of the tone and level controls. Do not overload the amplifier for more than a few seconds. Before performing this test on a transistor amplifier, check with the manufacturer to see whether the circuit will withstand this type of overload.

There should be no sign of oscillation under any of the conditions described above. If the amplifier you test has pockets of oscillation, they do not mean that your amplifier has a manufacturing defect.

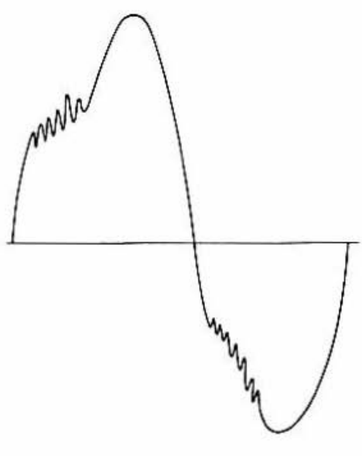


Fig. 10-5. Example of how parasites may appear as pockets of oscillation on a sine wave.

Some designs have not been made completely free from parasitics or other forms of oscillation or tendencies to oscillate.

Instability can also be made evident when the amplifier is driven from a 5000-Hz square wave at a relatively low power output level. Feed the square wave to an auxiliary input on the amplifier. Adjust the generator so that the output from the amplifier is about one-fourth the maximum the unit can deliver. Next, vary the load across the

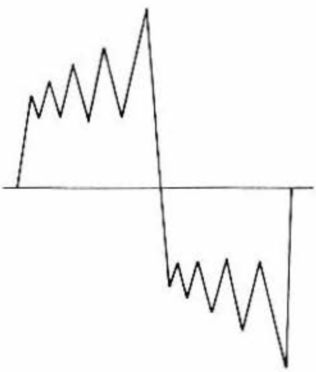


Fig. 10-6. Example of amplifier with tendency to oscillate. Notice the increasing amplitude of ringing on a square wave.

output for the amplifier from 1 ohm to an open circuit. Load the amplifier with different capacitors ranging from 100 pf to 10 mfd. Nonpolarized electrolytics can be used for the higher capacitor values. If nonpolarized capacitors are not available, use two identical polarized capacitors in series, connected positive to positive lead or negative to negative lead. The total capacity is one-half the capacity of either unit.

Finally, load the amplifier output with inductors ranging from 10μh to 1 h. During any of these tests no sine wave, nor damped sine wave, should appear riding on the top and/or bottom of the square wave. If the sinusoidal excursions are constant or rise with time, as shown in Fig. 10-6 the amplifier is either unstable or has a tendency toward instability. The amplifier is also unstable if it breaks into complete oscillation with any of these loads. The total curve should not bounce around on the screen of the scope at any time. If it does, this is a sign of low-frequency instability.

DAMPING FACTOR

A major factor in clarity of reproduction is definition. Definition is a measure of the capability of a system to separate two successive signal pulses. For good definition, a speaker will reproduce a signal only for the duration of the signal. Some speaker cones may continue to vibrate after the signal has been removed. Consequently, a musical note will continue to be produced even though no signal is present. This is known as overhang.

Should there be no or very little overhang, the speaker is said to be well damped. That is, once it is set in vibration, it will stop moving the moment the source of the vibration is removed. A low impedance across the speaker leads is a strong positive factor in helping the speaker overcome its overhang characteristic. Good-quality high-fidelity amplifiers are designed to present a low impedance to the speakers. If the amplifier has an impedance of about one-fourth that of the speaker system or less, the speaker is as well damped as it can be.

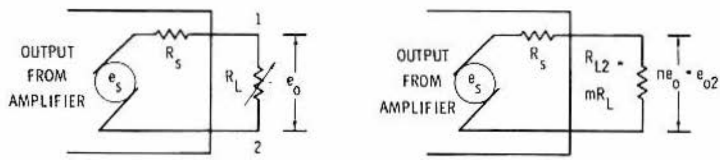


Fig. 10-7. Two setups used to derive a formula for damping factor.

The ratio of the speaker impedance, R_L, to the impedance the speaker sees in the amplifier, R_s, is known as the damping factor. Any amplifier with a damping factor of four more over the entire audio-frequency band is satisfactory. With the aid of Fig. 10-7, the formula for determining damping factor may be derived. The test setup is that of Fig. 9-1, with R_L being variable. The output voltage, e_s, should be limited to a maximum of about 1 volt as measured on the a-c vtvm output instrument.

In Fig. 10-7, e_s is the output voltage from the amplifier, R_s is the output impedance of the amplifier, and R_L is the load (representing the speaker or power resistor, as required). Using this circuit, there are several methods for measuring the damping factor.

In one method, first measure the voltage between points 1 and 2 (the output terminals of the amplifier) with the load, R_L, removed. Call this voltage V_{NL}. Now, insert a variable R_L into the circuit, as shown, and vary it until the output voltage is ½ V_{NL}. Under this condition, R_L is equal to R_s. Measure R_L, which is now equal to the internal resistance of the amplifier. The damping factor is the rated output impedance of the amplifier (usually 4, 8, or 16 ohms) divided by the measured R_L.

Two precautions: First, observe the signal on an oscilloscope when making all measurements. If there is any distortion, decrease the output level and repeat the test. Second, do not use this test on transistorized power amplifiers; the low-valued output load resistors may damage the transistors.

Although this method is simple, it has one major disadvantage. Because R_L is so small when it is equal to R_s, the characteristics of the circuit we are trying to evaluate will frequently change under test. The resistance of R_L will be only a vague indication of the internal impedance. A better method is to use the rated impedance of the amplifier and an impedance not too far removed from the rated impedance.

The formula (derived in Appendix M):

$$\frac{R_L}{R_s} = \frac{m - n}{mn - m} \qquad (10\text{-}3)$$

where,

R_L is the rated load resistor used at the output of the amplifier,
R_s is the internal impedance of the amplifier as seen by the load,
m is a factor by which the load, R_L, has been increased during the test,
n is a factor by which the output voltage has increased during the test when the higher-impedance load was put into the circuit.

The procedure is simple. Adjust the signal generator so that the amplifier will deliver 1 to 5 undistorted watts with the rated load at the output. Do not touch any controls after this adjustment is made. Now measure the output voltage, e_o, using the rated load resistor, R_L, required at the output of the amplifier. Increase the load resistor by a factor of *m*, equal to any value between 1.25 and 2. The new load resistor is now mR_L and is equal to R_{L2}. Note the output voltage using this load. Determine *n* from the formula

$$n = e_{o2}/e_o \qquad (10\text{-}4)$$

Substitute the values for *m* and *n* in equations 10-3 to calculate the damping factor.

A good approximation can be obtained by using the equation:

$$R_L/R_s = e_o/(e_{o2} - e_o) \qquad (10\text{-}5)$$

In this equation, e_o is the voltage measured at the output with the standard load in the circuit, and e_{o2} is the voltage measured with an open load when R_{L2} is infinite.

FEEDBACK

The damping factor is related to the output impedance of an amplifier, and the output inpedance is in turn affected by the feedback. The output inpedance with series-fed feedback, R_{sf}, is related to the output impedance without feedback, R_s, by the relationship

$$R_{sf} = \frac{R_s}{1 - \beta K} \qquad (10\text{-}6)$$

where,

K is the gain of the amplifier,

β is the portion of output signal fed back.

If the feedback is negative, βK is negative and R_{sf} is less than R_s.

The feedback around the power amplifier can be measured in several ways. The simplest and most direct is to disconnect the feedback loop from the circuit and note how much the gain increases in db. The circuit in Fig. 9-1 should be used. The scope should be connected across the output to note whether the amplifier will overload when the feedback loop is opened. If it does overload, reduce the amount of signal fed to the amplifier.

In most amplifiers, the feedback loop is completed from the output transformer to the cathode (or emitter) in the first voltage amplifier of the power-amplifier section. Instead of an open loop, the cathode or emitter resistors may be bypassed with a large electrolytic capacitor. The increase in gain when the resistor is bypassed indicates the feedback in the circuit. It also takes into account the local feedback due to the resistor in the cathode or emitter circuit.

Which method to use is an academic question. The latter method will result in a larger number and a better-looking specification. Although possibly misleading, the figure does include all the feedback in the power amplifier. The first method only considers the feedback in the loop from the output.

