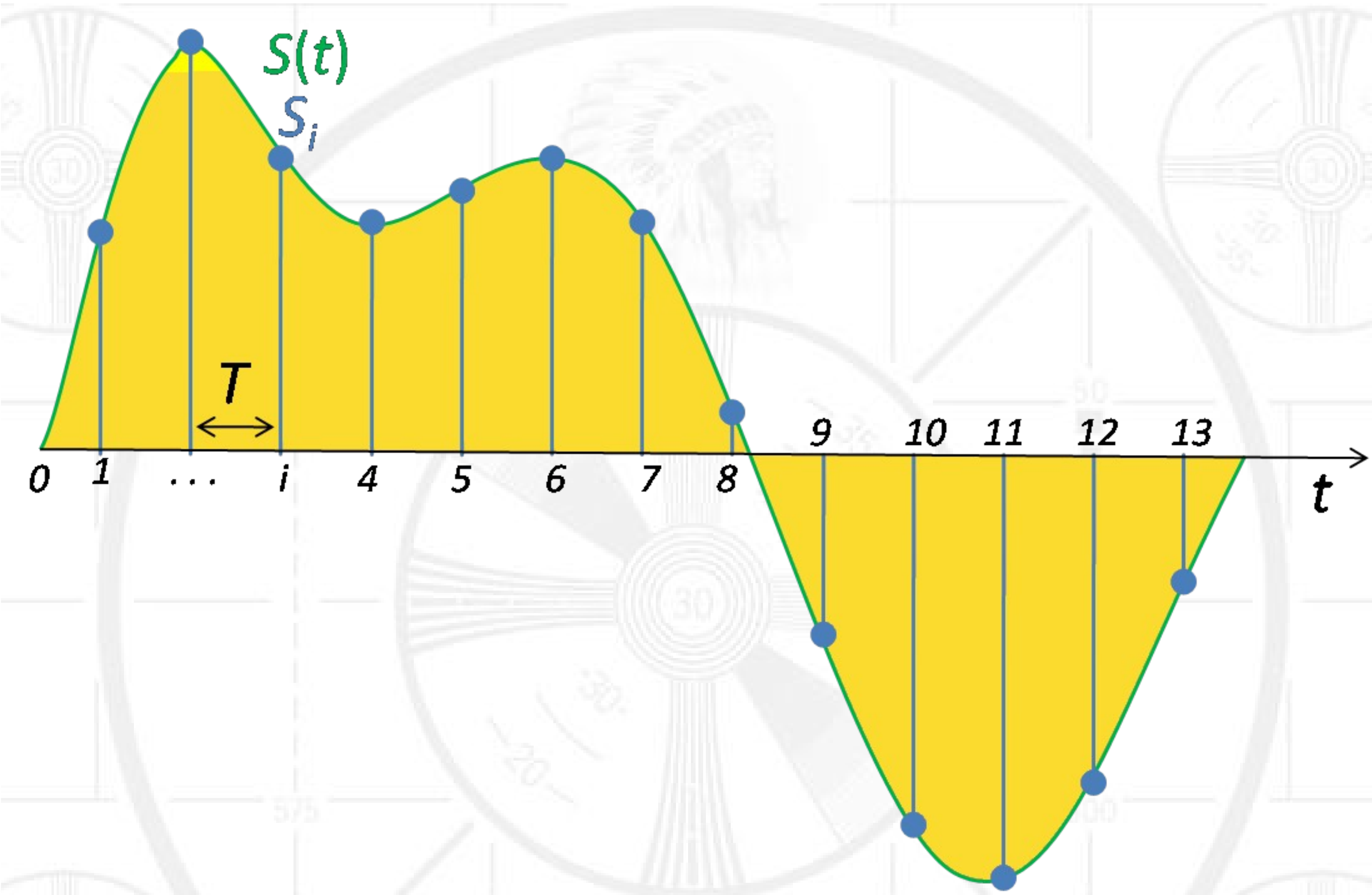
The Media Streaming Journal

June 2017



Covering Audio and Video Internet Broadcasting

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Welcome to The Media Streaming Journal

Greetings,

Audio is not a single user experience; it is an experience that can bring great enjoyment to scores of people. Broadcast audio engineering is not magic, but an art form that changes with each location setting, and audio equipment used. Each audio element must be seamlessly mastered and combined to provide an overall exquisitely delicious experience for the listening audience's ear. You must take the time to understand each detail and its ability to enhance or degrade the other elements.

Audio must be suitably produced and engineered to retain a station's audience. Audio that is hard to understand, poorly engineered or haphazardly thrown together will be a detriment and dissuade people from returning and listening to your station.

To the average listener: no matter how great the band, if the sound engineer sucks, the band sucks.
Anonymous

Please feel free to contact either the Publication Director (Derek Bullard) or myself if you have any questions or comments regarding The Media Streaming Journal.

Namaste

David Childers

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The Grand Master of Digital Disaster
(Editor In Chief)

The Media Streaming Journal

What is in this edition of the Media Streaming Journal

Excerpts from:

The Recording And Reproduction Of Sound
Olivery Read
Howard W. Sams & Co., Inc.
1952

Excerpts Edited By:

David Childers

Contents:

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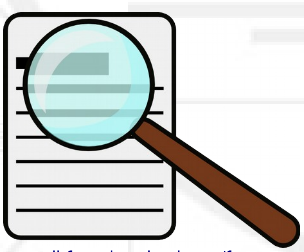
Scenic Television is an Internet television station that broadcasts the sights and sounds of nature 24 hours a day. Savor exotic tropical beaches, or relax in a remote rain forest. Meditate at a bubbling stream, or relish the view of soft rolling waves at a lake. We have beautiful nature video from locations all around the world.

Scenic Television originates from the Gulf coast of South Alabama and broadcasts to a global audience. The television broadcast is accessible on any device with an Internet connection. Such electronic devices include desktop computers, laptops, tablets, smartphones, game platforms, and Internet-connected televisions.

http://television.scenicradio.com



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The Recording and Reproduction of Sound

•

OLIVER READ, D.Sc., D.Litt.
Editor, Radio and Television News
and
Radio-Electronic Engineering

•

HOWARD W. SAMS & CO., INC.
INDIANAPOLIS 5, INDIANA
1952

CHAPTER 1

Behavior of Sound Waves

The behavior of sound waves—introduction to the history, development, and applications for the recording and reproduction of sound. Mechanical, electrical, and electronic methods will be covered, including sound on wire, disc, tape, film, etc.

● To fully understand the many problems which enter into the recording and reproduction of sound, we must first take into consideration the basic concept of sound. We must understand and fully appreciate the characteristics of sound or we cannot apply satisfactory techniques to obtain a true facsimile of the original sounds we expect to reproduce.

Webster defines *sound* as follows: "Sensation due to stimulation of the auditory nerves and auditory centers of the brain, usually by vibrations transmitted in a material medium (commonly air) affecting the organ of hearing-vibration energy which occasions such a sensation. Sound is propagated by progressive longitudinal vibratory disturbances (sound waves)."

Thus, it becomes apparent that the source of sound lies in the realm of physics, while the effect of sound is a physiological consideration. The engineering of sound consists of controlling the cause so as to produce the desired effect.

The theory of *sound waves* may be explained in simplified form thusly: If a small stone is dropped into a pool of still water, waves will be set up which will travel in all directions away from the point of impact. If our original stone were small in physical size, only waves small in height would result. However, if a large stone were dropped into the still water, we would discover that we have generated waves having a greater height. The up and down movement of the waves represents the *amplitude* or the intensity of the waves.

Differing from the behavior of water waves, sound particles of air do not move up and down and across in the pattern in which waves move but all move in the same direction. These particles of disturbed air literally bump one another as they travel through space. The air, therefore, is alternately compressed and rarefied.

The number of waves passing any fixed point per second represents the *frequency* or number of waves per second. Therefore, the *frequency* depends upon the number of waves traveling past one spot during an interval of one second. If we tie a string to a stone which is immersed in still water and move it up and down at the rate of 100 times per second, we will send out waves from the point of impact at the rate of 100 per second. In the study of acoustics we refer to *frequency* in terms of cycles per second, abbreviated cps.

Now suppose we took a large stone and moved it up and down very slowly in and out of the water, we would then set up waves at a lower frequency, but due to the greater displacement of the water by the object, we would create waves of greater amplitude.

The top of the wave is referred to as the *crest,* while the bottom is known as the *trough.* One *cycle,* as far as one wave is concerned, would be one complete wave beginning from its normal still point, building up to a crest, passing again through its still point, down to the lowest point, the trough, and its return to its normal position. See Fig. 1-1D.

Sound waves are not limited to but one frequency. In fact the speaking voice is made up of a variety of *complex waves.* These are continually varying in both frequency and amplitude. In other words, if we raise our voices in *pitch,* we are actually increasing the *frequency* and the *louder* we talk, the greater will be the *amplitude* of the sound waves sent out by our speaking mechanism.

All sound waves are composed of frequency, intensity, periodicity, and waveform.

1. *Frequency* is the speed of vibration or number of complete cycles per second. Frequency also determines pitch. If we double the speed of vibration, we raise the pitch one octave. For example, a note having 1000 vibrations (cycles per second) would be raised by one octave if the frequency were doubled to 2000 cps.

2. *Intensity* is the amplitude or power of vibration. Intensity therefore determines the loudness of a sound. If the pressure of a sound wave is doubled in intensity, we increase the power by about 3 decibels. The *decibel* is a ratio of power, voltage or current and not a quantity.

In the behavior of sound waves various ranges of intensities and pressures are so great that it is necessary to have some means which will conveniently measure volume or amplitude. The *decibel* is a unit used for expressing the magnitude of a change in either a sound level or a signal level. One decibel is the amount that the pressure of a pure sine wave tone must be changed in order for the change to be just barely detectable by the average human ear. The amount of change in power level expressed in decibels is equal to ten times the common logarithm of the ratio of the two powers.

$$\text{db.} = 10 \log_{10} P_1/P_2$$

3. *Periodicity* is the lack of, or the existence of, rhythm in sound. Therefore, musical sounds are *periodic,* while street noises, the jingle of keys on a chain, etc. are *non-periodic.*

4. *Waveform* is a direct pattern of vibration. Most fundamental sound waves are modified by secondary vibrations. The *timbre* of sound is determined by the waveform. Thus it is possible to distinguish particular notes played on various musical instruments such as the violin and the flute.

Distortion

If a microphone, radio tuner, pickup, amplifier or speaker is incapable of reproducing a true picture of the original sound waves, *distortion* will occur. This distortion may come from either mechanical or electrical defects in the system. *Mechanical distortion* may be caused from such conditions as having a needle in a phonograph pickup move because it has not been tightened thoroughly by means of the needle set screw. In other words, vibrations set up within the pickup could not transmit or move the needle in perfect cadence with the original electrical vibrations. The needle would be permitted to move freely and to follow random vibrations of its own choice instead of moving in perfect unison with the armature in the pickup.

Many reproducer (loudspeaker) cones vibrate at some particular frequency not found in the music that is fed to the speaker for reproduction. The resulting buzzing effect would be a form of *mechanical distortion.* The above illustrations are typical but many others may be encountered throughout the equipment if care is not exercised to

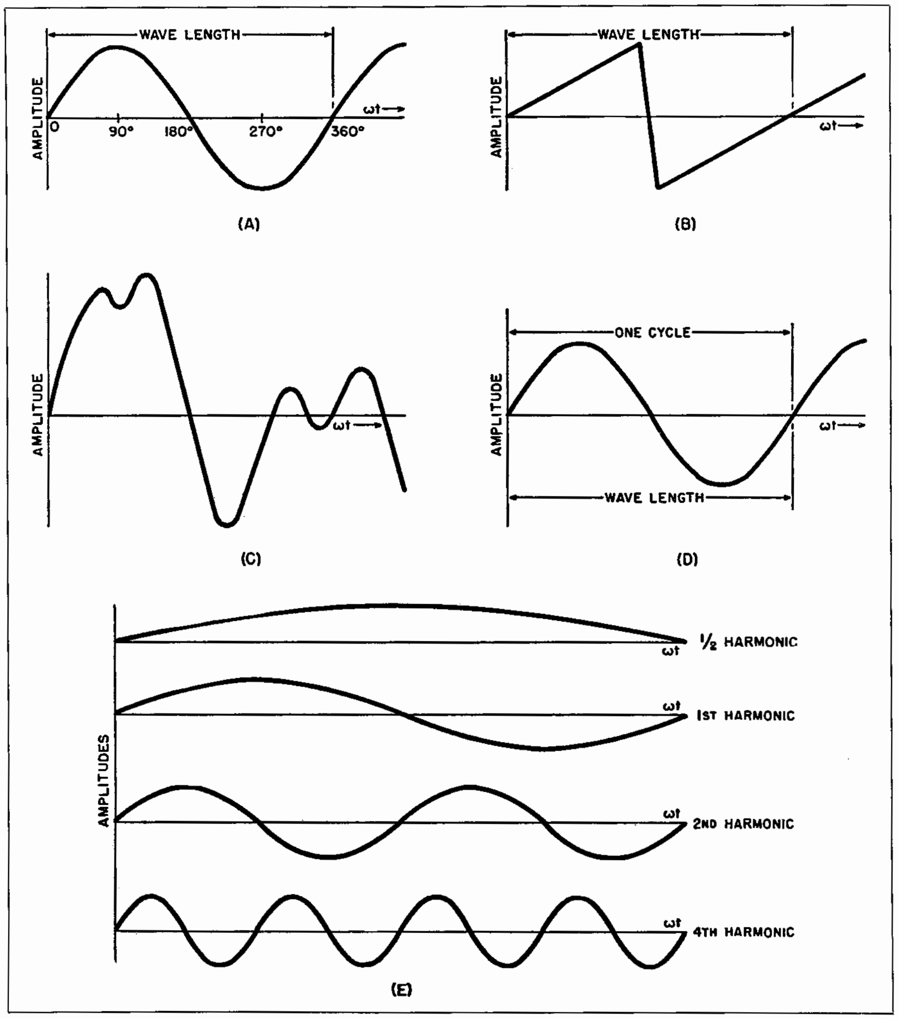


Fig. 1-1. Illustrating the characteristic behavior of wave motion; (A) sine, (B) periodic, (C) random, (D) one cycle and (E) harmonics.

adjust each and every part which might cause mechanical distortion.

Generally speaking, a sound is said to be distorted when the waveform is altered in transmission or when the intensity of any frequency is suppressed or exaggerated out of its natural proportions.

Electrical distortion is caused by the inability of the microphone, amplifier or speaker to give a true reproduction (facsimile) of frequency. For example; the delicate diaphragm on a microphone may become warped from excessive heat, etc. As pressure from the sound waves strikes this distorted diaphragm, the resulting currents would become distorted. There are many forms of electrical distortion. A poorly designed amplifier, overloaded audio stages, impedance mismatching and improper biasing will all result in electrical distortion.

The Ear

There are three main divisions to the human ear: (a) The outer ear, which is made up of the visible portion of the ear and the canal which is not visible. (b) The middle ear, which is the receiver for sound approaching the eardrum and the conducting media for sounds through a chain of three small bones (the ossicles), to the inner ear. (c) The inner ear is the delicate container filled with fluid in which is immersed the nerve ends which perceive sound and which also transmit messages to the brain. The part of the inner ear that perceives sound is called the cochlea. There is still another part of the inner ear in the form of semicircular canals which give us our sense of balance.

In the transmission and reproduction of sound, consideration must be given to the human ear as a transducer (here the sound waves are transformed into a stimulus for the auditory system). The ear exhibits certain non-linear characteristics which affect the fidelity with which sounds are received. For instance, harmonics of a fundamental tone are generated at different intensity levels. This may be compared in a sense to higher resonant modes of vibrating mechanical systems and their general effect is to reduce fidelity of the original tone.

There is a definite phase effect existing between the sound wave and generated harmonics in the ear and the effect contributes to the distortion experienced in hearing. As may be expected, there are definite intensity limits between which speech and music are reproduced with perfect fidelity.

There are many factors affecting the quality of a tone. Among these are the harmonic content, pitch, and intensity. The *quality* of a tone is also called *timbre* and should be distinguished from *fidelity*. Timbre is a relative quality of sound while fidelity is a true measure of the accuracy with which a sound is reproduced. While much has been done in measuring timbre, the results are subjective. "Full," "Rich," "brassy," "metallic" are some of the terms used in describing the timbre of a sound.

Speech has a frequency range of from approximately 100 to 8000 cps.

Sounds having very low frequencies possess the most power and result in naturalness and apparent loudness. High frequency sounds provide intelligibility. If we eliminate all sounds below 1000 cps we take but little from the clarity of the sound, but we do notice that the sound appears to be unnatural. However, if we eliminate sounds above 1000 cycles, we find that the speech is unintelligible but, as far as volume is concerned, there is little, if any, apparent change.