INPUT IMPEDANCE

The measurements described in this chapter were all concerned with the output section of the amplifier. Considering the input, an important measurement is the input impedance. Feedback increases the input impedance by a factor of $1 - \beta K$, when the feedback voltage is in series with the input signal. If equalization circuits are used in a feedback loop connected to the first stage of an amplifier, the input impedance is frequently frequency-sensitive. The importance of input impedance has been discussed in previous chapters concerning phono cartridges, tape heads, microphones, etc.

It is quite simple to measure the input impedance. Connect the output from a very low-impedance signal generator (several ohms) into the circuit shown in Fig. 3-7. Connect a variable resistor in series with the hot input signal lead of the amplifier, between the signal gen-

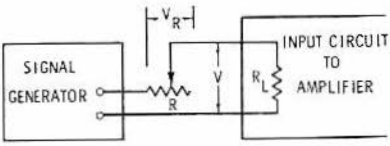


Fig. 10-8. Equivalent circuit for checking the input resistance of amplifier.

erator and amplifier. (This variable resistor, R, is not shown in the figure.) The equivalent input circuit becomes as shown in Fig. 10-8.

First, make R equal to zero and note the output voltage on the meter in Fig. 3-7. Next, vary R until the voltage is one-half the reading. Because of the voltage-divider action, R is now equal to the input resistance of the amplifier. If the input is reactive, this method cannot be used. Adjust R so that the voltage across it is equal to the voltage across Z_i, the input impedance. Only then will resistor R equal the input impedance of the amplifier.

This method cannot be used with transistor amplifiers, for the size of R reflects into the emitter circuit and affects the feedback. Here, it is best to use a low-impedance signal generator in series with a fixed resistor R.

It is desirable to measure the input impedance under the condition in which the amplifier will normally be used. In order for this to be done R should be made approximately equal to the impedance of the signal source ordinarily connected at the input. Using the circuit in Fig. 10-8, measure the voltage, V, across the input circuit of the transistor. Next, measure voltage V_R across the resistor R. The current, I_R, flowing through the resistor is V_R/R. The input impedance at the transistor is voltage V at the input of the transistor, divided by current V_R/R flowing through the input and R, namely:

$$\frac{V}{V_R} \times R = Z_{input} \qquad (10\text{-}7)$$

SEPARATION

Separation is concerned with the amount of undesirable audio-signal leakage between the two channels of a stereo amplifier. For the maximum stereo effect, no signal injected into one stereo channel should appear at the output of the second channel. The setup for making the test of how much signal will appear in the second channel is shown in Fig. 10-9.

A signal is fed to the left channel of the stereo amplifier. The input jack for the right channel is loaded with a resistor representing the equivalent impedance of the source. Thus, if it is a phono input, the resistor will be about 3300 ohms; a ceramic microphone input requires a resistor of about 470,000 ohms, etc. The outputs of both channels are each loaded properly with a resistor, R_L. A double-pole double-throw switch is used to select the output voltage the meter will read.

With the signal fed to the left channel, throw the switch so that the meter will read the output at the left channel. Be certain that all amplifier controls are set for a maximum output and for maximum flat frequency response. The output should be below the distortion level of the amplifier. The shape of the output signal may be monitored with a scope.

Now, throw the switch so that the meter will read the output from the right channel. The difference in db between the two readings is the channel separation. Repeat these readings over the entire audio band to check the separation at all frequencies. It is generally considered that the amount of separation below 100 Hz and above 10 kHz contributes little to the stereo effect.

Next, check the separation in the reverse condition. Feed the signal to the right channel, and load the left-channel input with a proper load resistor. Check the difference in output in db between the two channels. This figure is also valid in specifying separation. It is frequently different from the figure found in the original form of the setup in Fig. 10-9.

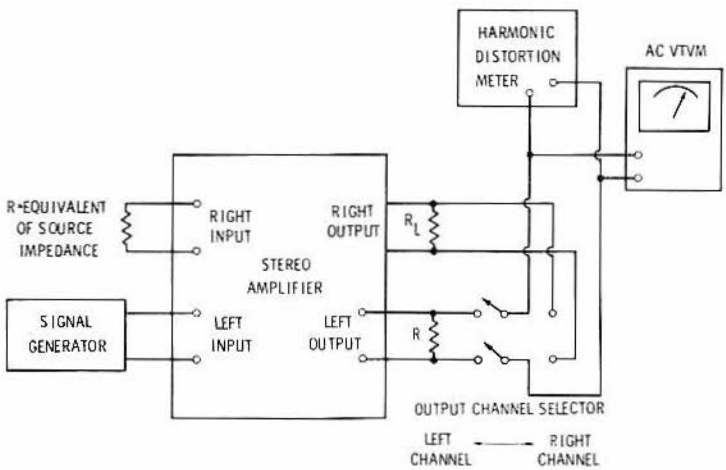


Fig. 10-9. Setup used to test separation between stereo channels.

CROSS TALK

Cross talk is similar to the concept of separation. However, cross talk does not measure the total signal leakage between channels. It measures only the harmonic components that leaked into the unused channel. To check for cross talk, feed a signal to one channel and set the level on your harmonic-distortion meter. Now switch the distortion meter to the second channel and adjust it for minimum output, reading the percentage of distortion. Note the difference, in db, between the first and second measurements. This is the cross talk.

The reading should be repeated for various frequencies and with the channels interchanged. A similar measurement can be made on a monaural amplifier for interaction between the various inputs on the one channel. Suppose a signal is fed to a phono input. Set the level on the harmonic-distortion meter. Then switch the input selector control to the tuner input, and adjust the distortion meter to measure the amount of harmonics. Cross talk is the difference between the two readings, in db, just made on the harmonic-distortion meter. This is a measure of how much of the harmonics from the phono signal will leak through when the input selector is set to reproduce the tuner channel.

Once again, the setup in Fig. 9-1 may be used, with the appropriate load resistor, R_L, connected in the circuit. This test should be repeated at various frequencies over the audio band and for all combinations of inputs.

APPENDIX A

Input and Output Voltages at the Wien Bridge

Refer to the output voltage, E_{out}, E_{R2} and E_{Z2} in Fig. 1-3. If both E_{R2} and E_{Z2} are in phase with E_{in}, then E_{out} is in phase with E_{in}. E_{R2} is definitely in phase with E_{in}, since E_{R2} equals E_{in} $(R_2/R_1 + R_2)$, which is pure resistance.

E_{Z2} is in phase with E_{in} and can be derived as follows: Assume the bridge is balanced at the angular frequency

$$\omega = \frac{1}{RC} \tag{1}$$

or:

$$\omega RC = 1 \tag{2}$$

where,

ω is $2\pi f$.

Then Z_1 is equal to:

$$Z_1 = R - \frac{j}{\omega C} = \frac{\omega RC - j}{\omega C} = \frac{1 - j}{\omega C} \tag{3}$$

where,

Z_1 is the impedance of R and C in series.

Z_2 is equal to:

$$Z_2 = \frac{jR}{j - R\omega C} = \frac{jR}{j - 1} \text{ or } \left(\frac{1}{Z_2} = \frac{1}{R} - \frac{\omega C}{j}\right) \tag{4}$$

where,

Z_2 is the impedance of R and C in parallel.

E_{Z2} is equal to:

$$E_{Z2} = E_{in}\left(\frac{Z_2}{Z_1 + Z_2}\right) = E_{in}\left(\frac{jR/(j-1)}{\frac{1-j}{\omega C} + \frac{jR}{j-1}}\right)$$

$$= E_{in}\left(\frac{\frac{jR}{j-1}}{\frac{(1-j)(j-1) + j\omega RC}{\omega C(j-1)}}\right)$$

$$= E_{in}\left(\frac{j\omega RC}{2j + j\omega RC}\right) = E_{in}\left(\frac{1}{2+1}\right) = \frac{1}{3} E_{in} \qquad (5)$$

where,

E_{in} is the input voltage.
Voltage without a *j* factor has phase shift between E_{Z2} and E_{in}.
Criterion 1 for oscillation has been fully met.

Effect of Stray Capacitance and Setting of Level Control on the Frequency Response of an Audio Amplifier

Assume the control in Fig. 2-5B is set at a point so that the upper portion has a resistance R_1 and the lower portion has a resistance R_2. The admittance of the lower portion is:

$$Y_2 = \frac{1}{X_2} = \frac{1}{R_2} + j\omega C = \frac{1 + j\omega CR_2}{R_2}$$

so that the reactance becomes:

$$X_2 = \frac{R_2}{1 + j\omega CR_2} \tag{1}$$

where,

X_2 is the reactance of R_2 in parallel with C,
Y_2 is the admittance of R_2 in parallel with C,
ω is $2\pi f$.

The impedance of the upper portion is R_1. Treating this circuit as a voltage divider:

$$\frac{e_{out}}{e_{in}} = \frac{R_2/(1 + j\omega CR_2)}{[R_2/(1 + j\omega CR_2)] + R_1} \tag{2}$$

Multiply the numerator and denominator in equation 2 by $(1 + j\omega CR_2)$ to yield:

$$\frac{e_{out}}{e_{in}} = \frac{R_2}{R_1 + R_2 + j\omega CR_1R_2} \tag{3}$$

Multiplying this equation by:

$$\frac{R_1 + R_2}{R_1 + R_2}$$

results in:

$$\frac{e_{out}}{e_{in}} = \frac{\dfrac{R_2}{R_1 + R_2}}{\dfrac{R_1 + R_2}{R_1 + R_2} + \dfrac{j\omega CR_1R_2}{R_1 + R_2}} = \frac{\dfrac{R_2}{R_1 + R_2}}{1 + \dfrac{j\omega CR_1R_2}{R_1 + R_2}} \qquad (4)$$

The frequency at which the response is 3 db from the center value is reached when the denominator takes the form $(1 + j)$, or:

$$\frac{j\omega CR_1R_2}{R_1 + R_2} = j \qquad (5)$$

and:

$$\omega = \frac{R_1 + R_2}{R_1R_2C}$$

The frequency response is a direct function of the relative values of resistors R_1 and R_2.

APPENDIX C

Relationship Between Rise Time and the Upper Frequency Limit of an Audio Amplifier

Just what the relationship is between rise time and the upper frequency limit can be determined with the help of the figure. We can see that the upper frequencies are limited by some configuration(s) similar to that shown. Here, the capacitor at the output results in a rolloff. The frequency at which the gain is down 3 db is $\frac{1}{2\pi RC}$. This can be seen when we consider the network as a voltage divider, where:

$$\frac{e_{out}}{e_{in}} = \frac{\frac{1}{j\omega C}}{R + \frac{1}{j\omega C}} = \frac{1}{j\omega RC + 1}$$

where,

j is an operator indicating a 90° phase shift,
ω is an angular frequency equal to $2\pi f$.

The output is 3 db down when the denominators is equal to $(1 + j)$, or:

$$\omega = \frac{1}{RC} \text{ and } f = \frac{1}{2\pi RC} \qquad (1)$$

Now, assume that the leading edge of a square wave enters the network shown in the figure, causing the capacitor to charge gradually. The equation for this network is:

$$e_{in} = V_C + V_R$$

$$V_C = \frac{1}{C}\int i\,dt \text{ and } V_R = iR \qquad (2)$$

Therefore:

$$e_{in} = \frac{1}{C}\int idt + iR \tag{3}$$

Rewriting this equation:

$$iR + \frac{1}{C}\int idt = e_{in} \tag{4}$$

where,

V_C is the voltage across the capacitor,
V_R is the voltage across the resistor,
j is the instantaneous current through the network,
dt is the time over which the capacitor is being charged.

The complete solution involves both the steady-state and the transient solutions. The force-free transient solution can be found by setting $e = 0$, resulting in:

$$iR + \frac{1}{C}\int idt = 0 \tag{4A}$$

Assume:

$$i = A\epsilon^{pt} \tag{4B}$$

as a solution to equation 4A. Substituting this gives:

$$iR + \frac{1}{C}\int A\epsilon^{pt} = 0$$

$$iR + \frac{1}{pc}A\epsilon^{pt} = 0$$

$$iR + \frac{i}{pc} = 0 \text{ (for } i = A\epsilon^{pt}\text{)}$$

$$i\left(R + \frac{1}{pc}\right) = 0$$

where,

A and p are quantities to be determined,
ϵ is a constant equal to 2.718,
t is an instant of time.

Solving for p results in:

$$p = -\frac{1}{RC}$$

Substituting this into equation 4B leaves as the solution for i:

$$i = A\epsilon^{-t/RC} \tag{4C}$$

At the start of the impulse, all of the current is across R. The current through the resistor at this instant is E/R. Writing this algebraically:

$$i = A\epsilon^{-0/RC} = A(1) = E/R$$

So that equation 4C becomes:

$$i = \frac{E}{R}\,\epsilon^{-t/RC} \qquad (4D)$$

where,

E is the instantaneous voltage across the resistor at the start of the square-wave pulse.

The steady-state solution for this is $i = 0$; the transient solution for the voltage across the capacitor is:

$$e_c = \frac{1}{C}\int \frac{E}{R}\,\epsilon^{-t/RC}\,dt \quad \left(\text{for } i = \frac{E}{R}\,\epsilon^{-t/RC}\right)$$

$$e_c = \frac{E}{RC}\,[-RC]\,\epsilon^{-t/RC} + E\epsilon^{-t/RC} + A \qquad (4E)$$

When $t = 0$, $e_c = 0$. At this time, equation 4E becomes:

$$0 = E\epsilon^{-0/RC} + A = E + A \text{ or } A = E$$

Substituting this into equation 4E yields:

$$e_c - E\epsilon^{-t/RC} + E \text{ or } e_c = E(1 - \epsilon^{-t/RC})$$

which can be written as:

$$e_o = E(1 - \epsilon^{-t/RC}) \qquad (5)$$

where,

e_o is the instantaneous voltage across the capacitor at any moment of time after the leading edge of the pulse has been applied,
E is the final voltage after an infinite time,
ϵ is a constant equal to 2.72.

We can now find the time it takes for the voltage to rise from 10 to 90 percent of its final value.

For convenience, let us assume that E in equation 5, the final voltage across the capacitor is 1. At the 90 percent portion of the final voltage, e_o must be equal to 0.9. Substituting these into equation 5;

$$0.9 = 1\ (1 - \epsilon^{-t/RC})$$
$$+\,0.1 = +\,\epsilon^{-t/RC}$$

Putting this into logarithmic form gives:

$$\mathrm{Log}_\epsilon\, 0.1 = {}^{-t/RC}$$

$$(-RC)\ \mathrm{Log}_\epsilon\, 0.1 =$$

$$t = (-\,2.3\ RC)\ \mathrm{Log}_{10}\, 0.1\ (\text{for } \mathrm{Log}_\epsilon = (2.3)\ \mathrm{Log}_{10})$$

$$t = -\,2.3\ RC\ (-1) = 2.3\ RC \qquad (6)$$

The time that it takes the voltage to reach 10 percent of its final value can be found by substituting 0.1 for e_o in equation 5:

$$0.1 = 1\ (1 - \epsilon^{-t/RC})$$

$$+\,0.9 = +\,\epsilon^{-t/RC}$$

Putting this into logarithmic form gives:

$$\mathrm{Log}_\epsilon\, 0.9 = {}^{-t/RC}$$

$$(-\,RC)\ \mathrm{Log}_\epsilon\, 0.9 = t = (-\,2.3\ RC)\ \mathrm{Log}_{10}\, 0.9$$

$$t = -\,2.3\ RC\ (-\,1 + .9542) = 0.105\ RC \qquad (7)$$

The time required for the voltage to rise from 10 to 90 percent of full value is the difference between equation 6 and 7.

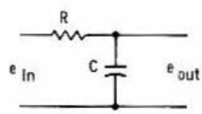
$$\text{Rise time} = t_r = (2.3 - .105)\ RC$$

$$= \frac{2.2}{\omega} = \frac{2.2}{2\pi f}\ \left(\omega = \frac{1}{RC}\right)$$

The frequency where the response is down 3 db is then:

$$f = \frac{2.2}{2\pi t_r} = \frac{0.35}{t_r}$$

Equation 8 will yield the 3-db point for frequency response at the high end of the band. This equation will give the 3-db point from actual measurement, whereas equation 1 will give the point from component calculations.



APPENDIX D

Solution of Equation for Network Placed Between a Constant-Amplitude Cartridge and Constant-Velocity Phono Preamplifier

With reference to Fig. 3-11, the equivalent circuit of the cartridge remains a voltage source, e_{in}, in series with a capacitor, C_2. The remainder of the circuit shown is designed to reproduce a record properly through an amplifier equalized for a velocity-type cartridge, while using one of the ceramic variety.