Speech may be considered as a series of *periodic* sound waves, emitted in a certain sequence. Association of particular sound sequences with particular ideas is the distinguishing feature between speech and noise. The so-called "scrambled" speech systems serve as an excellent illustration of this point. In such a system, spoken sounds are distorted by specially designed circuits and then transmitted in the distorted form. The receiver must have a rectifying circuit to convert the wave impulses back to their original undistorted form rendering it intelligible to the human ear.

Music

That sound we choose to call *music* consists of a sequence of single tones or a sequence of several tones played in unison. Pure single tones are harmonic in waveform and the key or pitch is

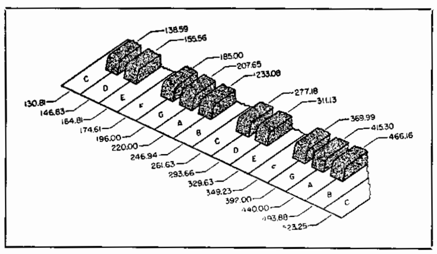


Fig. 1-2. Diagram illustrates keys of a piano keyboard and shows tempered scale for one octave on each side of middle C.

determined by the frequency (with certain modifications to be discussed later). The scale is an ascending series of tones with definite frequency intervals. Fig. 1-2 shows a portion of a standard piano keyboard for one octave on each side of middle C. The frequencies noted represent a so-called tempered scale which divides an octave into twelve equal intervals. The relative frequencies of the tempered scale (omitting sharps and flats) are shown below for one octave:

Key	*Relative Frequency*
C	1.000
D	1.122
E	1.260
F	1.325
G	1.498
A	1.682
B	1.887
C	2.000

The combinations of two or more tones make up a periodic composite sound. Here again the psychological phase of music shows up, for if the new sound is pleasing to the ear, it is called harmonious. If, on the other hand, the sound is irritating, the combination of tones is said to be in discord.

Frequency Range

If a tone is maintained at a constant intensity but its frequency is raised and lowered, high and low frequency points will be reached beyond which there will be no sensory perception by the ear. Generally this *frequency range* extends from 30 to 16,000 cps. The frequency range for conversational speech is from about 100 to 8000 cps, as mentioned previously. In terms of power, conversational speech will vary over about 50 db. and the intensity of the loudest sound will be about 100,000 times that of the weakest.

A large symphony orchestra, with instruments producing an abundance of bass notes and overtones extends over practically the entire frequency range of the ear—30 to 16,000 cps. The over-all volume range of a symphony orchestra from the softest passages through the loudest passages and peaks runs about 70 db. or an intensity range of 10,000,000 times.

Reproducing systems for highest fidelity should have a frequency range that is uniform from about 30 to 16,000 cps and a volume range of 70 db.

Of equal importance, the original sounds must be reproduced at the same power levels as were present in the studio. This is important if proper *balance* between the lower and higher frequencies is to be maintained.

When a sound is reproduced at a higher level than that of the original, the low bass frequencies will appear to be *accentuated.* If the sound is reproduced at a lower level than the original, bass frequencies will appear to be *attenuated* (suppressed) with respect to other portions of the frequency spectrum.

Noise

Noise is usually considered to be *random* sound waves with little or no periodicity. That does not completely define noise for certain "noises" are associated with certain commonplace events; for instance, the creaking of doors or gates, the clicking sound made by walking on a hard surface with leather heels, etc. The hum of an electric motor is considered noise, yet an analysis of the sound by means of an oscillogram would indicate a definite periodicity. Again it seems that the distinction between the classification of sounds is psychological in origin and relative to the observer.

The Technical Aspects of Sound

Sound is usually characterized as wave motion of which there may be three forms, namely, harmonic, periodic, and random.

Waves are propagated disturbances and may be *transverse* or *longitudinal,* depending upon the direction of the disturbance.

In a *transverse* wave, the particles of the medium vibrate in a direction perpendicular to the direction of propagation. For example, in a water wave, the individual particles of water move up and down, while the direction of wave motion is along the surface, or perpendicular to the particle motion. Thus, a water wave is a transverse wave.

A *longitudinal* wave is so named because the particles of the medium vibrate in a direction parallel to the direction of propagation. In a sound wave, for example, the individual particles of air move back and forth in the same direction that the wave is traveling. Thus, a sound wave is a longitudinal wave.

A *wavefront* may be classified as *plane* or *spherical*. Since a wave in general spreads out uniformly in all directions from its source or origin, the wavefront will be spherical close to the source. However, at some distance from the source the curvature is practically zero, and the wave is considered to be a plane wave. If a pebble is dropped in a still pond, circular waves will spread out in all directions, but by the time they have traveled a considerable distance, the waves will be essentially straight. Thus, a wave which starts out spherical (or circular in two dimensions) eventually becomes essentially a plane wave.

Harmonic Motion is a wave pattern of the sine curve type illustrated by Fig. 1-1A. All harmonic motion can be described by an equation of the form: $y = A \sin \omega t$ where: y = displacement of the disturbance, A = maximum displacement (which occurs at 90° and 270°), ω = circular frequency (to be defined in detail) in radians per second, t = time after disturbance is initiated.

Periodic Motion is a wave pattern compounded from two or more harmonic motions of different frequencies as in Fig. 1-1B. Periodic motions are analyzed by determining the various harmonic components. The character of any periodic motion is controlled by the number and magnitude of its various harmonic components.

Random Motion is a conglomeration of harmonic motions exhibiting little or no periodicity. Speech, squeaks, scraping, etc., all produce random motion unless they are sustained for long periods. The number of harmonic components is so great and their duration is so irregular as to make the harmonic analysis of random motion a practical impossibility. A typical wave pattern of random motion is shown in Fig. 1-1C.

Harmonic and periodic motions have certain defining characteristics which must be considered. These are:

Cycle, which is a sequence of events or motions that recur in exactly the same order in certain time intervals. For example, the motion of a pendulum started from one side, allowed to swing to the opposite and back to the starting side, comprises one complete cycle. This is illustrated in Fig. 1-1D.

Period is the time required for the motion to complete one set of recurring values, or one cycle. It is usually defined by the symbol T.

Frequency is the rate at which the cycles recur and is usually presented in one of two ways. If the frequency is given as ω, it is known as the circular frequency and has the dimensions, radians per second. The frequency may also be given as cycles per second or the symbol f in which case it is connected to the circular frequency by the relation: $f = \omega/2\pi$.

Wavelength is the distance between two corresponding points on harmonic or periodic waves (Fig. 1-1A) and is denoted by the symbol λ.

Fundamental Frequency is the lowest component frequency of a periodic wave motion.

Harmonic Frequency or overtone, when applied to music, is that component of a periodic wave motion whose frequency is an integral multiple of the fundamental frequency.

Sub-Harmonic Frequency is that component of a periodic wave whose frequency is an integral submultiple of the fundamental frequency.

An *Octave* is the interval between two waves whose frequencies are in the ratio of 1:2.

Each of these above four terms is illustrated in Fig. 1-1E.

Behavior of Sound Waves

From a purely physical concept, sound can be considered as an alteration in pressure, displacement of particles, or the velocity of particles in elastic media. These pressure or velocity changes are produced by a *transducer,* which is an electromechanical or electroacoustic system for converting electrical vibrations into mechanical or acoustical vibrations respectively. Microphones and loudspeakers are typical transducers as is the sounding diaphragm of Fig. 1-3. The motivating source for the diaphragm can be mechanical (as shown) or electrical as in the case of a loudspeaker. Since sound is a particular type of wave motion, it becomes desirable to consider particular behavior of waves.

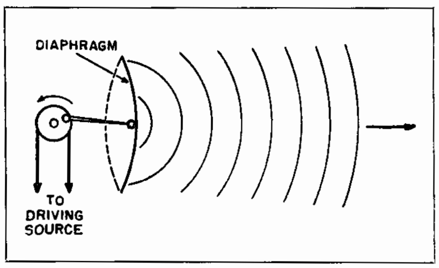


Fig. 1-3. Sound waves created by the movement of a mechanically-driven diaphrogm.

1. If two harmonic waves pass through the same point while vibrating at the same frequency, the resultant wave formed by adding the displacements will also be a harmonic wave.

2. Conversely, any harmonic wave can be resolved into a number of component harmonic waves. This principle is covered by *Fourier's Theorem.*

3. If two harmonic waves start at the same point and experience similar displacements at the same time, they are said to be *in phase* with each other. However, if one wave lags or leads the other, then they are said to be *out of phase* (Fig. 1-4B). It is customary to denote phase difference in angular units as that part of 360 degrees or 2π radians.

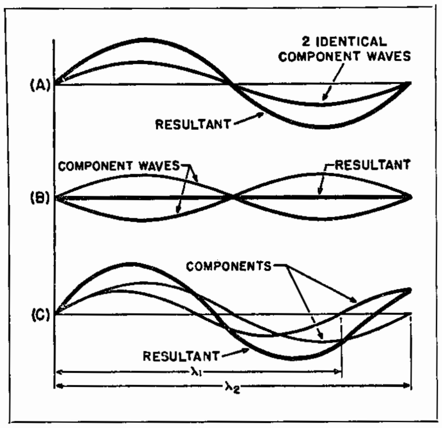


Fig. 1-4. The combinotion of two harmonic waves vibrating at the same frequency.

Almost everyone has witnessed some example of wave interference. For instance, in Fig. 1-4A, the component waves (both are identical and appear as one) form a harmonic wave at twice the original amplitude. This is said to be *constructive interference.* In Fig. 1-4B the component waves are of the same frequency and amplitude but 180 degrees out of phase, and therefore, cancel each other in what is called *destructive interference.* The resulting wave caused by the superposition of two waves of almost, but not quite the same frequency as in Fig. 1-4C has the characteristic of *beats,* i.e., alternate periods of constructive and destructive interference. Beat phenomena are particularly important in their application to superheterodyne reception.

When two identical waves travel with the same speed but in opposite directions, the resulting wave is known as a stationary or *standing wave.* While there is no translational motion, the vibratory displacements persist. The displacements are of equal magnitudes for distances one-half wavelength apart and the wave takes the form shown in Fig. 1-5A at some particular time and then reverses as shown in Fig. 1-5B. As

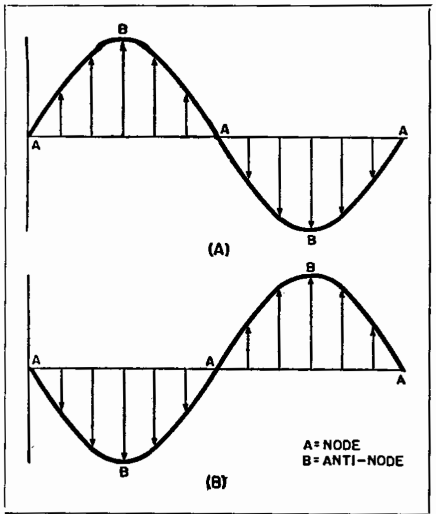


Fig. 1-5. Illustrates displacement of waves.

can be seen from the figure, certain portions of the wave experience little or no displacement. The points of maximum displacement are called *anti-nodal points* or *loops* and the points of zero displacement are called *nodal points.* Standing waves are of great importance in the theory of antenna design.

The Speed of Sound

The speed of sound in any medium is a function of the density and elastic qualities of the medium.

The variation of the speed of sound with temperature is expressed by the formula:

$$V_t = V_o \sqrt{1 + \frac{t}{273}}$$

where:

V_t = velocity at temperature t
V_o = standard velocity at 0° C
t = temperature in degrees C

The speed of sound in air at one atmosphere pressure at 0° C is approximately 1088 ft/sec. while at 100 atmospheres the speed of sound is 1150 ft/sec.

Pitch and Intensity

The pitch of a sound is best characterized by its frequency (Fig. 1-6) except for certain psychological effects. The audible range is known to be between 16 and 20,000 cycles per second, but the average ear does not hear below 30 cycles per second nor above 16,000 cycles. However the pitch, as received by the human ear, is not a direct function of the frequency but varies with the intensity. This effect has been investigated in considerable detail by Fletcher.[1]

While intensity is popularly considered to be synonymous with loudness, a distinct difference exists between the two. The former is a pure physical quantity while the latter is psychological in origin. In a strict physical sense, intensity is defined as the average rate of flow of energy per unit area in a direction normal to the rate of flow. It is expressed by the formula:

$$I = \frac{(P_{max})^2}{2\rho_o c}$$

where P_{max} = the maximum pressure developed above the steady pressure for no disturbance, c = wave velocity, ρ_o = medium density, I = intensity in watts/sq.cm.

The equation may also be written in the form:

$$I = \frac{1}{2} \rho_o c \ \omega^2 \ d_{max}$$

where d_{max} = maximum displacement and ω = circular frequency = $2\pi f$.

For acoustic measurements, the choice of the equation depends on whether the recording instruments respond to pressure or displacement variations.

Current practice now employs the decibel as a relative measure of intensity. The bel (so named after Alexander Graham Bell) is defined as follows:

$$b = \log_{10} \frac{I_1}{I_2}$$

where I_1 and I_2 are the absolute intensities.

The bel is a large unit, an intensity ratio of 10 to 1 being equivalent to only 1 bel. This would mean dealing

[1]Fletcher, H., "Speech and Hearing," D. Van Nostrand, New York, 1929.

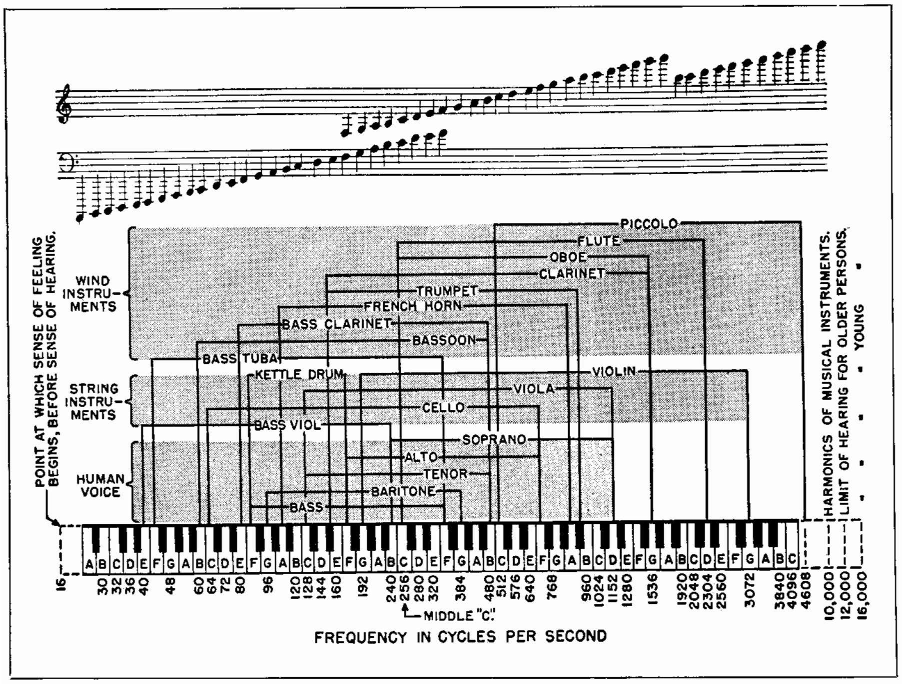


Fig. 1-6. Frequency range of musical instruments and those covering the human voice.

with rather small numbers, so it is customary to use the decibel, which is one tenth of a bel. Thus the numbers with which we deal are 10 times as large when speaking of bels. To convert to decibels, the value in bels must be multiplied by 10. Therefore:

$$\text{db.} = 10 \log_{10} \frac{I_1}{I_2}$$

An intensity ratio of 10 to 1 would thus be equivalent to 10 db., and 3 db. would be equivalent to an intensity ratio of 2 to 1.

On the other hand, the loudness of a tone is a function of its frequency and intensity and can be defined as the magnitude or stimulus it creates in the auditory system. This implies a relative measure of loudness and some *reference level* must be established for comparison. The tone emitted at 1000 cycles at 0 db. is taken as this level and the loudness of any particular tone is determined by adjusting the reference tone until it sounds as loud as the one being tested. The increase in intensity of the reference tone is interpreted in terms of loudness units. Fletcher and Munson have done much work along this line and their results have been published in the *"Journal of the Acoustical Society of America."* Obviously, a certain amount of human error is bound to be injected into such measurements; however, correlation by statistical methods has yielded some very valuable information.

CHAPTER 8

The Decibel (Volume and Power Level Meters)

A study of the decibel (db., vu, dbm.) its definition, how used, and how calculated.

● The measurement of gain in an amplifier or loss in a transmission line or attenuator circuit is usually expressed in db. (decibels) rather than in watts or volts. The reason for this is the fact that the human ear responds to intensities of sound logarithmically rather than linearly.

A sound having 10 watts of power is 10 times as loud to the ear as a sound of 1 watt or a sound of 100 watts is 10 times as loud as a sound of 10 watts but only 20 times as loud as a sound of 1 watt.

A sound of 1000 watts (1 kilowatt) is 20 times as loud as a sound of 10 watts or 30 times as loud as a sound of 1 watt. We must, therefore, use a "mental yardstick" having a logarithmic scale and not the usual rule.

The standard transmission unit is the decibel (abbreviated db.) and is equal to one-tenth of a bel. Nominally, it supersedes the *TU* (transmission unit), although specifying exactly the same thing as the latter term. The bel, named in honor of Alexander Graham Bell, was adopted by an international convention of telephone engineers.

The decibel is a logarithmic expression of a ratio between two quantities. As a unit of measurement, it specifies no definite amount of current, voltage, power, or sound but represents merely a ratio between two magnitudes. It is therefore a relative unit. Since the db. is a logarithmic unit, successive gains or losses expressed by it may be added algebraically.

The db. may express a ratio between two values of either current, voltage, power, or sound energy. It thus becomes possible to determine the db. gain for a given amplifier from ratios that express either voltage, current, or power amplification. Gain is expressed as *plus* db., loss as *minus* db.

While the decibel may show gain or loss with respect to the power at some point in a system, it properly has no regard for the finite value of any reference power. However, it is accepted practice in some radio and telephone measurements to designate the power level of 0.006 watt (6 milliwatts) as *zero* db. and to express any other value of power as a certain number of db. above or below this reference level. Hence, the expression "db. up" and "db. down." (See Fig. 8-1.)

Determination of Ratios

Ratios are common to everyday scientific ratings. They express relative superiority, or inferiority, gains or losses in a concise and readily understandable manner. We state that a certain transformer permits a voltage step-up of 2 to 1, etc. Such simple ratios might easily be converted into convenient statements of db. gain and db. loss.

POWER LEVEL (db.)	VOLTS 500 OHM LINE	POWER WATTS 0 db. = 6 mw.
—20	0.1730	0.00006
—19	0.1990	0.00007
—18	0.2180	0.00009
—17	0.2340	0.00011
—16	0.2730	0.00015
—15	0.3000	0.00018
—14	0.3390	0.00023
—13	0.3890	0.00030
—12	0.4450	0.00039
—11	0.4860	0.00047
—10	0.5477	0.00060
— 9	0.6145	0.00070
— 8	0.6895	0.00090
— 7	0.7737	0.00110
— 6	0.8681	0.00150
— 5	0.9740	0.00180
— 4	1.0928	0.00230
— 3	1.2262	0.00300
— 2	1.3758	0.00370
— 1	1.5437	0.00470
0	1.7321	0.00600
+ 1	1.9434	0.00750
+ 2	2.1805	0.00950
+ 3	2.4466	0.01190
+ 4	2.7451	0.01500
+ 5	3.0801	0.01890
+ 6	3.4559	0.02380
+ 7	3.8776	0.03000
+ 8	4.3507	0.03800
+ 9	4.8680	0.04740
+10	5.4772	0.06000
+11	6.1600	0.07590
+12	6.8100	0.09480
+13	7.7368	0.11970
+14	8.7100	0.15070
+15	9.7400	0.18970
+16	10.9285	0.23880
+17	12.2620	0.30070
+18	13.7578	0.37850
+19	15.4369	0.47660
+20	17.3205	0.60000
+21	19.4340	0.75530
+22	21.8050	0.95090
+23	24.4660	1.19710
+24	27.4510	1.50710
+25	30.8010	1.89740
+26	34.5590	2.3886
+27	38.7760	3.0071
+28	43.5070	3.7857
+29	48.8160	4.7660
+30	54.7720	6.0000
+31	61.4550	7.5535
+32	68.9540	9.5093
+33	77.3680	11.9716
+34	86.8080	15.0713
+35	97.4000	18.9747
+36	109.2850	23.8865
+37	122.6200	30.0710
+38	137.5820	37.8570
+39	154.3690	47.6600
+40	173.2050	60.0000
+41	194.3400	75.5350
+42	218.0500	95.0930
+43	244.6600	119.7160
+44	274.5100	150.7130
+45	308.0100	189.7470
+46	345.5900	238.8650
+47	389.0700	300.7100
+48	435.6000	379.5000
+49	487.0100	474.3700
+50	547.7200	600.0000
+51	616.0300	759.0000
+52	688.7400	948.7500
+53	770.4000	1185.9400
+54	871.2000	1518.0000
+55	974.0300	1897.5000

Impedance		DB.
4000 ohms	Subtract	9
2000 ohms	Subtract	6
600 ohms	Subtract	1
500 ohms	....	0
250 ohms	Add	3
24 ohms	Add	13
15 ohms	Add	15
10 ohms	Add	17
8 ohms	Add	18
6 ohms	Add	19
5 ohms	Add	20
4 ohms	Add	21
2.5 ohms	Add	23
2 ohms	Add	24

$E = \sqrt{WZ}$
where:
E = ac volts across load
Z = loaded line impedance.
W = watts.
Cond. in Series with db. meter = .25 mfd.

Fig. 8-1. Conversion Chart for 500 ohm db. meter across impedances from 2-4000 ohms.

The ratio of two values of power, for example, P_1 and P_2 is represented as:

$$P_1/P_2 \ldots\ldots\ldots\ldots (1)$$

Or, the larger power is divided by the smaller. It is easily seen that the actual instantaeous magnitudes of P_1 and P_2 might extend over a wide range of values but would be of no concern to the quotient as long as they remained in the same proportion.

The number of decibels represented by such a ratio is obtained by multiplying the logarithm of the indicated quotient by 10:

$$no.\ db. = 10 \log_{10} (P_1/P_2) \ldots\ldots (2)$$

Observe that the common logarithm of the quotient (ratio) is employed, i.e., the logarithm to the base 10. Hence the form, $\log_{10}$. From equation (2) the following rule may be stated:

Rule A: The number of db. is numerically equal to 10 times the common logarithm of a power ratio.

Voltage and current ratios may also be expressed in terms of decibels. If the two values of voltage E_1 and E_2 are measured across the same or identical impedances, or if the two values of current (I_1 and I_2) are taken through the same identical impedances:

$$no.\ db. = 20 \log_{10} (E_1/E_2) = 20 \log_{10} (I_1/I_2) \ldots\ldots\ldots (3)$$

Observe that in equation (3) the logarithm of the ratio is multiplied by 20 instead of 10. This is because power varies directly as the square of the current, and a logarithmic expression obtained in the manner of equation (2) needs to be multiplied again by 2, since doubling a logarithm is equivalent to squaring the number.

If, as is occasionally the case, the current or voltage values in the ratio are not associated with the same or identical impedances, our decibel computation must take into consideration the absolute magnitudes of the corresponding impedances and power factors of the impedances:

$$no.\ db. = 20 \log_{10} (E_1/E_2) + 10 \log_{10} (Z_2/Z_1) + 10 \log_{10} (f_1/f_2) \ldots\ldots (4)$$

and

$$no.\ db. = 20 \log_{10} (I_1/I_2) + 10 \log_{10} (Z_1/Z_2) + 10 \log_{10} (f_1/f_2) \ldots\ldots (5)$$

Z_1 and Z_2 are the impedances in which the voltages and currents operate, and f_1 and f_2 are the values of the corresponding power factors of the impedances.

From equations (3), (4), and (5) the following rules may be stated:

Rule B: When a voltage or current ratio shows values associated with the same or identical impedances, the number of db. is numerically equal to 20 times the common logarithm of the ratio.

Rule C: When a voltage ratio shows values associated with unequal impedances, the number of db. is numerically equal to a sum of three logarithmic expressions: 20 times the log of the voltage ratio plus 10 times the log of the impedance radio inverted plus 10 times the log of the power factor ratio.

Rule D: When a current ratio shows values associated with unequal impedances, the number of db. is numerically equal to the sum of three logarithmic expressions: 20 times the log of the current ratio plus 10 times the log of the impedance ratio plus 10 times the log of the power factor ratio.

Relationships

By definition, the common logarithm of a number is the exponent denoting the power to which 10 must be raised to equal the given number. Thus, 3 is the common log of 1000 since 10 must be raised to the third power to equal 1000, 5 is the common log of 100,000; 6 of 1,000,000.

Column 2 of Fig. 8-2 lists common logs corresponding to a few of the even-numbered ratios frequently encountered in radio work. The ratios are given in column 1. These logs are multiplied by 10 (in column 3) to give db. for power ratios, and by 20 (in column 4) to give decibels for current or voltage ratios.

It is readily seen from the chart that a power ratio of 100 to 1 corresponds to 20 db., while a current or voltage ratio of the same magnitude corresponds to 40 db. A ratio of 1,000,000 is equivalent to 60 db. for power, but 120 db. for current or voltage. It is also seen that a power, voltage, or current ratio must be squared in order to double

(1) RATIO	(2) LOG OF RATIO	(3) DB FOR POWER RATIO	(4) DB FOR CURRENT OR VOLTAGE RATIO
1	0	0	0
10	1	10	20
100	2	20	40
1000	3	30	60
10,000	4	40	80
100,000	5	50	100
1,000,000	6	60	120

Fig. 8-2. Common Log Table.

the number of decibels; and that increasing a power ratio to 10 times its original value is equivalent to adding 10 db., while the same increase in a current or voltage ratio is equivalent to adding 20 db.

It should be apparent to the reader that the same number of db. may be obtained from each of a number of ratios with widely divergent numerator and denominator values, as long as the same proportion exists between the two terms. A clear understanding of this condition not only explains why an extremely low-powered system can show the same number of db. gain as one of high power, but at the same time also places illustrative emphasis upon the relatively of the transmission unit.

Consider for example, three af amplifiers—one delivering 1 watt output with 0.1 milliwatt input; the second delivering 10 volts output for 100 millivolts input, both input and output circuits operating into 500 ohms impedance; and the third delivering 50 watts output for 5 milliwatts input. The gain in each case is 40 db.

Applications

1. *Amplifier Power Gain or Loss.* (a) Measure af or rf input watts, (b) measure af or rf output watts, (c) apply equation (2) and *Rule A*.

2. *Amplifier Voltage Gain or Loss.* Input and Output Impedances Equal. (a) Measure af or rf input voltage, (b) measure af or rf output voltage, (c) apply voltage equation (3) and *Rule B*.

3. *Amplifier Current Gain or Loss.* Input and Output Impedances Equal. (a) Measure af or rf input current, (b) measure af or rf output current, (c) apply current equation (3) and *Rule B*.

4. *Amplifier Current or Voltage Gain or Loss.* Unequal Input and Output Impedances. (a) Measure af or rf input voltage or current, (b) measure af or rf output voltage or current, (c) determine absolute values of input and output impedances, (d) determine absolute values of power factor for the two impedances, (e) apply voltage equation (4) and *Rule C* or current equation (5) and *Rule D*.

Note:—Any of the foregoing amplifier characteristics may be taken for the entire amplifier (over-all) a single stage (per stage) or any cascaded group of stages. A *plus db.* rating indicates gain; *minus db.*, loss.

5. *Amplifier Output Level, or AF Line Level.* This is stated by engineers and manufacturers as so many db., the number of decibels above or below 6 milliwatts (zero db.) being assumed. Apply the equation:

$$P \text{ in watts} = .006 \times \text{antilog } (\text{db.}/10) \qquad (6)$$

The term db. is the figure stated for amplifier or line. An *antilog* is the figure or number which corresponds to a certain log. (Looking up an antilog in the log tables is the reverse of the process of looking up a log.)

6. *Microphone, Hum, Input Signal or Noise Level.* This is generally stated as so many negative db., or "db. down," and like Example 5 refers to zero db. as 6 milliwatts. Since we are dealing with ratios only, we can work out the problem assuming a positive db. level to get the power ratio, then divide the reference level, 6 milliwatts, by this ratio to get the actual level. For example, a level of — 19 db. will be assumed for the purposes of calculation. Taking the antilog of (19/10) or 1.9 gives 79.4. The actual power level represented by — 19 db. would then be .006 divided by 79.4 or .000076 watts, or .076 milliwatts. This method simplifies calculation somewhat as it avoids the necessity of dealing with negative logarithms.

Use of the Slide Rule

Problems involving decibels are easily solved on the slide rule, making use of the *L* logarithm scale. Converting a power ratio to decibels is accomplished by setting the indicator to the power ratio on the *D* scale and reading the number of db. on the *L* scale. Example: the power ratio corresponding to 5 is 6.99 db. When the power ratio is higher than 10, divide by 10, 100, etc., until the quotient is less than 10. Find the corresponding db. and add 10 decibels for each place the decimal point had to be moved in order to bring the ratio within the range 1-10. Example: What is the db. gain corresponding to a power gain of 5530? Moving the decimal point three places to the left, we obtain 5.53. Set the indicator to 5.53 on the *D* scale and read 7.42 on *L*. Add 30 db. to the result, which gives 37.42 db.

If the power ratio is less than 1, the *CI* and *L* scale should be employed. Finding the db. gain corresponding to voltage ratios—if the impedances and power factors are the same in each case—proceed as described in the foregoing but multiply the result by 2.

The "log-log" slide rule offers an alternative method of finding decibels. Set the index on the slide to 10 on *LL*3. Opposite the power ratio on *LL*2 and *LL*3 find the gain in db. on *C*. If the power ratio was greater than 10, all values found on *C* are between 10 and 100. If it was less than 10, the *C* scale may be read directly.

Finding decibels from the voltage ratio is accomplished by setting 2 on the *C* scale and 10 on *LL*3. Opposite the voltage ratio on *LL*2 or *LL*3 the db. may be read on *C*. If the voltage or current ratio is between 1 and *pi*, the db. gain is between 1 and 10. If the ratio exceeded 3.14, multiply the *C* indication by 10.

When converting power ratios less than 1 to decibels, set 1 (middle of *B* scale) to 0.10 on the *LLO* scale. The loss is then found on *B* opposite the power ratio on *LLO*. For current or voltage ratios, set 2 on *B* scale to 0.10 on the *LLO* scale and proceed as before.

To find the gain in db. directly from two values of current, voltage, or power, set the larger of the two on *C* to the smaller on *D*. Opposite the index of *C*, find db. on *L*.

Ready-Reference Charts

For the reader's convenience, two charts have been made up from calculations involving equations (2), (3), and (6).[1]

By referring to Fig. 8-3, the number of decibels corresponding to any power level between 6 micromicrowatts and 6 kilowatts may be found quickly, the necessity for performing equation (6) computations being eliminated in most practical instances. From Fig. 8-4, the number of decibels corresponding to any current, voltage or power ratio may be quickly located.

Particular notice should be taken of the subdivisions in the power column of Fig. 8-3. These graduations are uniformly spaced (as regards numerical value) except that the lowermost subdivision in each power group has not the same value as each of the upper five in the group. For this reason, we have numbered the lowermost subdivision in each group. Thus, the number line, 10 micromicrowatts is only 4 micromicrowatts removed from the 6 micromicrowatts major division, while each other subdivision up to 60 micromicrowatts is exactly 10 micromicrowatts higher than the previous one. Thus, we read, 10, 20, 30, 40, and 50 micromicrowatts between 6 and 60 micromicrowatts. Similarly, we read, 100, 200, 300, 400 and 500 micromicrowatts in the next highest power group, between 60 and 600 micromicrowatts.

To illustrate the use of Fig. 8-3, locate the db. output rating of a 6L6 amplifier, the audio output of which is 60 watts. Opposite the 60 watt line in the power column will be found the 40 line in the decibel column. On the basis of 6 milliwatts as zero db., a power level of 60 watts is 40 db.

The power output of a high quality microphone rated at minus 45 db. may be found in the same manner. Read 0.2 microwatt directly opposite the "minus 45 db." in Fig. 8-3.

A current, voltage, or power ratio is

[1]The Aerovox Research Worker, Vol. 12, No. 7, July, 1941.

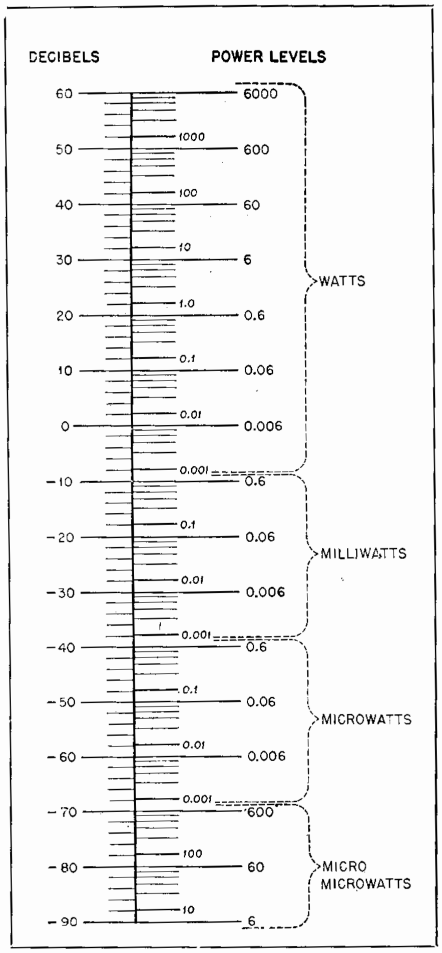


Fig. 8-3.