Considering the circuit as a voltage divider:

$$e_{out} = \frac{R_2}{X_{c2} + \dfrac{R_1 X_{c1}}{R_1 + X_{c1}} + R_2}(e_{in}) \qquad (1)$$

where $\dfrac{R_1 X_{c1}}{R_1 + X_{c1}}$ is the impedance of the parallel combination of R_1 and C_1, while $X_{c2} = \dfrac{1}{j\omega C_2}$. It also follows that:

$$\frac{R_1 X_{c1}}{R_1 + X_{c1}} = \frac{\dfrac{R_1}{j\omega C_1}}{R_1 + \dfrac{1}{j\omega C_1}} = \frac{R_1}{j\omega C_1 R_1 + 1}$$

where,

X_{c1} is the impedance of $C_1 = \dfrac{1}{j\omega C_1}$,

j is an operator indicating a 90° phase shift,

ω is angular frequency equal to $2\pi f$.

Substituting this into equation 1, and dividing both sides of the equation by e_{in}, yields:

$$\frac{e_{out}}{e_{in}} = \frac{R_2}{R_2 + \frac{1}{j\omega C_2} + \frac{R_2}{j\omega C_1 R_1 + 1}}$$

Simplifying this:

$$\frac{e_{out}}{e_{in}} = \frac{R_2\,(j\omega C_2)\,(1 + j\omega C_1 R_1)}{R_2\,(j\omega C_2)\,(j\omega C_1 R_1 + 1) + j\omega C_1 R_1 + 1 + j\omega C_2 R_2} \quad (2)$$

Now, let $k = C_2R_1$, $n = C_1R_1$, $m = C_2R_2$, and $S = j\omega$. Substituting these into equation 2, and simplifying all terms inside the parentheses, results in:

$$\frac{e_{out}}{e_{in}} = \frac{Sm\ (1 + Sn)}{S^2nm + Sm + Sn + 1 + Sk} \quad (3)$$

$$= \frac{Sm\ (1 + Sn)}{nmS^2 + S\ (m + n + k) + 1}$$

The two component factors in the denominator can be found by first letting the denominator be equal to zero. Thus:

$$S^2 + S\left(\frac{m + n + k}{nm}\right) + \frac{1}{nm} = 0$$

Letting $\frac{m + n + k}{nm} = \Sigma t$, and solving* for the two values of S, namely S_1 and S_2, yields:

$$S_1S_2 = \frac{-\Sigma t}{2} \pm \frac{\sqrt{(\Sigma t)^2 - 4/nm}}{2} \quad (4)$$

Now letting $\Delta = \sqrt{(\Sigma t)^2 - 4/nm}$, the two solutions for S are:

$$S_1 = \frac{-\Sigma t + \Delta}{2} \quad (5)$$

and:

$$S_2 = \frac{-\Sigma t - \Delta}{2} \quad (6)$$

Substituting this back into equation 3 produces:

$$\frac{e_{out}}{e_{in}} = \frac{Sm\ (1 + Sn)}{\left(S + \frac{\Sigma t + \Delta}{2}\right)\left(S - \frac{\Sigma t - \Delta}{2}\right)} = \frac{j\omega m\ (1 + j\omega n)}{\left(j\omega + \frac{\Sigma t + \Delta}{2}\right)\left(j\omega + \frac{\Sigma t - \Delta}{2}\right)}$$

---

*This type of solution can be found in most algebra books as a solution of the equation $ax^2 + bx + c = 0$. Here the two solutions of x are

$$x_{1,2} = \frac{-b \pm \sqrt{b^2 - 4ac}}{2a}$$

Multiplying the denominator by $\left(\frac{\frac{\Sigma t + \Delta}{2}}{\frac{\Sigma t + \Delta}{2}}\right)\left(\frac{\frac{\Sigma t - \Delta}{2}}{\frac{\Sigma t - \Delta}{2}}\right)$ and separating the equation into two factors yields:

$$\frac{e_{out}}{e_{in}} = \frac{j\omega m}{\left(\frac{\Sigma t - \Delta}{2}\right)\left(\frac{\Sigma t + \Delta}{2}\right)} \times \frac{(1 + j\omega n)}{\left(\frac{j\omega}{(\Sigma t + \Delta)/2} + 1\right)\left(\frac{j\omega}{(\Sigma t - \Delta)/2} + 1\right)} \quad (7)$$

Analyzing equation 7 results in the following four characteristics:

1. $\frac{j\omega m}{\left(\frac{\Sigma t - \Delta}{2}\right)\left(\frac{\Sigma t + \Delta}{2}\right)}$ is a constant factor and does not affect the frequency response. It shows that at zero cps the output is zero, and that the output rises at the rate of 6 db per octave from there.

2. The $(1 + j\omega n)$ indicates the point where the curve starts the second 6-db-per-octave rise. The frequency where the rise has advanced 3 db can be found by making $j\omega = j$ or $\omega = 1/n$. Call this ω

3 and 4. $1 + \frac{j\omega}{(\Sigma t + \Delta)/2}$ and $1 + \frac{j\omega}{(\Sigma t - \Delta)/2}$

where,

ω_1 is $2\pi f_1$.

indicate the point where the curve starts dropping at the rate of 6 db per octave. The angular frequencies where the output has dropped 3 db from its former plateau can be found by making the j terms in both factors equal to j.

Thus: $\frac{j\omega}{(\Sigma t + \Delta)/2} = j$ or $\omega = \frac{\Sigma t + \Delta}{2}$. Call this ω_2,

$\frac{j\omega}{(\Sigma t - \Delta)/2} = j$ or $\omega = \frac{\Sigma t - \Delta}{2}$. Call this ω_3.

where,

ω_2 is $2\pi f_2$,
ω_3 is $2\pi f_3$,
f_2 is a second frequency in hertz (Hz),
f_3 is a third frequency in hertz (Hz).

Using the constant called out in Fig. 3-11 to find ω_1 results in:

$$\omega_1 = \frac{1}{n} = \frac{1}{C_1R_2} = \frac{1}{(68 \times 10^3)(10^{-9})} = 14.7 \times 10^3$$

and: (8)

$$f_1 = \frac{\omega_1}{2\pi} = 2350 \text{ cps.}$$

To calculate f_2 and f_3, it must first be noted that R_2 can vary from 5000 to 30,000 ohms. It will be necessary to calculate f_2 and f_3 for both values of R_2. First, assume R_2 equals 5000 ohms. Then, if $n = 68 \times 10^{-6}$ (from previous calculations);

$$k = C_2R_1 = (5 \times 10^{-10})\ (68 \times 10^3) = 34 \times 10^{-6}$$

$$m = C_2R_2 = (5 \times 10^{-10})\ (5 \times 10^3) = 2.5 \times 10^{-6}$$

$$k + m + n = 1.045 \times 10^{-4} = 1.05 \times 10^{-4}$$

$$mn = (68 \times 10^{-6})\ (2.5 \times 10^{-4}) = 1.70 \times 10^{-10}$$

$$\Sigma t = \frac{k + m + n}{mn} = \frac{1.05 \times 10^{-4}}{1.7 \times 10^{-10}} = 6.17 \times 10^5$$

$$\Delta = \sqrt{(\Sigma t)^2 - \frac{4}{mn}} = \sqrt{38 \times 10^{10} - \frac{4}{1.7 \times 10^{-10}}} = 5.97 \times 10^5$$

Substituting these figures into the equations for ω_2 and ω_3 gives:

$$\omega_2 = \frac{6.17 \times 10^5 + 5.97 \times 10^5}{2} = 6.07 \times 10^5$$

$$\omega_3 = \frac{6.17 \times 10^5 - 5.97 \times 10^5}{2} = 10^4$$

so that:

$$f_2 = \frac{\omega_2}{2\pi} = 96{,}500 \text{ cps} \qquad (9)$$

and:

$$f_3 = \frac{\omega_3}{2\pi} = 1600 \text{ cps} \qquad (10)$$

while f_1 equals 2350 cps from equation 8.