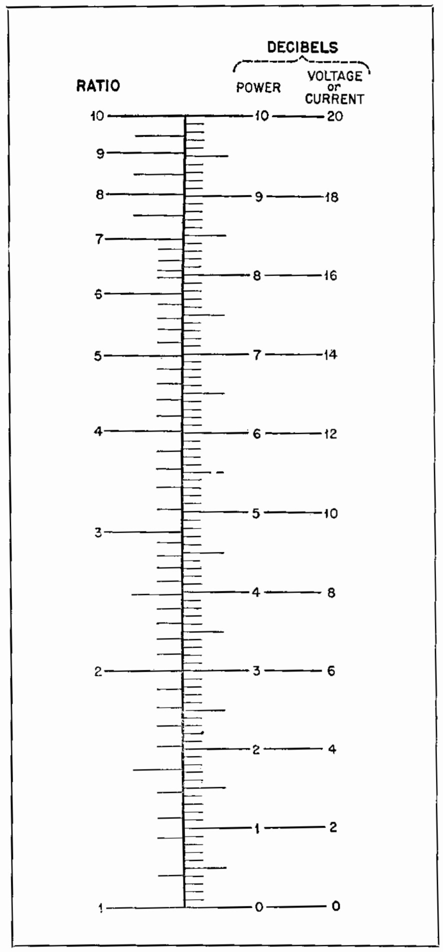


Fig. 8-4.

located in the ratio column of Fig. 8-4 and the number of decibels read directly opposite in the power column or current-voltage column, depending upon the nature of the ratio. For example: a power ratio of 4 is seen to correspond to 6 db. while a current or voltage ratio of the same value equals 12 db.

The use of Fig. 8-4 can be extended beyond the current or voltage ratio of 10 by adding 20 db. for each place the decimal point has been moved to the right to make the figures in the ratio column correspond to those in the ratio desired. For example: to find the db. equivalent to a current or voltage ratio of 44, locate 4.4 in the ratio column of Fig. 8-2. Read the equivalent 12.8 db. in the current-voltage db. column. The decimal point was moved one place in 4.4 to convert it into the ratio, 44. Therefore, add 20 db. to the result. 12.8 plus 20 equal 32.8 db.

The use of Fig. 8-4 may similarly be extended beyond the power ratio of 10 by adding 10 db. for each place the decimal point is moved to the right. For example: look up the power ratio 160 as 1.6 in the ratio column. This would correspond to 2 db. But the decimal point was shifted two places to change 1.6 to 160 and 10 db. must be added for each place. The result, therefore is 2 plus 20, or 22 db.

Power Level and Volume Indicator Meters

The measurement of power or program levels in sound circuits is of primary importance to the audio engineer. Meters designed for this purpose are known as "volume indicator (vi)" meters and have been in use by the telephone company, radio broadcasting stations, recording companies, and the motion picture industry for many years.[2]

The purpose of these meters, or volume indicators, is to indicate the variation in electrical sound levels. This does not mean that the meter measures the actual volume of sound, but rather the level of voltages generated by the reproducing equipment. Measurement by electrical means is made possible by the fact that the sound intensity is directly proportional to the voltage amplitude in an electrical circuit.

As the voltages generated are of an alternating nature, an ac type meter will be required for their measurement. The conventional type ac meter used for power measurement is not suitable for this purpose because of its characteristic low impedance and limited response to frequency.

Volume indicator meters are, in reality, sensitive high impedance ac voltmeters of wide frequency range, calibrated either in decibels or volume units relative to one of the two existing reference levels in use by the electronic industry. The device may be either electronic or of the copper-oxide rectifier type.

From a study of the waveforms involved in the transmission of speech and music, at first thought it appears that the measuring instrument should respond to the peak or crest of the wave, rather than the rms value. Further study, however, indicates that either type meter will be satisfactory. Also, it has been found that phase distortion affects the readings of peak indicating meters, although it may not be apparent to the ear, thus possibly causing overloading of the system because of incorrect readings.

Two terms are in general usage in regard to meters used for audio frequency measurements. They are the volume level indicator and the power level indicator. The first term applies only to meters used for monitoring program material of complex waveforms which is varying in both amplitude and frequency. The second term is applied to meters used for the measurement of sound circuits where steady-state conditions are maintained by using sine wave power.

Power Level Indicator

The principal difference between the volume level indicator and the power level meter is in damping. In the volume level indicator definite characteristics have been standardized by the industry and are built into the meter movement to control its sensitivity, damping, frequency range, and input impedance. In the power level meter, only the frequency range and sensitivity are of importance. Volume level meters may be used to measure both steady-state conditions and complex waveforms; however, this does not apply to power level meters.

In the early days of broadcasting and recording, the volume level indicator was of the electronic type, having characteristics which were considered by its designers to be the most suitable. This situation led to considerable confusion throughout the industry as measurements made in one connection could not be correlated with other measurements. In addition, the meter employed one or more vacuum tubes and required both a high and low voltage source for its operation. This feature restricted the use of these meters to the laboratory or studio control room.

Around 1929 the copper-oxide rectifier volume level indicator made its appearance. This meter employed a 1 ma. dc D'Arsonval type movement which utilized a half-wave copper-oxide recti-

[2] H. M. Tremaine, "Power Level and Volume Indicator Meters," Radio & Television News, Nov. 1950, pp. 84-96.

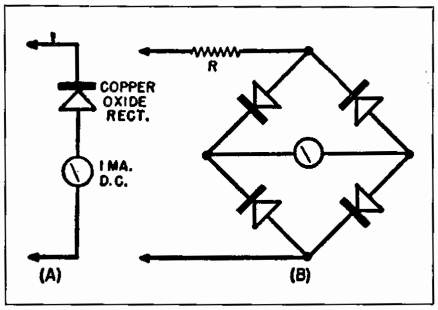


Fig. 8-5. (A) Half-wave rectifier and (B) a full-wave bridge rectifier.

fier connected in series with one side of the meter. See Fig. 8-5A. This resulted in a fairly sensitive meter which was handicapped, however, by a limited frequency response and poor damping. The use of the half-wave rectifier resulted in a non-symmetrical response to complex waveforms. Later the rectifier was changed to a full-wave bridge type, Fig. 8-5B, which eliminated the distortion introduced by the half-wave rectifier.

Full-wave rectification is essential in meters used for audio frequency measurements since the human voice consistently generates peaks of greater amplitude for one polarity than the other. Under these conditions it would be impracticable to use half-wave rectification unless the circuits were poled and the meter polarity was fixed. Furthermore, if a bi-directional microphone was in use and it was reversed 180 degrees, the peaks, as indicated by the meter, would also be reversed, and depending on the original polarity, might cause overloading of the electrical circuits due to the incorrect meter readings.

With the adoption of the full-wave rectifier, the meter sensitivity was increased by replacing the 1 ma. movement with one of 500 microampere sensitivity. A resistance was placed in series with the movement to increase the meter impedance to 5000 ohms and at the same time adjust its sensitivity to some definite reference level. When this meter was introduced, the reference level for audio frequency measurements was 6 milliwatts of power in a 500 ohm circuit. For this amount of power, 1.73 volts is developed across the circuit. The meter movement indicated this 1.73 volt reference point on the scale slightly to the left of its center position.

Power level meters are used for setting the levels of recording or reproducing systems and are generally connected in the circuit at a point where the signal is distributed to other portions of the system. This point is, as a rule, the bridging bus.

The power level indicator meter is used strictly for steady-state sine wave measurements and is not suitable for monitoring program material as its damping is such that the meter tends to overshoot its mark, thus giving rise to false indications with changing levels.

Three types of meter movements are available for power level indicators: 1. the "high speed" or HS movement used for high speed measurements; 2. the "general purpose" or GP movement work; and 3. the "low speed" or SS movement which is generally employed in making acoustic measurements. The different types of movements may be identified by the symbols "GP," "SS," or "HS" which appear on the meter face.

To increase the operating range of the meter movement above the reference level, a 5000 ohm "L" type attenuator is placed ahead of the meter. See Fig. 8-6. The attenuator performs two functions, i.e., it extends the range above the reference level, and permits this extension without affecting the

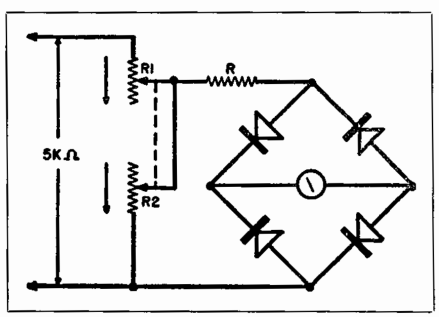


Fig. 8-6. Circuit diagram of a rectifier-type meter with an "L" type variable attenuator.

5000 ohm input impedance. A constant input impedance is essential as it allows the input attenuator setting to be changed during operation without upsetting the impedance of the circuit bridged by the meter.

The "L" type attenuator does not present a constant impedance looking in both directions, but only to the source bridged. The arms R_1 and R_2 are connected mechanically by the common shaft and are varied inversely. Therefore, R_2 may drop to only a few ohms when the attenuator is set to the higher ranges. The loss potentiometer should be thought of as a meter multiplier. For the power level meter, the settings of the attenuator usually start at zero and continue in steps of 1 or 2 db. to at least +30 db. above the reference level. The meter face is calibrated from —10 db. to +6 db. Adding the meter dial readings and the attenuator settings algebraically results in a range of from —10 db. to + 36 db., or a total spread of 46 db.

Copper-oxide instrument rectifiers are approximately 70% efficient and are excellent for rectification at audio frequencies, providing the voltage across any one element in the bridge does not exceed 11 volts and that the temperature does not rise above 160 degrees F. With the attenuator placed ahead of the meter, the voltage across the rectifier elements, under normal operating conditions, does not exceed 4 volts.

The frequency response of the early instrument rectifiers was deficient at frequencies above 6000 cycles, dropping off as much as 4 db. at 10,000 cycles. At first this was attributed to the capacity between the plates of the rectifying elements, but subsequent investigation proved that this was not entirely the case. While it is true that some effect was contributed by the capacity, the major factor was the rectifying qualities of the elements at the higher frequencies.

Present day instrument rectifiers are highly developed and will respond to several hundred thousand cycles-per-second. The frequency response of a typical copper-oxide rectifier is shown in Fig. 8-7.

The VU Meter

Because of the inaccuracies inherent in the copper-oxide rectifier power level meter and because of the fact that it was not satisfactory for program monitoring, the development of an entirely new meter was jointly undertaken by the *Bell Telephone Laboratories, Columbia Broadcasting System,* and the *National Broadcasting Company.* The result of this research was the development not only of a new type volume indicator meter but also a new reference level of 1 milliwatt, a unit which was adopted by the electronic industry in May 1939.

This new meter, shown in Fig. 8-8, was termed a "volume unit (vu)" meter. This instrument is calibrated in volume units numerically equal to the number of decibels above the reference level of 1 milliwatt of power in a 600 ohm circuit. With this new meter came a new term, "zero dbm.," meaning that the level under discussion is in relation to 1 milliwatt of power in a 600 ohm cir-

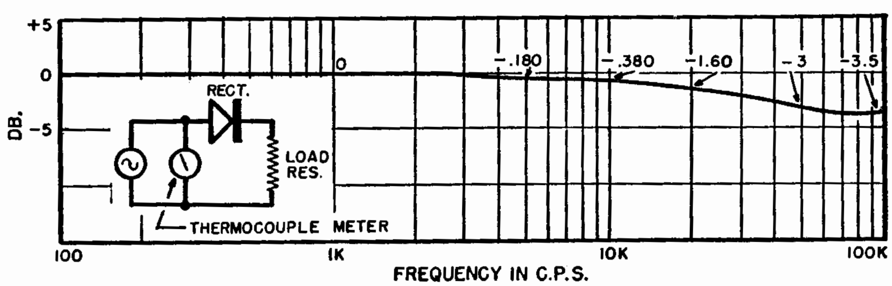


Fig. 8-7. Frequency response of a conventional copper-oxide instrument rectifier.

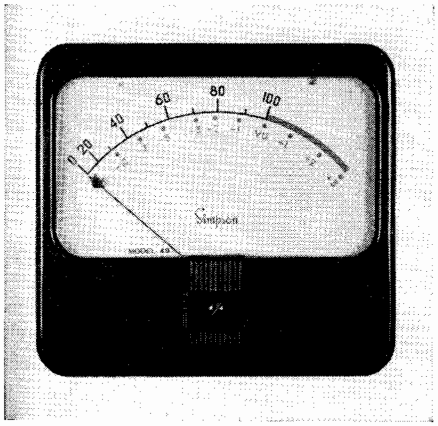


Fig. 8-8. Standard VU meter, Simpson 49, with B scale. (Courtesy, Simpson Elect.)

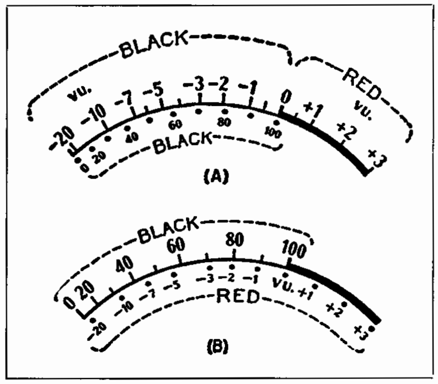


Fig. 8-9. (A) Standard VU type A scale stressing VU and (B) Standard VU scale type B stressing modulation.

cuit. In discussing reference levels, the term "dbm." applies only to the 1 milliwatt reference level, while "db." is used to express levels referring to the 6 milliwatt reference level.

The 1 milliwatt reference level has three distinct advantages. It is a unit quantity, hence it is readily applicable to the decimal system, being related to the watt by the factor 10^{-3} which results in positive values for the majority of measurements. A further advantage of the vu meter is that all meters of this type are exactly alike in construction and characteristics and, when several are connected across the same circuit, may be tested by the application of a 1000 cycle signal for checking their operation.

VU Meter Characteristics

The characteristics for the vu meter were adopted after careful consideration as being most suitable for all applications.

1. General. The volume indicator employs a dc meter movement with a non-corrosive, full-wave, copper-oxide rectifier unit, and responds approximately to the rms value of the impressed voltage. This will vary somewhat depending on the waveforms and the per-cent harmonics present in the signal.

2. Scale of instrument. The face of the instrument may have either of the two scale cards shown in Fig. 8-9. Each card has two scales, one a vu scale ranging from —20 to +3 vu, and the other a per-cent voltage scale ranging from 0 to 100 with the 100 point coinciding with the 0 point on the vu scale. The normal point for reading volume levels is at the 0 vu or 100 scale point which is located to the right of the center at about 71% of the full-scale arc.

3. Dynamic characteristics. With the instrument connected across a 600 ohm external resistance, the sudden application of a sine wave voltage sufficient to give a steady-state deflection at the 0 vu or 100 scale point, should cause the pointer to overshoot not less than 1% nor more than 1.5% (.15 db.). It should be capable of reaching 99 on the per-cent voltage scale in .3 second.

4. Response vs. frequency. The instrument sensitivity should not depart from that at 1000 cps by more than .2 db. between 35 and 10,000 cps nor more than .5 db. between 25 and 16,000 cps.

5. Sensitivity. The application of a sinusoidal potential of 1.228 volts (4 db. above 1 milliwatt in 600 ohms) to the instrument, in series with the proper external resistance (3600 ohms),

will cause a deflection to the 0 vu or 100% point.

6. Impedance. For bridging across a line the volume indicator, including the instrument and proper series resistance (3600 ohms), should have an impedance of 7500 ohms when measured with a sinusoidal voltage sufficient to deflect the meter to the 0 vu or 100% scale point.

7. Harmonic distortion. The harmonic distortion introduced in a 600 ohm circuit when a volume indicator is bridged across it is less than .3% under the worst possible condition (when there is no loss in the variable attenuator).

8. Overload. The instrument must be capable of withstanding, without injury or effect on the calibration, peaks of ten times the voltage equivalent to a reading of 0 vu or 100% for .5 second, and a continuous overload of five times that giving a reading of 0 vu.

The selection of the scale for the meter face will depend on its application. For broadcasting and recording purposes where it is desirable to know the percentages of modulation the scale of Fig. 8-8 is used. For laboratory and general test work the scale of Fig. 8-9A is employed because most of the readings will be in vu.

The circuit diagram of the complete vu meter is shown in Fig. 8-10. The meter movement consists of a 50 microampere dc D'Arsonval movement and a full-wave bridge rectifier. The meter impedance, including the rectifier, is 3900 ohms.

A 3900 ohm variable attenuator and a 3600 ohm series resistor are connected ahead of the meter. As a rule, devices which are to be bridged across circuits of 600 ohms should have at least ten times the impedance of the circuit being bridged. Connecting the 3600 ohm resistor ahead of the attenuator provides a bridging impedance of 7500 ohms. See Fig. 8-10.

Attenuators designed for use with vu meters are constructed with the attenuator dial calibrations starting at +4 dbm. and continuing in steps of 2 db. up to +44 dbm. No 0 dbm. position is provided. The reason for this is that by placing a 3600 ohm resistor ahead of the attenuator to raise the input impedance, a loss of 4 db. is incurred.

If the 3900 ohm meter movement and the attenuator (the attenuator in its +4 dbm. position has zero loss) is placed across a 600 ohm circuit in which 1 milliwatt of power is flowing, the pointer will be deflected to the 100% or 0 dbm. mark. When the 3600 ohm series resistor is inserted in the circuit the sensitivity of the meter is lowered by 4 db. To bring the deflection back to the 100% mark will now require +4 dbm. at the input.

It is the practice of the telephone company to allow signal levels from +4 dbm. to +8 dbm. to be transmitted over the average cable pair. Thus, with the meter connected as shown in Fig. 8-10 the 100% mark will indicate a +4 dbm. signal level.

The advantage to be obtained by the use of the vu meter over the power level and older type v.i. meters are: a 7500 ohm bridging impedance; the 100% or 0 dbm. mark may be set to represent a maximum signal level permitted by the telephone company and if associated with a line feeding a radio transmitter or recording system, it will indicate the percentage of modulation; uniform damping; and frequency response which allows the correlation of measurements with other activities.

An important point to remember in using the attenuator is that the zero calibration point on the meter always

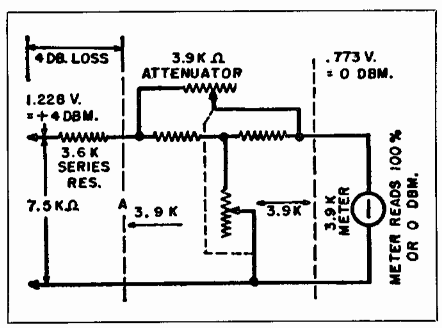


Fig. 8-10. Standard attenuating network for use with NARTB meter.

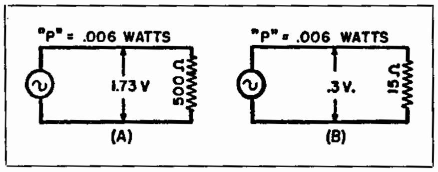


Fig. 8-11. Comparison of voltages in a 500 ohm and 15 ohm circuit for the same power.

becomes that of the attenuator setting and the meter indications are in reference to that setting. The greatest accuracy for either the power level or volume indicator meter lies between the —1 db. and the +1 db. calibration points on the meter scale. Thus, all readings should be made between these points, if possible, by adjusting the input attenuator and then adding the readings algebraically.

Power Level Meters

Power level meters are generally calibrated in reference to 6 milliwatts of power in a 500 ohm circuit and if placed across an impedance other than 500 ohms will not read correctly. To obtain the correct reading a "correction factor" is applied to the indicated reading to obtain the true power in the circuit. Fig. 8-11 demonstrates why the meter does not read correctly.

In the diagram a 500 ohm and a 15 ohm circuit are shown. It will be noted that the same amount of power is being dissipated in both circuits; however, the voltage across the circuits is not the same. If a power level meter, calibrated for 500 ohms, is placed across the 500 ohm circuit, it will be deflected to its zero reference mark. Connecting the same meter across the 15 ohm circuit will cause it to read low by 15.22 db. This is true because for the same amount of power the voltage across the 15 ohm circuit is only .3 volt. Thus the meter will not be deflected the same amount as for the 500 ohm circuit where 1.73 volts may be obtained.

CORRECTION FACTORS		
LINE Z (in ohms)	METER CAL. (500 ohms) (in db.)	METER CAL. (600 ohms) (in db.)
600	—0.791	0.000
500	0.000	+0.791
250	+3.01	+3.80
200	+3.97	+4.77
125	+6.02	+6.81
100	+6.99	+7.78
50	+10.00	+10.79
30	+12.22	+13.01
15	+15.22	+16.02
8	+17.96	+18.75
4	+20.97	+21.76

Fig. 8-12. Some commonly used correction factors for power level meters.

The correction factor for any impedance may be obtained with the aid of the following equation: db. correction $= 10 \log_{10} (Z_1/Z_2)$ or $(10 \log_{10} (Z_2/Z_1)$ if the impedance of the circuit is higher than that for which the meter was calibrated) where Z_1 is the impedance for which the meter is calibrated and Z_2 the impedance to be bridged.

As an example, assume that a db. meter calibrated for 500 ohms is connected across a line impedance of 8 ohms with its attenuator set to zero. The signal deflects the pointer to the zero mark. What is the true power level of the circuit? It is db. $= 10 \log_{10} (Z_1/Z_2) = 10 \log_{10} (500/8) = 10 \log_{10} 62.5 = 10 \times 1.796 = 17.96$ db.

The correction factor 17.96 db. is added to the meter reading, thus when the meter indicates zero the true level is +17.96 db. If the meter is placed across an impedance higher than that for which it was originally calibrated, the second equation is used and the correction factor is subtracted from the meter reading. The most commonly used correction factors are given in Figs. 8-1 and 8-12.

When a meter is bridged across a circuit a slight power loss takes place. This is called "bridging loss." To determine how much the circuit level will be affected when the meter is bridged across it, the following equation may be used: db. $= 20 \log_{10} [(2BR + R)/2BR]$ where BR is the meter impedance and R is the impedance of the circuit bridged.

For example, assume a v.i. meter of 5000 ohms is bridged across a circuit of 500 ohms. What is the bridging loss in decibels? It is db. $= 20 \log_{10} [(2BR + R)/2BR] = 20 \log_{10} [(10{,}000 + 500)/10{,}000] = 20 \log_{10} 1.05 = 20 \times .0212$

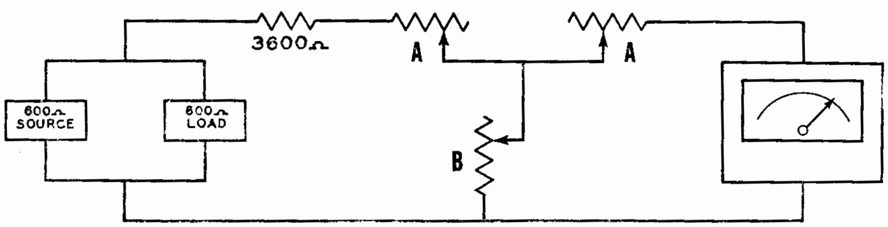
= .424 db. Thus, when the meter is bridged across the circuit the original level will be lowered by .424 db. Generally speaking, unless very precise measurements are to be made, this small loss in level may be ignored. Bridging losses for lines of different impedances and v.i. meters of various bridging impedances are shown in Fig. 8-14.

The difference in decibels between any two reference levels may be determined by the equation db. $= 10 \log_{10} P_1/P_2$. For example, the difference between a reference level of 1 milliwatt and one of 6 milliwatts would be: db. $= 10 \log_{10} P_1/P_2 = 10 \log_{10} 6/1 = 10 \log_{10} 6 = 10 \times .778 = 7.78$ db.

Conclusion

In closing it might be well to mention that copper-oxide rectifier meters, irrespective of type, will at times introduce distortion into the circuit being bridged, if used in conjunction with distortion measuring equipment. The vu attenuator should be set to its maximum position after the proper level has been obtained and then the distortion reading should be taken. Specifications for an attenuation network for the

ATTENUATION NETWORK FOR VOLUME LEVEL INDICATOR



Attenuator Loss — db.	Level vu*	Arm A Ohms	Arm B Ohms	Attenuator Loss — db.	Level vu*	Arm A Ohms	Arm B Ohms
0	+ 4	0	Open	24	+28	3487	494.1
1	+ 5	224.3	33801	25	+29	3485	440.0
2	+ 6	447.1	16788	26	+30	3528	391.9
3	+ 7	666.9	11070	27	+31	3566	349.1
4	+ 8	882.5	8177	28	+32	3601	311.0
5	+ 9	1093	6415	29	+33	3633	277.1
6	+10	1296	5221	30	+34	3661	246.9
7	+11	1492	4352	31	+35	3686	220.0
8	+12	1679	3690	32	+36	3708	196.1
9	+13	1857	3166	33	+37	3729	174.7
10	+14	2026	2741	34	+38	3747	155.7
11	+15	2185	2388	35	+39	3764	138.7
12	+16	2334	2091	36	+40	3778	123.7
13	+17	2473	1838	37	+41	3791	110.2
14	+18	2603	1621	38	+42	3803	98.21
15	+19	2722	1432	39	+43	3813	87.53
16	+20	2833	1268	40	+44	3823	78.01
17	+21	2935	1124	41	+45	3831	69.52
18	+22	3028	997.8	42	+46	3839	61.96
19	+23	3113	886.3	43	+47	3845	55.22
20	+24	3191	787.8	44	+48	3851	49.21
21	+25	3262	700.8	45	+49	3857	43.86
22	+26	3326	623.5	46	+50	3861	39.09
23	+27	3384	555.0				

*vu—Numerically equal to number of db. above 1 mw reference level.

Fig. 8-13. Circuitry and specifications for varible attenuators used with standard volume level indicators.

meter of Fig. 8-8 is shown on the chart Fig. 8-13.

The distortion introduced by the vu meter is caused by the rectifier clipping the peaks of the impressed waveforms. The amount of distortion introduced will depend on how much isolation is provided by the attenuator network resistance.

BRIDGING LOSSES		
METER Z (in ohms)	LINE Z (in ohms)	BRIDGING LOSS (in db.)
5000	500	0.42
5000	600	0.51
7500	500	0.28
7500	600	0.34
10,000	500	0.21
10,000	600	0.26

Fig. 8-14. Bridging losses for lines of different impedonces and vi meters of various bridging impedances.

Another precaution which should be taken is that since the vu meter is designed to be mounted on dural or aluminum panels, because of the high coercive force magnet used in the movement, the shunting effect of the steel on the magnetic circuit reduces the flux and affects the meter characteristics if the meters are mounted on steel panels.

CHAPTER 16

Dividing Networks and Filters

Design data covering series and parallel connected filter type networks and constant resistance networks for use with audio amplifiers.

Introduction

● In multiple channel systems the characteristics of the dividing network are as important as any other link in the chain of units. Simple arrangements use the inductance of the low frequency voice coil as well as the large mass of a relatively heavy cone as the low pass filter, and the high frequency section is fed with a series capacitance. More complex arrangements include inductance/capacitance networks for both sections. The characteristics may vary from a very slow roll-off with a great deal of overlap to cut-offs as sharp as 18 decibels per octave with very little overlap in commercially available units. A sharper cut-off than this is likely to introduce transient distortion and is found in practice to be unnecessary. The selection of crossover frequency is dictated by many considerations. If the crossover is low, then (a) the power handling requirement of the high frequency section is increased; (b) the length of the high frequency horn must be increased proportionately; and (c) very large capacitors are required for the high frequency section. If the cross-over is high, then (a) the low frequency speaker must handle a wider range with the probability of reaching into the region where the response is not smooth, and (b) if a low frequency horn is used, absorption in the walls may cause a droop in response ahead of cross-over.

In home installations where music only is to be reproduced, the division between speakers does not usually introduce a directional problem of consequence, nor is there difficulty from this source where the speaker units are designed with satisfactory coaxial orientation. However, where speech is to be satisfactorily reproduced from radio broadcast or other signal sources, it is undesirable to place the cross-over point within the central portion of the speech spectrum. This is one reason why some systems that sound excellent in reproducing music have undesirable characteristics in reproducing speech. The same kind of fuzziness and lack of presence may be observed under these conditions with sharply percussive sounds. This phenomenon is observed quite commonly but the source of trouble is often not recognized.

Where space is of relatively little consequence and a satisfactorily long high frequency horn may be used, it is probably desirable to place the cross-over point below 400 cycles per second. Another approach to this problem, of course, is to use a three-way system so that the mid-frequency range is handled by a single unit, eliminating the problem mentioned above in connection with speech and percussion reproduction and permitting optimum operation of both extreme high and low frequency units. This is not intended to indicate that very high quality results cannot be obtained with cross-over points selected in the mid-frequency range for two-way systems, but to emphasize the fact that optimum conditions satisfying all of the requirements are very difficult to obtain.

With some types of cross-over networks, where feedback is taken from the secondary of the output transformer, a sufficiently large reactive load may be introduced to cause serious instability in the feedback circuits resulting in oscillation. While this can be eliminated by careful design of the feedback circuits, it almost certainly is an indication of other effects that are undesirable though not so apparent. Many commercial crossover networks have tone control arrangements, usually for adjustment of the high frequency response. In some designs this is a roll-off control while in others it introduces a resistive loss that lowers the drive to the high frequency section. The roll-off type of control often involves networks that present an undesirable type of load for the amplifier. The stepped type of control may be desirable in some installations, but it is important that the user realize the characteristics of this system. A tone control in an amplifier that simply lifted or lowered an entire section of the spectrum would be considered entirely unsatisfactory. Such a control puts a distinct step in the response curve. Where the control is used to compensate for differences in efficiency between the high and low frequency sections of a system, it may be of real advantage but it should not be considered or used as a conventional tone control.

The efficiency of loudspeaker systems is a consideration of greater importance than is always recognized. It is all too common a concept that electrical power is cheap and that for this reason the efficiency of a loudspeaker system is not of first order importance. Electrical power within distortion limits satisfactory for high quality reproduction is by no means so inexpensive that it is of no importance. It is also true that the characteristics of power amplifiers are such that power output is increased in fairly substantial steps, and in-between ratings are not practical. Thus, 10 watts, 15 watts, 30 watts, etc., are relatively common ratings. If 15 watts is not quite enough to allow adequate reserve power for peak passages, it generally means going to 30 watts. Note that increasing the electrical drive to a single speaker from fifteen to thirty watts amounts to only approximately 3 decibels. Under normal listening conditions this will be observed as only a small increase in loudness. Usually it will be found that two fifteen watt amplifiers driving separate loudspeakers will result in greater efficiency and a larger increase in apparent loudness. With this in mind it becomes important that a dividing network introduce a minimum of loss. The insertion loss of practical systems may be kept to one decibel without great difficulty, and it is well worthwhile to reduce it to the practical limit of half a decibel in most systems. One decibel sounds deceptively small, but when it is translated into actual watts it will be noted that it makes a large difference in the power required from the amplifier.

Dividing Networks

Many audio systems make use of separate high-frequency and low-frequency loudspeakers. In order to obtain maximum efficiency from this dual reproducing arrangement, *dividing networks* are connected between the amplifier output transformer and the voice coils of the tweeter (high-frequency speaker) and woofer (low-frequency speaker). See Fig. 16-1. These networks separate the frequency components in

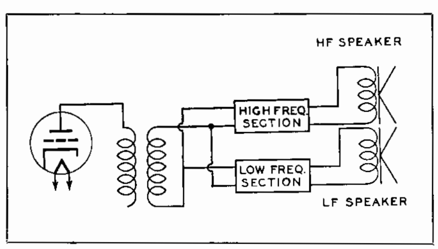


Fig. 16-1. Method for inserting dividing network between secondary of amplifier output transformer and loudspeaker voice coils.

the amplifier output voltage into two bands, so that only frequencies above a certain *crossover* frequency are transmitted to the tweeter, and only those below this crossover frequency are transmitted to the woofer. Each speaker thus operates only at those frequencies at which it is most efficient and faithful.

The crossover frequency may be selected at will, but most commercially available dividing networks operate at crossover frequencies of 400 to 2000 cycles. The basic facts concerning practical dividing networks may be summed up in the following brief comments:

(1) Each such network comprises a low-pass and high-pass filter with their input circuits connected either in series or in parallel. The output circuit of the high-pass filter section feeds the tweeter; that of the low-pass filter, the woofer.

(2) At the crossover frequency, the high and low-frequency power outputs are equal.

(3) With respect to the attenuation at the crossover frequency, the dividing network should provide 12 db minimum attenuation one octave from the crossover frequency.

(4) The *constant-resistance* type of dividing network is a specific form which, when terminated in the proper resistance load, will offer a constant input resistance over a frequency band. The constant-resistance type network is convenient in some instances, since each of its capacitive components are identical in value, as are each of its inductive components.