Now assume the other extreme, where R_2 is equal to 30,000 ohms:

$$n = 68 \times 10^{-6}$$

$$k = 34 \times 10^{-6}$$

$$m = C_2R_2 = (5 \times 10^{-10})\ (3 \times 10^4) = 15 \times 10^{-6}$$

$$k + m + n = 117 \times 10^{-6} = 1.17 \times 10^{-4}$$

$$mn = (68 \times 10^{-6})\ (15 \times 10^{-6}) = 1.02 \times 10^{-9}$$

$$\Sigma t = \frac{1.17 \times 10^{-4}}{1.02 \times 10^{-9}} = 1.17 \times 10^{5}$$

$$\Delta = \sqrt{(1.17 \times 10^{5})^2 - \frac{4}{1.02 \times 10^{-9}}} = 0.99 \times 10^{5}$$

Substituting these values into the equation for ω_2 and ω_3 gives:

$$\omega_2 = \frac{1.17 \times 10^{5} + 0.99 \times 10^{5}}{2} = 1.08 \times 10^{5}$$

$$\omega_3 = \frac{1.17 \times 10^{5} - 0.99 \times 10^{5}}{2} = 0.09 \times 10^{5} = 9 \times 10^{3}$$

so that:

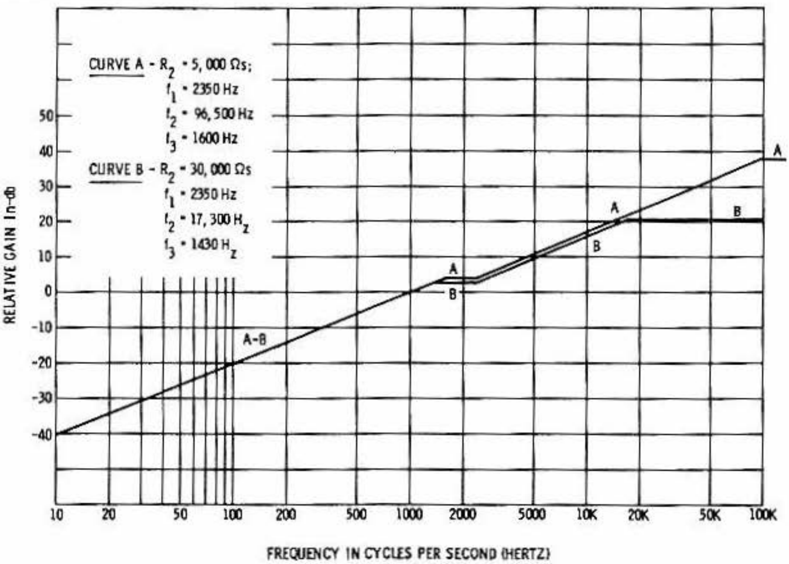
$$f_2 = \frac{\omega_2}{2\pi} = \frac{1.08 \times 10^{5}}{6.28} = 17{,}300 \text{ cps} \qquad (11)$$

and:

$$f_3 = \frac{\omega_3}{2\pi} = 1430 \text{ cps}$$

while f_1 equals 2350 cps from equation 8.

A plot of the curves using both values of resistors is shown in Fig. 1. Curves for circuits using resistors between 5,000 and 30,000 ohms fall between the two curves shown.



The break points, f_2 and f_3, involve resistor R_2. It is interesting to note how insignificant a part the variation of R_2 from 5000 to 30,000 ohms plays on the break points and consequently on the resulting frequency response of the network. When f_2 was calculated for a value of R_2 equal to 5000 ohms, the break point was at 96,500 Hz (equation 9). Calculated for R_2 equal to 30,000 ohms, it was 17,300 Hz (equation 11). For all practical considerations, both frequencies are outside the audio spectrum.

However, f_3 falls well within the audio spectrum. It is 1600 cycles when R_2 is 5,000 ohms, and 1430 cps when it is 30,000 ohms. Audibly, it is at about the same frequency. We shall compromise and consider f_3 to be 1500 cps. f_1 does not involve R_2 and remains at 2350 cps, regardless of what R_2 is. Variation of R_2 will not affect the frequency response to any noticeable degree. It will affect the output voltage from the network, however.

Derivation of Average Current in Half a Sine-Wave Cycle

The average current of a curve is equal to the area under the curve divided by the length along the X-(or ω) axis. Thus, for the sine-wave function $i = I_p \sin \theta$, the area under the curve for a half-cycle is:

$$\int_o^{\pi} I_p \sin \theta d\theta$$

where,

I_p is the peak current,
θ is the instantaneous angle of the sine wave,
$d\theta$ is a minute change in the angle,
π is a constant equal to 3.14.

and the distance along the X-axis is $\pi - o$. The equation for the average area under half a sine wave is:

$$\frac{1}{2\pi} I_p \int_o^{\pi} \sin \theta d\theta = \left[\frac{I_p}{2\pi} (- \cos \theta) \right]_o^{\pi}$$

$$= - \frac{I_p}{2\pi} (- 1 - 1)$$

$$= \frac{I_p}{\pi}$$

APPENDIX F

Proof That Differentiating Network Connected in Series with Audio Signal Emphasizes Harmonic Distortion

R and C in Fig. 6-7 form a differentiating circuit because the reactance, X_c, of C is considerably greater than the resistance of R. The current through R is thus determined by the reactance of C alone, and $i = C\frac{de}{dt}$, where e is the voltage across C and $\frac{de}{dt}$ is the change of voltage with time. Because X_c is large compared to R, the voltage across R is negligible, and e is effectively the total voltage applied to the network.

The voltage across R is the product of the current through the network, C de/dt, and R, which is:

$$e_o = RC\frac{de}{dt} \qquad (1)$$

where,

e_o is the output voltage.

The output voltage is then the derivative of the voltage at the input of the circuit.

Assume that a distorted signal is applied to the network from an audio amplifier:

$$e = E_1 \cos \omega t + E_2 \cos 2\omega t + E_3 \cos 3\omega t \qquad (2)$$

where,

E_1, E_2 and E_3 are the maximum voltages of three signal components,

$\cos \omega t$ represents the sinusoidal excursions of the fundamental,
$\cos 2\omega t$ represents the sinusoidal excursions of the second harmonic,
$\cos 3\omega t$ represents the sinusoidal excursions of the third harmonic,
ω is the angular frequency equal to $2\pi f = 6.28f$,
f is the frequency in cycles per second or Hz
t is on instant of time in the cycle.

Furthermore assume that the harmonic components are not visible on the scope. It is desirable to increase the size of the harmonic components when applied to the scope, so that they can be seen.

Put this signal, with the harmonics, through a differentiating network. Mathematically, it results in:

$$\frac{de}{dt} = \frac{d\,(E_1 \cos \omega t)}{dt} + \frac{d\,(E_2 \cos 2\omega t)}{dt} + \frac{d\,(E_3 \cos 3\omega t)}{dt}$$

$$= -\,\omega E, \sin \omega t - 2\omega E_2 \sin 2\omega t - 3\omega E_3 \sin 3\omega t \qquad (3)$$

Here, the relative size of the second harmonic is doubled and that of the third harmonic is tripled. Thus, with respect to the fundamental, the sizes of the harmonics are increased. When connected to a scope, the distortion now becomes more obvious.

APPENDIX G

Determining Second- and Third-Harmonic Distortion From Oscillograms

First assume an even-function type of signal which contains only second harmonics. The only pertinent terms from equation 6-7 are:

$$e = E_1 \cos \omega t + E_2 \cos 2\omega t \qquad (1)$$

where,

e is a voltage at an instant of time,
E_1 is the peak voltage of the fundamental,
E_2 is the peak voltage of the second harmonic,
$\cos \omega t$ represents the sinusoidal excursions of the fundamental,
$\cos 2\omega t$ represents the sinusoidal excursions of the second harmonic,
ω is the angular frequency equal to $2\pi f = 6.28f$,
f is the frequency in cycles per second or Hz,
t is an instant of time in the cycle.

A drawing of this equation and its component parts is shown in Fig. 6-11. In (A), the fundamental is shown; in (B), the second harmonic has been drawn; and in (C), the two are added to give the total of the two components, which is effectively equation 1.

In Fig. 6-11A, the amplitude of the signal at 2π is $e_1 = E_1 \cos \omega t = E_1 \cos 2\pi = E_1$; in (B), the amplitude of the harmonic at angle 2π is $e_2 = E_2 \cos 2\omega t = E_2 \cos 4\pi = E_2$.

where,

e_1 is the voltage of the fundamental at an instant of time,
e_2 is the voltage of the second harmonic at an instant of time.