Circuit diagrams of dividing networks are given in Figs. 16-2 and 16-3, together with the formulas for obtaining the values of their capacitive and inductive elements. The arrangements shown in Fig. 16-2 are the conventional series and parallel-connected filter-type networks. Those given in Fig. 16-3 are constant-resistance networks. In the latter groups, two of the circuits (A and C) will provide only about 6 db attenuation at 1 octave from the crossover frequency, and should be employed only in those specific cases where this low attenuation may be tolerated.

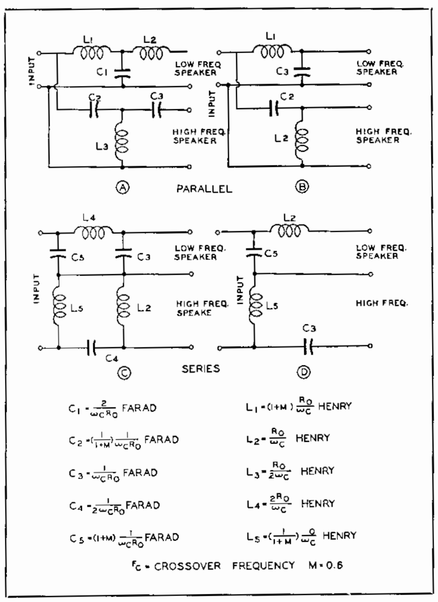


Fig. 16-2. Conventional series and parallel connected filter type networks along with the formulas for obtaining LC component values.

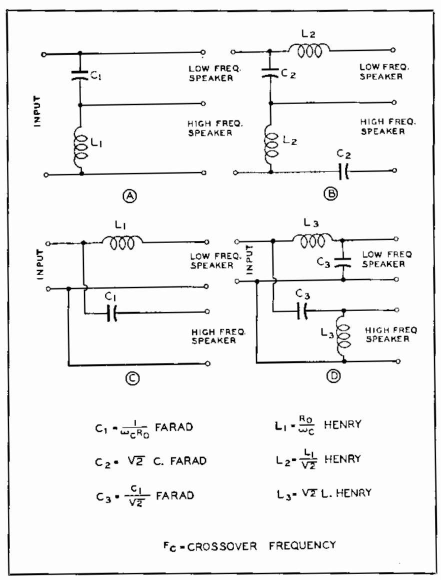


Fig. 16-3. Four types of constant-resistance networks and associated formulas for computing values. See note in text on A and B.

Fc =	400	500	600	700	800	900	1000	1100	1200	1300	1400	1500	1600	1700	1800	1900	2000
C1	79.62	63.70	53.08	45.48	39.80	35.38	31.85	28.95	26.54	24.49	22.74	21.23	19.90	18.75	17.69	16.76	16.59
C2	24.88	19.91	16.59	14.21	12.44	11.06	9.95	9.04	8.78	7.65	7.10	6.68	6.22	5.86	5.53	5.23	4.97
C3	39.81	31.85	26.54	22.74	19.90	17.69	15.91	14.47	13.27	12.24	11.37	10.61	9.95	9.37	8.84	8.38	7.95
C4	19.90	15.92	13.27	11.37	9.95	8.84	7.96	7.23	6.63	6.12	5.68	5.30	4.97	4.68	4.42	4.19	3.98
C5	63.69	50.96	42.46	36.38	31.84	28.30	25.46	23.16	21.23	19.59	18.19	16.98	15.95	15.00	14.15	13.41	12.73
L1	6.37	5.10	4.25	3.64	3.18	2.83	2.55	2.31	2.12	1.96	1.82	1.70	1.59	1.50	1.41	1.34	1.27
L2	3.98	3.18	2.65	2.27	1.99	1.77	1.59	1.44	1.32	1.22	1.13	1.06	.995	.93	.88	.84	.79
L3	1.99	1.59	1.33	1.14	.995	.884	.796	.725	.665	.61	.57	.530	.497	.468	.442	.419	.398
L4	7.96	6.37	5.31	4.55	3.98	3.54	3.18	2.89	2.65	2.45	2.27	2.12	1.99	1.87	1.77	1.62	1.59
L5	2.49	1.99	1.66	1.42	1.24	1.11	.995	.905	.830	.765	.710	.660	.620	.580	.550	.502	.497

C in μfds L in mh m = 0.6 Ro = 10 ohms F in cycles

Fig. 16-4. Conventional dividing networks. See Figure 16-2 for circuit.

Fc =	400	500	600	700	800	900	1000	1100	1200	1300	1400	1500	1600	1700	1800	1900	2000
C1	39.80	31.85	26.54	22.74	19.90	17.69	15.91	14.47	13.27	12.24	11.37	10.62	9.95	9.37	8.84	8.38	7.95
C2	56.29	45.03	37.53	32.15	28.14	25.01	22.50	20.46	18.76	17.31	16.07	15.01	14.07	13.25	12.50	11.84	11.25
C3	28.14	22.51	18.76	16.07	14.07	12.50	11.25	10.23	9.38	8.65	8.03	7.50	7.03	6.67	6.25	5.92	5.62
L1	3.98	3.18	2.65	2.27	1.99	1.77	1.59	1.44	1.32	1.22	1.13	1.06	.79	.93	.88	.84	.79
L2	2.81	2.25	1.87	1.60	1.41	1.25	1.12	1.02	.93	.86	.80	.75	.56	.66	.62	.59	.56
L3	5.62	4.50	3.74	3.20	2.82	2.50	2.24	2.04	1.87	1.73	1.60	1.50	1.12	1.32	1.25	1.19	1.12

C in μfds L in mh. Ro = 10 ohms f in cycles

Fig. 16-5. Constant-resistance dividing networks. See Figure 16-3 for circuit.

Position of Network in Amplifier

The band-separating action of the dividing network might be obtained at several points in a conventional audio amplifier. In standard practice, however, the dividing network is almost always connected between the secondary winding of the amplifier output transformer and the loudspeaker voice coils, as shown in Fig. 16-1. In this way, one output amplifier stage is made to serve both loudspeakers.

Each network section must carry the full power delivered to the loudspeaker which it supplies. Network components accordingly must be capable of handling these power levels safely. The condensers used may be low voltage types however, as the voltage across them never exceeds 25 volts. Electrolytic condensers should never be used. At the same time, the resistance of the inductors must be of the lowest possible value, consistent with required inductance, in order to minimize insertion losses.

The impedances out of, and into which the practical dividing network operates are identical with the rated voice coil impedance of the speakers. Thus the *Ro* value that appears in the formulas.

Use of Tables

All component values for dividing networks may be calculated by means of the formulas given in Figs. 16-2 and 16-3. However Figs. 16-4 and 16-5, which list these values calculated with sufficient accuracy for critical applications, are included herein for the reader's convenience.

Figs. 16-4 and 16-5 list all condenser and inductor values required, respectively, in conventional and constant-resistance type dividing networks. These tables are based upon an *Ro* value of 10 ohms and an *m* of 0.6. All capacitance values are given in microfarads and all inductance values in millihenries, for common crossover frequencies every 100 cycles from 400 to 2000 cycles.

When working with systems in which $Ro = 10$, all *C* and *L* values may be read in the corresponding frequency column directly from Fig. 16-4 for conventional networks, or from Fig. 16-5 for the constant-resistance type. For *Ro* values other than 10, the chart values may be operated upon to yield values required for the new impedance, thus, for a value (*Rx*) other than *Ro* (10 ohms) multiply all *L* values corresponding to the desired crossover frequency by *Rx*/*Ro*, and divide all *C* values by this same factor.

As an illustration of the use of the *Rx*/*Ro* factor, consider the following example: A conventional dividing network is required to work between 16 ohms at a crossover frequency of 1000 cycles. At 16 ohms, $Rx/Ro=16/10=1.6$. All *L* values in the 1000 cycle column of Fig. 16-4 must be multiplied by 1.6, and all *C* values in the same column must be divided by 1.6.

Design of Audio Networks

The use of electrical networks to give desired amplitude-frequency characteristics has become an extremely important part of electronic engineering. Since in general the uncorrected frequency-response characteristic of any given electrical circuit may not necessarily be the one which best gives the desired result, the method of frequency-response correction is seen to be of considerable importance in the design and construction of such equipment.

For example, considerable expense and difficulty may often be avoided in the design and construction of amplifiers by compensating for deficiencies in the over-all frequency response by means of an attenuation equalizer, rather than by attempting to make the amplifier perfect within itself. In a complete communications channel any component can be compensated for, and any desired frequency response may be attained by the

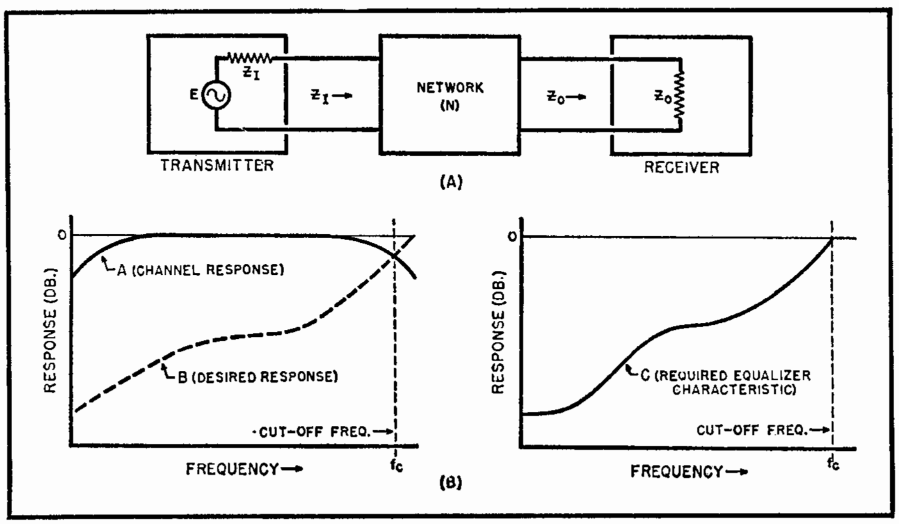


Fig. 16-6. (A) Generalized communication circuit including a 4-terminal network. (B) Frequency response curves showing required equalization to obtain desired response.

insertion of an equalizing network at some point in the channel. In transmission lines which are used to convey intelligence, the transmission losses are generally greater at higher frequencies, resulting in frequency distortion of the transmitted signal. To obtain faithful reproduction of the original signals at the receiving terminal, the transmission line is compensated by a correction network, generally at the receiving end, to make the resultant loss invariable with frequency over the required band. In the reproduction of sound, many different effects can be obtained by changing the amplitude relations of the different frequency components of the desired signal.

In addition to the extreme importance of network theory in audio-frequency work, it also finds considerable application at higher frequencies and particularly in ultra-high-frequency work at the present time. The same basic principles apply at uhf as at audio frequencies, and in addition, devices which depend in part on currents of uhf may have in them elements or sections which operate at much lower frequencies, as well as with direct current. Thus, much of audio-frequency network theory is of interest to many others besides audio-frequency engineers and technicians.

For some time there has been a need for a complete and concise practical treatment of network design from the viewpoint of the practical engineer who may not be a specialist in network theory, but is called upon to design networks for specific purposes. Some topics of network theory can only be found in theoretical books on the subject, while many others are widely scattered in the periodical literature, and still others have not been written about at all and are gained only by experience. This text will attempt to present, in a concise and unified form, a summary of the principles and design procedures of audio-frequency networks for the practical engineer and technician, with the actual design procedure reduced to a set of charts and curves wherever possible.

The general type of network which will give a desired frequency-response characteristic is the four-terminal network having two input terminals for connection to a transmission system to receive power, and two output terminals for connection to a load to deliver power. Such a network consists of an orderly array of two-terminal electrical elements whose arrangement is such as to produce a specified insertion-loss characteristic when connected between the proper terminal impedances. The two-terminal elements are made up of inductance, capacitance, and resistance in various combinations according to the function which the network is to perform. In wave filters, whose function is to let pass the desired frequency bands and to highly attenuate neighboring undesired frequency bands, and which therefore have frequency characteristics which vary rapidly near the cut-off frequency with no attentuation loss within the transmission band, the circuit elements are purely reactive. In frequency-response correction equalizers, whose frequency characteristics are desired to vary in a gradual manner with frequency, resistance as well as reactance elements are used in order to give a gradual variation of attenuation loss within the transmission band.

An important set of relations which arise in network theory are the equivalent and the inverse relationships between various two-terminal impedances. Two networks which have identical impedances at their terminals for any frequency (even though the circuit arrangements and element values may be different) are said to be equivalent. Two networks, Z_1 and Z_2, are said to be inverse with respect to an impedance Z_0 when they satisfy the relation that:

$$Z_1Z_2 = Z_0^2.$$

The specific impedance relationships which must be satisfied by two networks in order for them to be equivalent or inverse are summaried for convenient reference in Table 1.

The basic circuit setup for use of any type of four-terminal network is given in Fig. 16-6A. This schematic represents the insertion of the network at any point

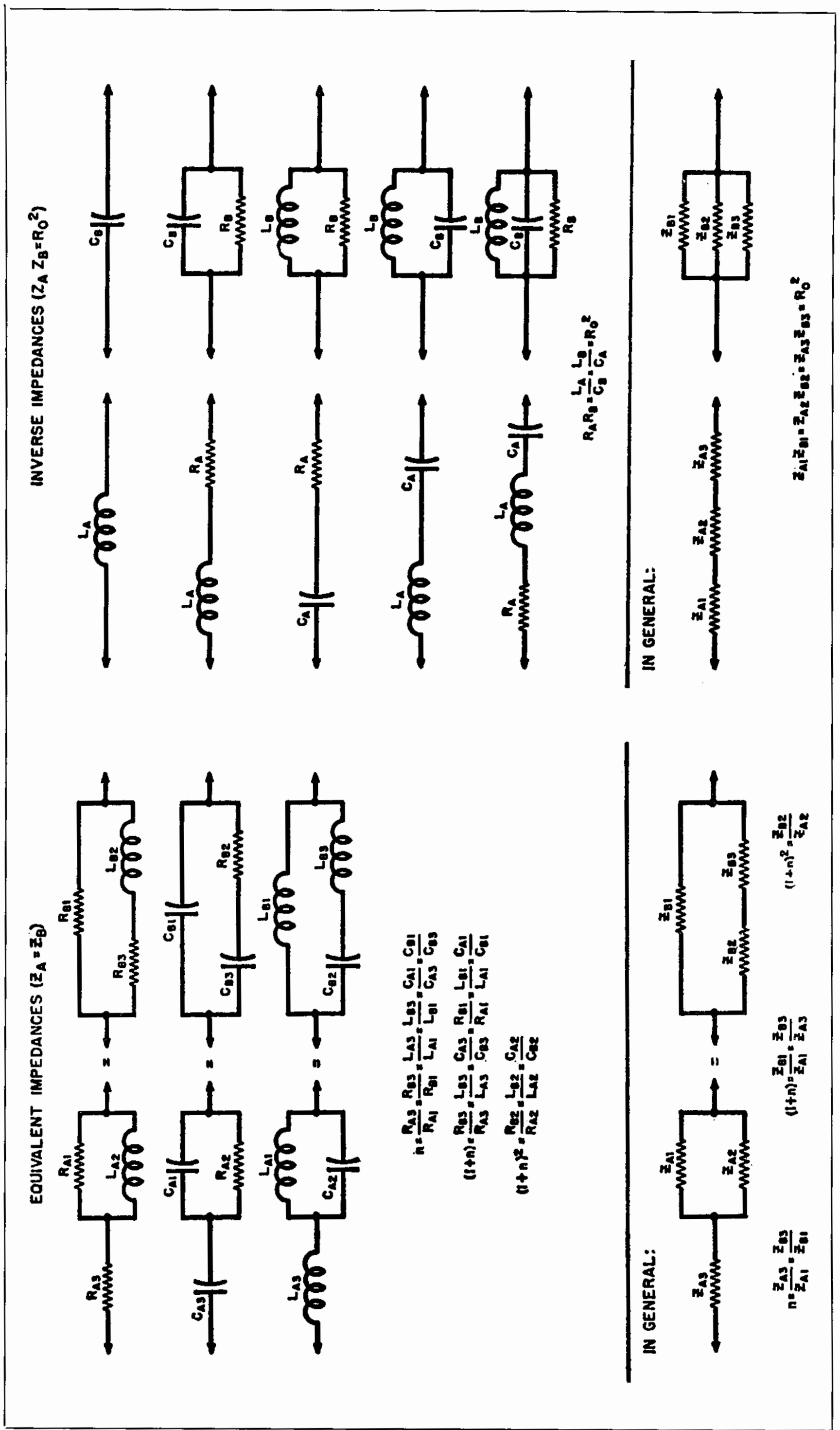


Table 1. Equivalent and inverse impedances, together with the general mathematical expressions for each.

in a communications channel, where the network represented by N operates between a transmitter of impedance Z_1 and a receiver of impedance Z_0. (Usually in audio network applications, $Z_1 = Z_0 = R_0$.) Then, if the circuit, taken alone without the network N, has the frequency response characteristic A in Fig. 16-6B, and the desired response curve is B, the response of the frequency-corrective network will be the difference between these two curves as represented by C.

In the use of four-terminal networks as illustrated in Fig. 16-6, it must always be kept clearly in mind that all networks consisting only of resistance, capacity and inductance with no sources of voltage (i.e., *passive* networks), can only *attenuate* frequencies, and cannot increase response at any frequency. Therefore, if it is desired to accentuate some particular frequency, this can be done only by attenuating all other frequencies; and the one frequency has therefore been accentuated at the expense of the over-all signal level.

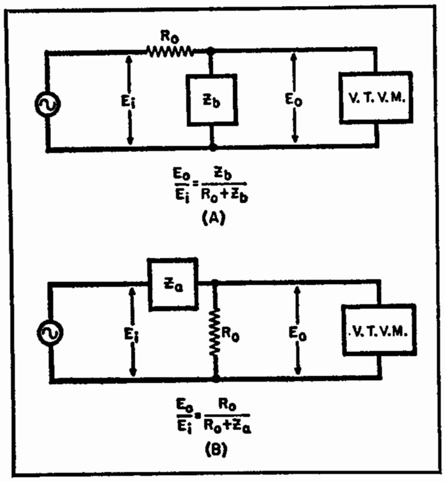


Fig. 16-7. Basic set-ups for experimental method of equalizer design.

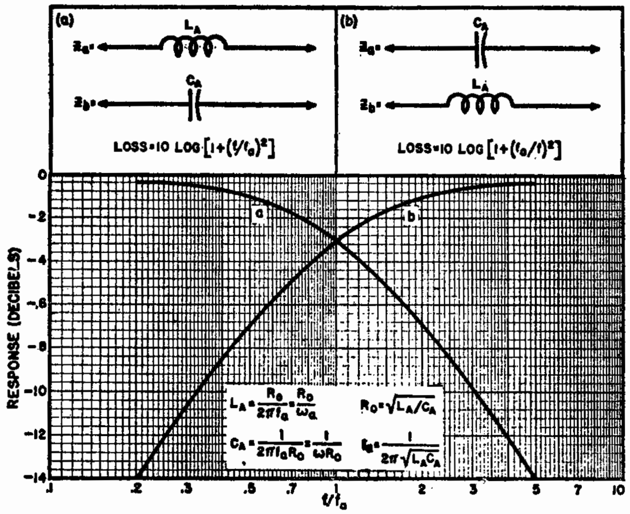
Attenuating Equalizers

An attenuation equalizer is a four-terminal network whose response varies more or less gradually in some desired manner over a given frequency range. Therefore, if a signal containing components of different frequencies is passed through the network, the relative amplitudes of the different components will have been altered in the desired manner when the signal is delivered to the load circuit. For example, in the response curve shown for the channel in Fig. 16-6B, the response drops off at the higher frequencies. Since in this case it is desired to correct the response to the cut-off frequency, the equalizer which is inserted in the circuit contains an attenuation characteristic inverse to that of the network to be corrected. The result obtained by adding the two response curves is the desired response.

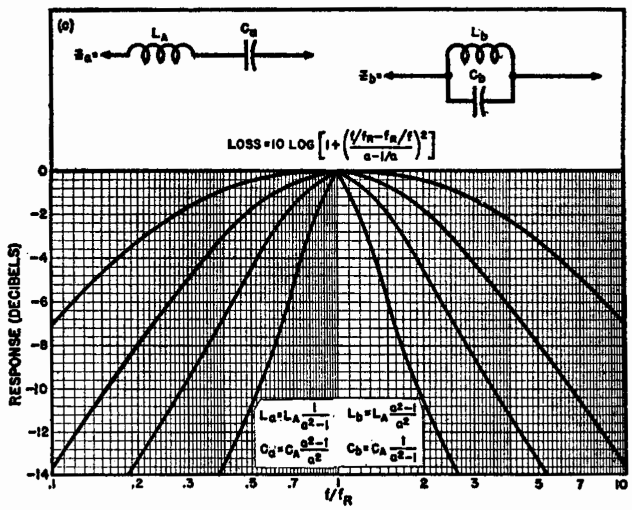
In the application which has just been described, the desired result was a certain response curve within the limits of the frequency band prescribed by the upper and lower cut-off frequencies. This is only one of the many different types of response curves which are of interest in audio-frequency work, and which may be attained by the use of attenuation equalizers. In some applications it is desired to have a response characteristic which rises toward the higher frequencies (for instance, in disc recording and FM broadcasting) to obtain a better signal-to-noise ratio. In many cases it is necessary to decrease or increase the response at some particular frequency (or narrow band of frequencies) in order to compensate for a peak or lack or response at that frequency. Almost all of the frequency response characteristics which are required in practical audio engineering can readily be attained by means of properly designed attenuation equalizers.

A great amount of work has been and is being done on the design of frequency-corrective networks and attenuation equalizers, and many different types of networks of varying complexity have been developed to produce the various results. The types of circuits which are of greatest importance and of greatest interest in audio-frequency work are described below:

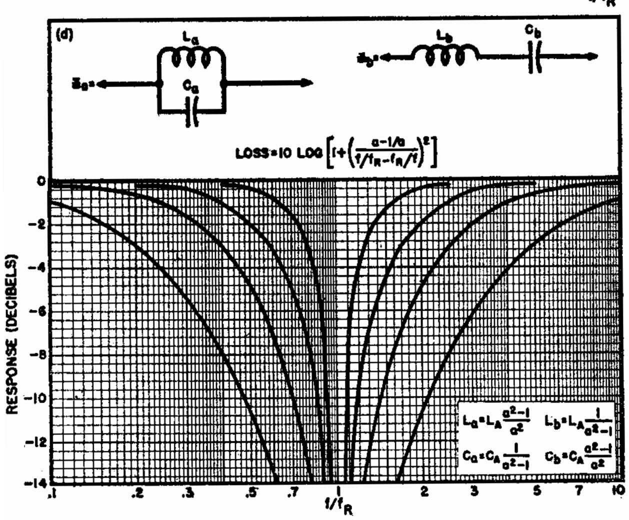
(a) *Simple RC frequency-corrective*


(a)
LOSS=10 LOG[1+(f/f_a)^2]
(b)
LOSS=10 LOG[1+(f_a/f)^2]
RESPONSE (DECIBELS)
L_A = R_o/(2πf_a) = R_o/ω_a
R_o = √(L_A/C_A)
C_A = 1/(2πf_aR_o) = 1/(ωR_o)
f_a = 1/(2π√(L_AC_A))
f/f_a

(A)


(c)
LOSS=10 LOG[1+((f/f_R−f_R/f)/(a−1/a))^2]
RESPONSE (DECIBELS)
f/f_R

(B)


(d)
LOSS=10 LOG[1+((a−1/a)/(f/f_R−f_R/f))^2]
RESPONSE (DECIBELS)
f/f_R

(C)

Notes for all types:

f_R = resonant freq. of Z_a and Z_b arms

f_a = freq. of 3 db insertion loss.

f_b = freq. where loss is ½ max. in db

f = any frequency

$b = f_R/f_b$, defined as > 1

R_o = equalizer resistance

Pad loss = max. loss = $20 \log_{10} K$

L = inductance in henrys

C = capacity in farads

Notes for types (e), (f), (g), (h):

$R_o = \sqrt{L_B/C_B}$, $R_a = R_o\,(K - 1)$

$R_b = R_o/(K - 1)$, $L_B = R_o/2\pi f_B = R_o/\omega_B$

$C_b = 1/2\pi f_B R_o = 1/\omega_B R_o$

$f_B = 1/2\pi\,\sqrt{L_B C_B}$

(D) Top

(E) Center

(F) Bottom

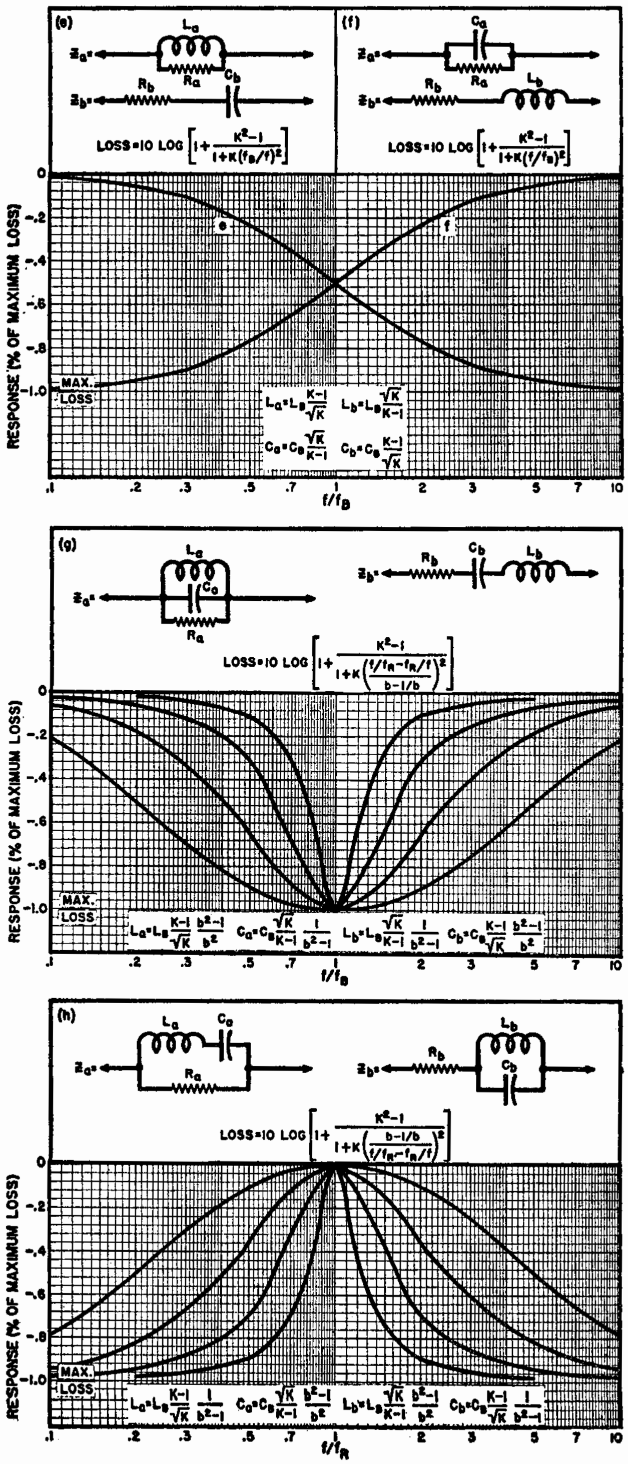


Table 2. Formulas and curves for equalizer design using several different networks for the Z_a and Z_b arms.

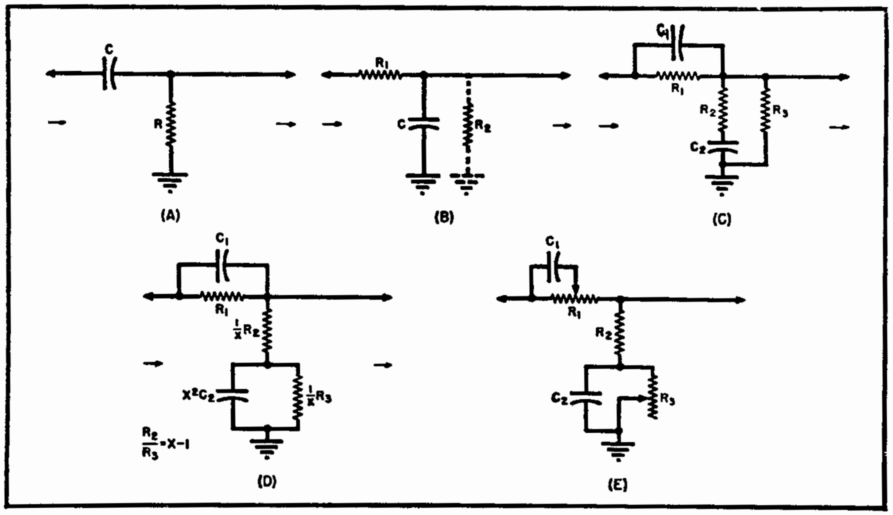
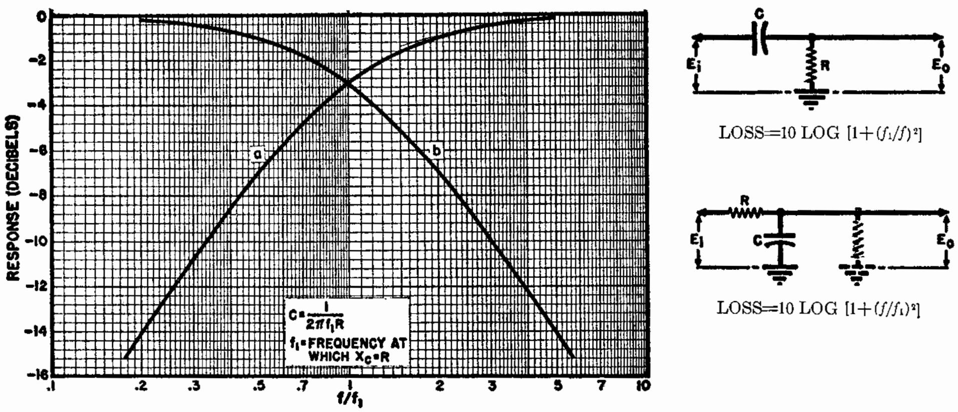


Fig. 16-8. Basic RC networks for frequency-response correction. (A) Low frequency attenuation. (B) High frequency attenuation. (C) and (D). Low and high frequency lift. (E) Variable low and high frequency lift.

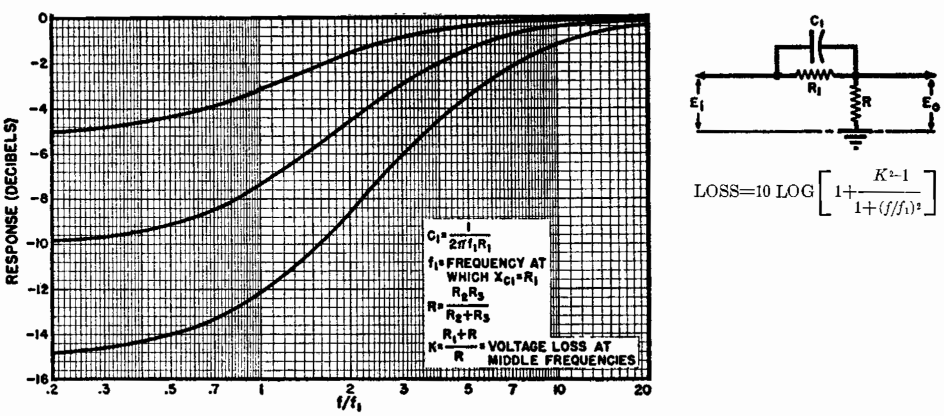
circuits. Many types of frequency correction can be obtained by simple circuits containing only resistance and capacity elements. The basic circuits for frequency-response correction by means of *RC* networks are shown in Fig. 16-8. All the basic circuits are simple ladder networks. For low-frequency attenuation the circuits consist of a condenser in the series arm with a resistance as the shunt arm; while for high-frequency attenuation, the circuit is a resistance in the series arm, with a condenser (in parallel with a resistance) as the shunt arm. For high and low-frequency lift, the network consists of a resistance and capacity in parallel as the series arm, and two resistors and a capacity to form the shunt arm. The input and output impedances vary with frequency, and these networks are most practical for use in high-impedance circuits such as interstage coupling and feedback in resistance-capacitance coupled amplifiers.

The equations for the frequency response characteristics of these circuits, together with a chart and a set of curves by which practical *RC* corrective networks may be designed for specified characteristics, are given in Table 3.