The amplitude of the signal in (C) at 2π is $e_{2\pi}$. It is equal to the sum of the amplitudes of the components at 2π, or

$$e_{2\pi} = E_1 + E_2 \qquad (2)$$

Similarly, at angle 3π, $e_1 = E_1 \cos 3\pi = - E_1$ in Fig. 6-11A. In Fig. 6-11B, $e_2 = E_2 \cos 2(3\pi) = E_2 \cos 6\pi = E_2$. Because the signal in Fig. 6-11C is equal to the sum of the signals in (A) and (B), the sum of the amplitudes of the components at angle 3π is equal to:

$$e_{3\pi} = - E_1 + E_2 \qquad (3)$$

Adding equation 2 and 3 will give the relative size of the second-harmonic component, which is:

$$E_2 = \frac{e_{2\pi} + e_{3\pi}}{2} \qquad (4)$$

Subtracting equation 3 from equation 2 gives the relative size of the fundamental, which is:

$$E_1 = \frac{e_{2\pi} - e_{3\pi}}{2} \qquad (5)$$

The percentage of the second harmonic can be calculated from equation 6-8, which will be:

$$\text{percent 2nd harmonic} = 100\ (E_2/E_1) \qquad (6)$$

If a wave exhibits third-harmonic characteristics, as determined from Fig. 6-6, a similar line of analysis may be pursued. The pertinent terms from equation 6-7 are:

$$e = E_1 \cos \omega t + E_3 \cos 3\omega t \qquad (7)$$

where,

E_3 is the peak voltage of the third harmonic,

$\cos 3\omega t$ represents the sinusoidal excursions of the third harmonic.

It is assumed that the curve can be made to fit an even function (no sine components), and that no even harmonics are present, nor any harmonics above the third. Finally, it is assumed that the d-c component is zero. All these conditions can be readily realized in practice. The drawing of the component parts of equation 7 is shown in Fig. 6-12.

While, from Fig. 6-11, two equally spaced ordinates in one cycle were required to find the second-harmonic components, three equally spaced ordinates must be determined from Fig. 6-12 to find the magnitude of the third-harmonic component. These three ordinates are at angle 2π or 360°, angle $8\pi/3$ or 480°, and angle $10\pi/3$ or 600°.

Consider the three significant ordinates in sequence. At angle 2π, the amplitude of the fundamental is $e_1 = E_1 \cos \omega t = E_1 \cos 2\pi = E_1$; the amplitude of the harmonic is $e_3 = E_3 \cos 3\omega t = E_3 \cos 6\pi = E_3$. Thus:

$$e_{2\pi} = E_1 + E_2 \qquad (8)$$

where,

$e_{2\pi}$ is the amplitude of the total signal at angle 2π.

At angle $8\pi/3$, which is identical to angle $2\pi/3$, the amplitude of the fundamental is $e_1 = E_1 \cos \omega t = E_1 \cos 2\pi/3 = -E_1/2$; the amplitude of the harmonic is $e_3 = E_3 \cos 3\omega t = E_3 \cos 6\pi/3 = E_3$. It follows that:

$$e_{8\pi/3} = -\frac{E_1}{2} + E_3 \tag{9}$$

where,

$e_{8\pi/3}$ is the amplitude of the total signal at angle $8\pi/3$.

At angle $10\pi/3$, which is identical to $4\pi/3$, the amplitude of the fundamental is $e_1 = E_1 \cos \omega t = E_1 \cos 4\pi/3 = -E_1/2$; the amplitude of the harmonic is $e_3 = E_3 \cos 3\omega t = E_3 \cos 4\pi = E_3$. Thus:

$$e_{10\pi/3} = -\frac{E_1}{2} + E_3 \tag{10}$$

where,

$e_{10\pi/3}$ is the amplitude of the total signal at angle $10\pi/3$.

The sum of equations 8, 9, and 10 gives the relative size of the third-harmonic component. This is:

$$E_3 = \frac{e_{2\pi} + e_{8\pi/3} + e_{10\pi/3}}{3} \tag{11}$$

The relative size of the fundamental can be found by subtracting equations 9 and 10 from two times equation 6-17. This gives a relative fundamental amplitude of:

$$E_1 = \frac{2e_{2\pi} - e_{8\pi/3} - e_{10\pi/3}}{3} \tag{12}$$

From equations 11, 12, and 6-8, the percent of third-harmonic distortion is:

$$\text{percent third harmonic} = 100\ (E_3/E_1) \tag{13}$$

Some readers may be concerned with the incomplete analysis, where sinusoidal components are omitted. Experience shows that most types of distortion found in an amplifier can be considered as an even-function phenomenon. Furthermore, the method described, though useful, is approximate. Neglecting odd functions should add little to any error.

APPENDIX H

Generation of Sidebands by Nonlinear Operation of the Amplifying Devices

If vacuum tubes (or transistors) are driven into nonlinear operation, the current at the plate (or collector) follows the well-known expansion:

$$i = a_1 e + a_2 e^2 \qquad (1)$$

where,

i is the plate current,
e is the instantaneous signal voltage,
a_1 and a_2 are constants.

Higher-order terms in equation 1 (such as $a_3e^3 + a_4e^4 + a_5e^5 + \ldots$ etc.) have been omitted from the calculation. Although they may be significant in determining all components in the distortion, that exact an analysis is not required here. Equation 1 is sufficient to calculate the type of frequencies present in a nonlinear amplifier when two signals, E_1 and E_2, are fed simultaneously to such a device.

Denote the sinusoidal signals fed simultaneously to an amplifier (for intermodulation tests) by the equation:

$$e = E_1 \cos \omega_1 t + E_2 \cos \omega_2 t \qquad (2)$$

where,

E_1 and E_2 are the peak voltages of signals 1 and 2, respectively,
ω_1 and ω_2 are the angular frequencies of each of these signals and are equal to $2\pi f_1$ ($6.28f_1$) and $2\pi f_2$ ($6.28f_2$) respectively,
f_1 and f_2 are the respective frequencies in cycles per second or Hz.

Squaring equation 2 gives:

$$e^2 = E_1^2 \cos^2 \omega_1 t + E_2^2 \cos^2 \omega_2 t + 2E_1 E_2 \cos \omega_1 t \cos \omega_2 t \qquad (3)$$

Substituting equation 2 and 3 into equation 1 shows i to be:

$$
\begin{aligned}
i = a_1 \; & (E_1 \cos \omega_1 t + E_2 \cos \omega_2 t) + \\
& a_2 \; (E_1{}^2 \cos^2 \omega_1 t + E_2{}^2 \cos^2 \omega_2 t) + \\
& 2 \, E_1 \, E_2 \cos \omega_1 \, t \cos \omega_2 \, t
\end{aligned}
\qquad (4)
$$

From trigonometry, it can be shown that:

$$\cos^2 \omega t = \frac{1}{2} + \frac{1}{2} \cos 2\omega t \qquad (5)$$

It can also be shown that:

$$\cos (\omega_1 t + \omega_2 t) = \cos \omega_1 t \cos \omega_2 t - \sin \omega_1 t \sin \omega_2 t \qquad (6)$$

and:

$$\cos (\omega_1 t + \omega_2 t) = \cos \omega_1 t \cos \omega_2 t + \sin \omega_1 t \sin \omega_2 t \qquad (7)$$

Adding equations 6 and 7 results in:

$$\cos (\omega_1 + \omega_2) \, t + \cos (\omega_1 - \omega_2) \, t = 2 \cos \omega_1 t \cos \omega_2 t \qquad (8)$$

Substituting equations 5 and 8 into equation 4 gives us:

$$
\begin{aligned}
i = \; & a_1 \; (E_1 \cos \omega_1 t \; + E_2 \cos \omega_2 t) \\
& + a_2 \Bigg[E_1{}^2 \left(\frac{1}{2} + \frac{1}{2} \cos 2\omega_1 t \right) + E_2{}^2 \left(\frac{1}{2} + \frac{1}{2} \cos 2\omega_2 t \right) \\
& + E_1 E_2 \; (\cos \; (\omega_1 + \omega_2) t + \cos \; (\omega_1 - \omega_2) t \Bigg]
\end{aligned}
\qquad (9)
$$

From equation 9, is it possible to find some of the frequencies present when there is square-law nonlinearity. ω_1 and ω_2 are the fundamental frequencies which appear at the output as well as the input, and $2\omega_1$ and $2\omega_2$ are harmonics of these frequencies. $(\omega_1 + \omega_2)$ and $(\omega_1 - \omega_2)$ are the sum and difference components of these frequencies. If the higher-order terms had been considered in equation 1, more combinations of sum and difference frequencies would be present in the final equation.

Equivalent Sine-Wave Power

The relation between actual power and measured power for the curve in Fig. 7-5B can be derived as follows. The voltage in the modulated cases are related by the ratio of 4:1.

$$E_{p1} = 4E_{p2} \qquad (1)$$

Substituting this into equation 7-3, which states $E_p = E_{p1} + E_{p2}$, we find that:

$$E_p = 4E_{p2} + E_{p2} = 5E_{p2}$$

or:

$$E_{p2} = E_{p/5} \qquad (2)$$

but:

$$E_{p1} = 4E_{p2}$$

therefore:

$$E_{p1} = 4E_{p/5} \qquad (3)$$

where,

E_p is the peak voltage of the composite signal,
E_{p1} is the peak voltage of the low-frequency signal,
E_{p2} is the peak voltage of the high frequency.

To find the power contributed by each of these voltages, E_{p1} and E_{p2}, across the load resistor, R_L, the peak voltages must be converted to rms values. This is easily accomplished by dividing each of these by $\sqrt{2}$, or, from equation 3,

$$E_{rms1} = \frac{E_{p1}}{\sqrt{2}} = \left(\frac{4}{5}\right)\frac{E_p}{\sqrt{2}} \qquad (4)$$

and from equation 2,

$$E_{rms2} = \frac{E_{p2}}{\sqrt{2}} = \left(\frac{1}{5}\right)\frac{E_p}{\sqrt{2}} \tag{5}$$

where,

E_{rms1} is the rms voltage contributed by the low frequency signal,
E_{rms2} is the rms voltage contributed by the high frequency signal.

The power ω_1 delivered by E_{rms1} is, from equations 7-2 and 4:

$$\omega_1 = \frac{E^2_{rms1}}{R_L} = \left(\frac{16}{25}\right)\left(\frac{E_p{}^2}{2}\right)\left(\frac{1}{R_L}\right) \tag{6}$$

and the power ω_2, delivered by E_{rms2} is, from equations 7-2 and 5:

$$\omega_2 = \frac{E^2_{rms2}}{R_L} = \left(\frac{1}{25}\right)\left(\frac{E_p{}^2}{2}\right)\left(\frac{1}{R_L}\right) \tag{7}$$

The total power delivered to the load, R_L, is the sum of these two individual values, or:

$$\omega_1 + \omega_2 = \left(\frac{17}{25}\right)\left(\frac{E_p{}^2}{2}\right)\left(\frac{1}{R_L}\right) \tag{8}$$

Substituting equation 7-3 into equation 8, we get:

$$\omega_1 + \omega_2 = \left(\frac{17}{25}\right)\omega_p \tag{9}$$

Thus, the power delivered to the load during an intermodulation test is 17/25 the power delivered during the sinusoidal harmonic distortion test, if both tests are to be conducted under identical voltage swing conditions. This is known as the equivalent sine-wave power; 17/25 is an important number to remember.

If the output meter responds to rms values of the wave shapes rather than peaks, the power at the output can be calculated using equation 9. Power is a function of the rms voltage developed across the load resistor. An rms measuring voltmeter indicates (by calculation) the power that is across the load. If the meter reads the signal drawn in Fig. 7-5B, the equivalent sine-wave power is 25/17 the power calculated from the meter reading.

Derivation of RMS Voltage and Average Power

The rms voltage of equation 8-5, and the average power of equation 8-3, can be derived directly from the equation defining 8-2 sine-wave voltage.

The meaning of the constants have been stated in Chapter 8.

Substituting this into equation 8-1, power at any specific instant of time during the cycle is:

$$P = \frac{e^2}{R} = \frac{E^2_{max}\ (\cos 2\pi ft)^2}{R} \qquad (1)$$

where,

P is the instantaneous power dissipated by the load resistor,
e is the instantaneous voltage across the load,
R is the resistance of the load,
E_{max} is the peak value of the voltage in the cycle,
f is the frequency of the sinusoidal signal,
t is an instant of time in the cycle,
π is a constant equal to 3.14.

The expression $(\cos 2\pi ft)^2$ can be expanded trigonometrically as follows:

$$(\cos 2\pi ft)^2 = \frac{1}{2}\left[1 + \cos 2\ (2\pi ft)\right] \qquad (2)$$

$$= \frac{1}{2}\ (1 + \cos 4\pi ft)$$

Substituting this into the above equation for power results in:

$$P = \frac{E^2_{max}}{R}\left(\frac{1}{2}\right)(1 + \cos 4\pi ft) \qquad (3)$$

This equation is an expression of the power at any specific instant of time during the complete cycle. The average power through the cycle, the actual measured value at the output of an amplifier, is derived by simply studying the last equation. Over a complete cycle, the term cos $4\pi ft$ becomes zero because of zero-axis symmetry. The preceding power equation becomes:

$$P_{av} = \left(\frac{E_{max}^2}{R}\right)\left(\frac{1}{2}\right)(1 + 0) \qquad (4)$$

$$= \frac{E_{max}^2}{2}\left(\frac{1}{R}\right) = \left(\frac{E_{max}^2}{\sqrt{2}}\right)\left(\frac{1}{R}\right)$$

The quantity in the parenthesis, $E_{max}/\sqrt{2}$, is the rms value for voltage as indicated in equation 8-3. The average power is:

$$P_{av} = \frac{E_{rms}^2}{2} \qquad (5)$$

APPENDIX K

Comparison of Sensitivities of Two Amplifiers With Different Power-Output Capabilities

Assume E_1 is the voltage developed across the load resistor, R_L, for an output power of P_1 watts from the smaller amplifier. E_2 is the voltage developed for an output power of P_2 watts from the larger unit. Then:

$$E_1 = (P_1R_L)^{1/2} \tag{1}$$

and:

$$E_2 = (P_2R_L)^{1/2} \tag{2}$$

The known gain, G_2, of the larger amplifier is:

$$\frac{E_2}{v_2} = \frac{(P_2R_L)^{1/2}}{v_2} = G_2 \tag{3}$$

where,

v_2 is the input voltage at the larger amplifier for P_2 watts out.

The gain of the larger amplifier (assuming linearity) is G_2, regardless of the input-signal strength. The voltage gain of this larger amplifier can be stated as:

$$\frac{E_1}{v_x} = \frac{(P_1R_L)^{1/2}}{v_x} = G_2 \tag{4}$$

when an input signal, v_x, is fed to the input of this larger amplifier for an output of E_1 volts or P_1 watts. Using equations 3 and 4 and performing the necessary simplifications results in:

$$v_x = v_2\left(\frac{P_1}{P_2}\right)^{1/2} \tag{5}$$

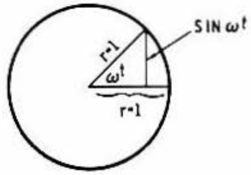
which is the expected exponential equation.

APPENDIX L

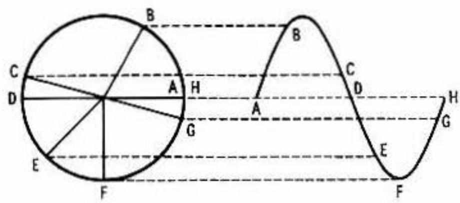
Derivation of Formula for Phase Measurements on a Scope Display

To facilitate the derivation of the formula for phase measurements, we must recall some elementary trigonometry. In Fig. 1A, a circle with a radius equal to "1" is shown. A horizontal radius is drawn, as well as a second radius rotated by an angle ωt. The length of the vertical line connected to the two radii is $\sin \omega t/1$, or simply $\sin \omega t$. Extending this construction for 360°, the well-known sine wave can be evolved. This is shown in Fig. 1B. A circle can thus be used to represent the sine wave.

Now, let us see just what will happen when we add two sine waves or circles, when one sine wave is applied to the vertical-deflection plates of the scope and the second to the horizontal. In the measurement, the scope is adjusted so that the resultant pattern will fit into a square on the cathode-ray tube screen. Hence, the vertical and horizontal amplitudes on the scope are equal. Several examples of this are shown in Fig. 2. Note that a zero-degree phase relationship shows



(A) A circle with a radius equal to one.

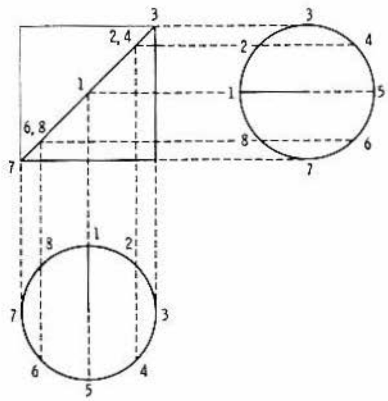


(B) A circle representing a sine wave.

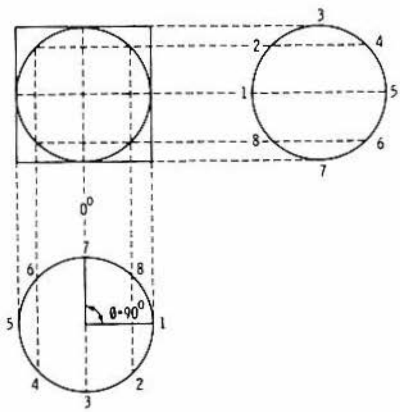
Fig. 1. Derivation of the sine wave.

up as a 90-degree difference between the two circles in the drawing. This is due to the fact that they are fed to two pairs of deflection plates (in the scope) which are oriented at 90 degrees with respect to each other. The period of each circle is divided into eight parts for ease of superposition in the final drawing in the square.

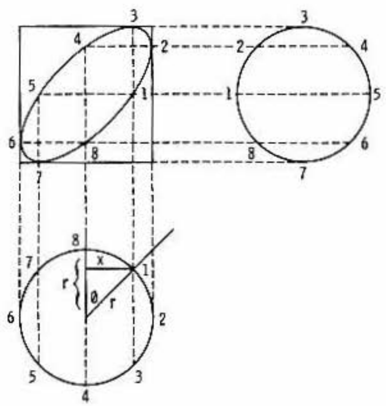
The phase difference between the two signals can be determined from the opening in the ellipse along the horizontal axis. In Fig. 2C, half of the opening of the ellipse along the horizontal axis is shown as x. The diameter of either circle as well as the length of each side of the square is 2r. The phase angle is created by the two radii, r. The sine of this angle is x/r.



(A) Zero-degree phase shift described by a straight line.



(B) A 90-degree phase shift described by a circle in a square.



(C) An ellipse with phase shift when ϕ is between 0 and 90 degrees.

Fig. 2. Resultant curves when two equal sinusoidal signals are added at 90 degrees to each other.

Thus, the relative phase of the two curves can be found from the equation:

$$\sin \phi = x/r, \text{ or } \phi = \text{arc} \sin x/r \qquad (1)$$

If the constructions in Fig. 2 were extended, the curves for ϕ from 0° to 90° and 270° to 360° would slant in the positive direction, while the curves for ϕ from 90° to 270° would slant in the negative direction. The elliptical shapes can be derived mathematically. If the voltages applied to the two sets of plates are controlled so that they both equal E, the deflection along the y axis will be:

$$y = E \cos \omega t \qquad (2)$$

and along the x axis, the deflection will be:

$$x = E \cos (\omega t + \phi) = E (\cos \omega t \cos \phi - \sin \omega t \sin \phi) = x_c - x_s \quad (3)$$

From equation 2:

$$\cos \omega t = y/E \qquad (4)$$

and from equation 3:

$$\sin \omega t = x_s/E \sin \phi \qquad (5)$$

Squaring equations 4 and 5 and adding them together gives:

$$y^2/E^2 + x_s^2/E^2 \sin^2 \phi = 1 \qquad (6)$$

which is the equation of an ellipse.

The other factor in the equation is:

$$x_c = E \cos \omega t \cos \phi = y \cos \phi \qquad (7)$$

a straight line.

The ellipse in equation 6 and the straight line in equations 7 make up the curves shown in Fig. 2.

Derivation of Formula for Determining Damping Factor

Referring to Fig. 10-7A, let R_L equal the rated output impedance of an amplifier, R_s the internal (Thevenon) impedance of the amplifier, and e_s the internal (Thevenon) voltage of the amplifier. The voltage across this output impedance is:

$$e_o = \frac{R_L}{R_s + R_L} e_s \tag{1}$$

Now increase the output impedance, R_L, by some small (1.25 to 2) factor, m. Consequently, the output voltage would be increased by a factor n. The equation for this, shown in Fig. 10-7b, is:

$$ne_o = \frac{mR_L}{R_s + mR_L} e_s \tag{2}$$

where,

m is the ratio of the new value of R_L to the old value of R_L. $m = \frac{R_L\ (\text{new})}{R_L\ (\text{old})} = 1.25$ to 2,

n is the ratio of the voltage across the new value of R_L to the voltage across the old value of R_L. $n = \frac{e_{rms}\ (\text{across new } R_L)}{e_{rms}\ (\text{across old } R_L)}$.

The change in voltage is due to the change in R_L. This change is across R_L.

Substituting equation 1 into equation 2 will yield:

$$\frac{nR_Le_s}{R_s + R_L} = \frac{mR_Le_s}{R_s + mR_L} \tag{3}$$

Canceling terms and cross-multiplying:

$$mR_s + mR_L = nR_s + nmR_L \tag{3A}$$

$$R_L\,(nm - m) = R_s\,(m - n) \tag{4}$$

$$\frac{R_L}{R_s} = \frac{m - n}{nm - m} = \text{damping factor} \tag{5}$$

which by definition is the damping factor. Equation 5 and the more accurate test method can be applied to any amplifier, whether it is of the transistor or vacuum tube type.

Equation 5 can be changed into the form:

$$\text{damping factor} = \text{D.F.} = \frac{1 - n/m}{n - 1} \tag{6}$$

if the numerator and denominator are divided by m. Letting m approach infinity, which is the same as saying let R_{L2} in Fig. 10-7B be an open circuit, changes equation 6 into:

$$\text{D.F.} = \frac{1}{n - 1} \tag{7}$$

Substituting $n = \frac{e_{o2}}{e_o}$ into the equation gives:

$$\text{D.F.} = \frac{1}{\frac{e_{o2}}{e_o} - 1} = \frac{e_o}{e_{o2} - e_o} \tag{8}$$

where,

e_o is the rms output voltage across the rated load resistor, R_L,
e_{o2} is the rms output voltage when R_L is open or made equal to infinity.

The damping factor is the voltage across the required load divided by the difference between the voltage across an infinite load and the voltage across the required load.

Index

T

V

W

MEASURING HI-FI AMPLIFIERS
by Mannie Horowitz

Most audio-amplifier circuits, no matter how complex, stem from just a few fundamental designs. Thus, to understand how these circuits operate and the various characteristics of each unit, you need a working knowledge of the basic principles. Moreover, few amplifiers are sold which do not have a long list of data that define the characteristics of these amplifiers.

Measuring HI-FI Amplifiers will supply these basic principles and will clarify the manufacturers' data by giving the meanings behind them. The book contains a comparison of various types of instruments used in measuring audio equipment. Also included are graphs and test setups to aid you in understanding the fundamental operation of hi-fi amplifiers. Various subjects such as checking frequency response, harmonic and intermodulation distortion, sensitivity and overload, measuring and matching phono, tape playback, and microphone equalization curves are discussed. Vacuum-tube and transistor amplifiers are both covered extensively.

Measuring Hi-Fi Amplifiers contains the answers to many questions. It is intended for the technician, engineer, or audiophile. Mathematics is used throughout, but may be bypassed without loss of continuity. Detailed deviations are relegated to appendices, where they may be referred to as desired.

ABOUT THE AUTHOR

Mannie Horowitz has written many articles and several books about audio, hi-fi, and other electronics subjects. He was graduated from Brooklyn College with a B.A. degree in physics, and studied electrical engineering at the City College of New York. Formerly project engineer at EICO and Harman Kardon, he is a member of the IEEE and their Professional Group on Audio and Electroacoustics, the Audio Engineering Society, and the Association of Orthodox Jewish Scientists. He is author of *Troubleshooting Audio Equipment*, another SAMS book.

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