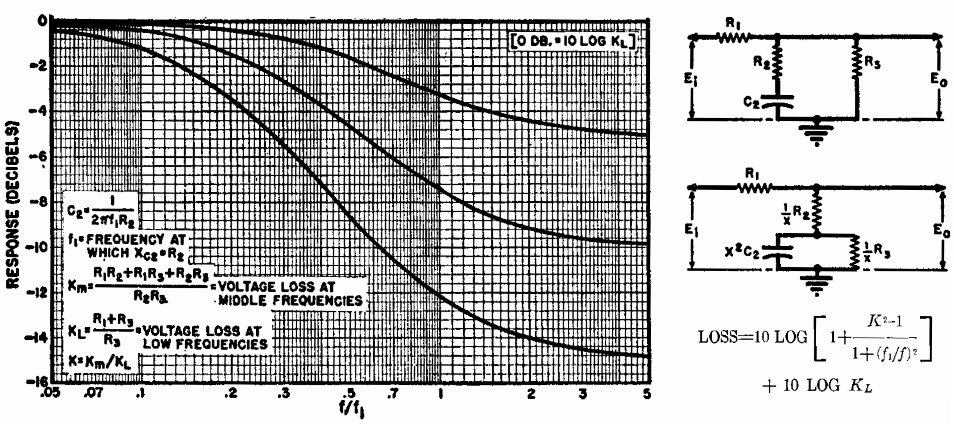
Low-frequency and high-frequency attenuations are obtained from the circuits in Fig. 16-8A and B in the following manner: In network A, the output is determined by the voltage divider consisting of the series capacity and the shunt resistance. At high frequencies, the impedance of the condenser is small compared to the resistance, and essentially the entire input voltage is delivered to the load; below a certain crossover frequency determined by the relative values of the resistance and capacity, the impedance of the condenser becomes large compared to the resistance, and as the frequency becomes lower the voltage delivered to the output becomes progressively less as the impedance of the condenser increases. In network B the output is determined by the voltage divider consisting of the series resistance and the parallel resistance and capacity in shunt. At low frequencies the condenser has essentially infinite impedance, and the output is determined by the resistances; above some crossover frequency, the impedance of the condenser begins to short-circuit the shunt resistance, and as the frequency increases the output voltage becomes



A



B



C

Table 3. The design of RC equalizers.

progressively less as the impedance of the condenser decreases.

Low-frequency and high-frequency lift are obtained from the circuits in Fig. 16-8C and D in the following manner: At the middle frequencies the circuit is essentially a resistive voltage divider consisting of R_1 in series with R (where R is the combined resistance in the shunt arm when the condenser is short-circuited), since at these frequencies condenser C_1 is effectively an infinite impedance and C_2 a short circuit. Thus, the insertion loss of the network is determined by the output of this voltage divider, and this also determines the maximum amount of frequency correction since the maximum voltage output cannot be more than the input voltage.

At higher frequencies the reactance of condenser C_1 becomes smaller until at sufficiently high frequencies resistor R_1 is effectively short-circuited and the entire input voltage appears across the output terminals. At low frequencies, the series arm R_1 remains constant, but the reactance of C_2 in the shunt arm of the network increases as the frequency decreases. Therefore at low frequencies the shunt arm of the voltage divider becomes relatively a higher impedance compared to R_1, and a greater proportion of the input voltage appears across the output terminals. The maximum amount of low-frequency lift is determined by the relative values of resistors R_1 and R_3.

In this circuit, the specific amounts of high- and low-frequency equalization are determined by the relative values of resistors R_1, R_2 and R_3. The frequencies at which the lift occurs are determined by the value of C_1 relative to R_1, and C_2 relative to R_2. The design curves in Table 3 are therefore plotted in terms of frequency ratio (i.e., relative to f_1 at which X_{c1} is equal to R_1, and to f_2 at which X_{c2} is equal to R_2), and in decibels of equalization. The formulae by which the values of resistance and capacity are determined from the curves are given in the table. When they are drawn in this manner the curves and the table are thus universal in application to the design of equalizers of this type.

In designing an equalizer in a practical problem, the desired curve is compared with the curves in Table 3, and the most suitable one is selected or interpolated from the given curves. This comparison gives the values of high- and low-frequency equalization, the middle-frequency insertion loss, and the frequencies f_1 and f_2. From these, values of resistance and capacity are found by means of the formulae.

One additional piece of information is required before the high- or low-frequency lift equalizer can be designed; one of the resistors R_2 or R_3 must be selected as the reference impedance for the network. These values are generally dictated by the requirements of the circuit in which the network is to be used. The network has its minimum input impedance (equal to R) at high frequencies and its highest output impedance (equal to R_3 at low frequencies. In most cases at least one of these values is determined by the circuit in which the network is used, and this furnishes complete data for the design of the *RC* equalizer.

The basic circuit for high- and low-frequency lift has been given in two different forms in Fig. 16-8 and in Table 3. Usually the circuit in the form C is more convenient to use when the amount of low-frequency equalization required is constant, as in interstage coupling in an amplifier to correct for deficiencies elsewhere in the system; while form D of the circuit is more convenient to use when a variable amount of equalization is required, as in the tone control of a radio receiver or a phonograph amplifier. An example of how this network may be used as a variable equalizer is given in Fig. 16-8E. This variable equalizer has the feature that the high- and low-frequency equalization may be varied independently and the middle-frequency level remains constant regardless of the amount of equalization. The basic circuit which has been described may also be used for either high or low-frequency correction alone. For low-frequency lift, condenser C_1 is omitted. For high-frequency lift, condenser C_2 is removed and the two resistors R_2 and R_3 are replaced by a single resistor R having the appropriate value.

(b) *Constant-impedance and conventional L,C,R equalizers.* Much greater variety of frequency response and flexibility of design are offered by the constant-impedance and other types of conventional equalizers containing inductive as well as resistance and capacity elements. In the equalizer circuits which fall into this category, the transmission characteristics in general are made to depend upon a series impedance Z_a and a shunt impedance Z_b. These two impedances must satisfy the condition that they are inverse to each other with respect to the line impedance Z_o; that is $Z_a Z_b = Z_o$.

Equalizers of this type fall into seven different circuit arrangements which have been found most satisfactory for general use. These various circuit arrangements are summarized briefly in Table 4. Because Z_o is almost always resistive, it is assumed that these equalizers are to be used in a line whose impedance is resistive and has a value equal to R_o (so that $Z_a Z_b = R_o^2$). The first two circuit arrangements, the series impedance and the shunt impedance, do not present constant impedance to either the input circuit or to the load. The full series and the full shunt present constant impedance R_o to the input circuit; while the bridged-T, the T and the lattice arrangements are symmetrical and present constant impedance R_o both the input circuit and the load. (This is indicated on the chart by the arrow and the symbol R_o to indicate when the circuit presents a constant impedance R_o). For the same impedances Z_a and Z_b, when driven from a source of impedance R_o and terminated in a resistive load R_o, the frequency response characteristics of all seven types of equalizer circuits shown in this chart will be identical.

The frequency characteristic of the equalizer may be determined from the fundamental equation for the insertion loss:

$$I.L. = 20 \log \frac{R_o + Z_a}{R_o} = 20 \log \frac{R_o + Z_b}{Z_b} \quad (1)$$

Thus, the manner in which the response varies over the frequency range depends entirely upon the manner in which Z_a and Z_b vary with respect to R_o, and can be calculated from either the series or the shunt impedance arm. From this insertion loss equation, it can be seen that the frequency-attenuation curve of this class of equalizer depends only upon the impedance configuration of the series and shunt arms, and since they are inverse to one another, only one of them need be known and the entire response of the equalizer is specified.

The number of different possible two-terminal impedances which might be used as the series and shunt arms in an equalizer is almost unlimited, but in actual practice it has been found that most of the desired results can be accomplished by one or more of a few simple impedances. The most useful of these are:

(a) Z_a and inductance and Z_b a capacity.

(b) Z_a a capacity and Z_b an inductance.

(c) Z_a a series-resonant circuit and Z_b parallel-resonant

(d) Z_a parallel-resonant and Z_b series-resonant

(e) Z_a an inductance in parallel with a resistance and Z_b a capacity in series with a resistance

(f) Z_a a capacity in parallel with a resistance and Z_b an inductance in series with a resistance

(g) Z_a a parallel-resonant circuit in parallel with a resistance and Z_b a series-resonant circuit in series with a resistance

(h) Z_a a series-resonant circuit in parallel with a resistance and Z_b a parallel-resonant cricuit in series with a resistance.

The different frequency responses obtained when these various impedances are used are shown in Table 2. The equations for the curves, giving the exact insertion loss in decibels for any frequency (also included in this chart) are obtained by substituting the impedance of Z_a or Z_b into the general insertion loss equation given above.

As a practical aid in the engineering design of attenuation equalizers, the response of the various types of equalizers are accurately plotted in the six sets of curves in Table 2. The curves have been plotted on a universal scale so that they can be used for the general design of filters having various frequency-response characteristics to meet the requirements of particular engineering problems. The frequency scale in these curves is in terms of a ratio with respect to a reference frequency determined by the type of equalizer. (The manner in which these reference frequencies are chosen is indicated in the chart). In the first three sets of curves, representing the response of equalizer types (a), (b), (c) and (d) the vertical scale is an absolute scale in terms of insertion loss in decibels. In the remaining three sets of curves, representing the response of equalizer types (e), (f), (g) and (h) the vertical scale is a relative one in terms of percentage of maximum insertion loss (which will vary for different equalizer designs). These last three sets of curves are not strictly accurate in this form, but they give a very close approximation.

The formulae by means of which the electrical elements of an equalizer are determined from the various parameters of these curves are included in the chart in Table 2. The procedure in using these curves to design an equalizer to have a desired frequency response is to find the curve in the chart which best matches the required response, taking proper account of the relative frequency and insertion loss scales, then to calculate the values of the elements by means of the formulae in the table.

It may be noted that the total amount of equalization is determined by the values of the two resistors R_a and R_b. This affords a convenient method of designing equalizers for variable amounts of equalization, by varying the resistance values (either by means of a multipole switch, or by means of commercially available variable attenuators) while maintaining the relationship $R_aR_b = R_o^2$.

The fundamental equalizer insertion loss equation (1) indicates a convenient experimental method of equalizer design, in addition to the method of design by means of the curves given in Table 2. The basic principles of this method may be understood from the schematic block diagrams in Fig. 16-7. It can be seen in the circuits as set up in Fig. 16-7 that the output of the voltage dividers, consisting of R_o and of the impedance arms of the equalizer, satisfies the general equation for the insertion loss of the equalizer types under discussion. Thus, the equalizer may be designed completely by experimental measurement, by setting up either of the voltage divider networks shown in Fig. 16-7 and adjusting the impedance until the desired output frequency response curve is obtained. Once the one impedance arm has been determined, the other may be found by simple calculation from the formula for inverse impedances, i.e., $Z_a\ Z_b = R_o^2$. This method is completely general, and applies to all types of impedance configurations for Z_a and Z_b, including those covered in Table 3.

(c) *Constant-B equalizers.* It may be seen from Table 2 that in equalizers of types (g) and (h) the curve has two parameters which may be varied to change both the shape and the maximum insertion loss. The maximum loss may

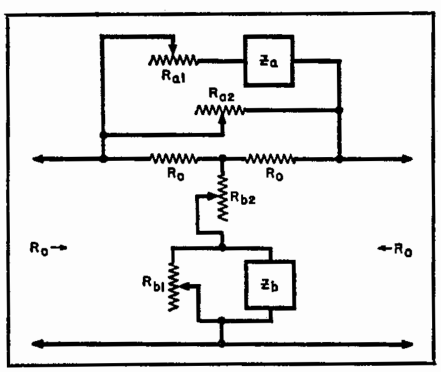


Fig. 16-9. Circuit diagram of bridged-T constant B variable equalizer.

easily be changed by changing the values of the resistors R_a and R_b in accordance with the formulae given in Table 2. However, when this is done the shape of the curve (represented by the parameter b) changes at the same time. In other words, the two parameters are not independent in these types of equalizers. In a great many variable equalizer applications this is extremely inconvenient. However, by choosing the impedances Z_a and Z_b in the proper manner, the shape of the curve may be made to remain constant as the amount of equalization is changed.

The circuit of this type of equalizer is shown in the schematic diagram in Fig. 16-9. The impedance configuration is seen to be similar to that of types (g) and (h), except for the addition of the two resistors R_{a1} and R_{b1}, whose function is to maintain the shape of the curve constant for different amounts of equalization. As in the conventional types of equalizers, the series and shunt arms are inverse to each other. This condition is represented by the equations: $R_{a1}\,R_{b1} = R_o^2$, $R_{a2}\,R_{b2} = R_{o2}$, $(jX_1)\ (jX_2) = R_o^2$. The various design equations covering the operation of the constant-B type of equalizer may be summarized in the following manner:

Since the series and shunt reactances are inverse, when either is zero the other is infinite. At frequencies for which $jX_1 = 0$ and $jX_2 = \infty$, the resistances R_{a1} and R_{a2} are in parallel, R_{b1} and R_{b2} are in series; at these frequencies the circuit has minimum insertion loss. At frequencies for which $jX_1 = \infty$ and $jX_2 = 0$, neither R_{a1} nor R_{b1} has any effect in the circuit; at these frequencies the circuit has maximum insertion loss. Inserting these conditions in Eqt. (1), the basic equation for the insertion loss of a constant-impedance equalizer, gives the quantitative result that:

$$\text{Max. loss} = 20 \log \left[1 + \frac{R_{a2}}{R_o}\right] \quad \ldots\ldots (2)$$

$$\text{Min. loss} = 20 \log \left[1 + \frac{R_p}{R_o}\right] \quad \ldots\ldots (3)$$

where R_p is the parallel resistance of R_{a1} and R_{a2}. The difference between maximum and minimum loss is the amount of equalization of the circuit.

Thus:

Equalization = max. loss — min. loss

$$= 20 \log \left[\frac{1 + R_{a2}/R_o}{1 + R_p/R_o}\right] \quad \ldots\ldots (4)$$

In the design of a variable equalizer, the amount of equalization is, in general, varied in known steps from zero to some maximum amount at the top step of the control dial, therefore the equalization and the manner in which it is to be varied may be assumed to be known design information. For a definite amount of equalization, the shape of the curve (represented by the parameter b) may be varied by adjusting the maximum and minimum loss while keeping their difference constant. In constant-B equalizers, the object is to keep f_b constant for all steps of the equalizer. The condition for this to be true is that:

$$\sinh^2 \left[\frac{\begin{matrix}\text{Max. loss on}\\ \text{any step}\end{matrix}}{2 \times 8.68}\right] = \sinh \left[\frac{\begin{matrix}\text{Equalization}\\ \text{on same step}\end{matrix}}{2 \times 8.68}\right] \times \sinh \left[\frac{\begin{matrix}\text{Max. equalization}\\ \text{on top step}\end{matrix}}{2 \times 8.68}\right] \quad \ldots\ldots (5)$$

Since the two factors on the right are known, the factor on the left may then be determined. In most cases Eqt. (5) can be simplified to avoid the use of hyperbolic functions. For equalizers having maximum losses not greater than 15 or 20 db, the hyperbolic angles in the equation are small, so that the sinh of the angle is approximately equal to the angle, Thus, as an approximation to Eqt. (5):

$$\begin{pmatrix}\text{Max. loss on}\\ \text{any step}\end{pmatrix}^2 = \begin{pmatrix}\text{Equalization on}\\ \text{same step}\end{pmatrix} \times \begin{pmatrix}\text{Max. equalization}\\ \text{on top step}\end{pmatrix} \quad \ldots\ldots (6)$$

Eqts. (2) to (6) give all the information necessary for the design of the resistance attenuator portion (i.e., R_{a1}, R_{a2}, R_{b1}, and R_{b2}) of the constant-B equalizer. The procedure and the steps which should be followed in the design of such an equalizer from these equations may be summarized briefly in the following manner:

(1) Note that at the top step the equalization is maximum, and is seen from Eqt. (6) also to be equal to the maximum loss. Thus, on the top step the minimum loss is zero, $R_{a1} = 0$, and this is therefore a conventional equalizer on the top step. Therefore, determine first the reactance values and the top step resistances by designing the conventional equalizer for the top step.

(2) On all other steps, find the maximum loss by Eqt. (6) from the desired equalization and the equalization on the top step.

(3) From the maximum loss on any step, find R_{a2} from Eqt. (2)

(4) From the equalization and the maximum loss, find R_p by using Eqt. (3).

(5) Find R_{a1} from R_p and R_{a2}.

(6) Determine R_{b1} and R_{b2} from R_{a1} and R_{a2} by using the inverse relationships, i.e. $R_{a1}\ R_{b1} = R_o^2$ and $R_{a2}\ R_{b2} = R_o^2$.

From this information the equalizer is then completely specified.

Filter Networks

A wave filter is a four-terminal network which has negligible attenuation for a certain band of frequencies, and high attenuation for other frequencies. Since there are many applications where it is desirable or necessary to permit only certain bands of frequencies or to eliminate certain frequencies, wave filters are of considerable importance in electronic design.

The various frequency-selective attenuation characteristics of wave filters are classified into four different categories:

(1) low-pass filters, which pass all frequencies up to some finite cutoff frequency and attenuate all higher frequencies,

(2) high-pass filters, which transmit all frequencies above the cutoff frequency and attenuate all lower frequencies,

(3) band-pass filters, which transmit

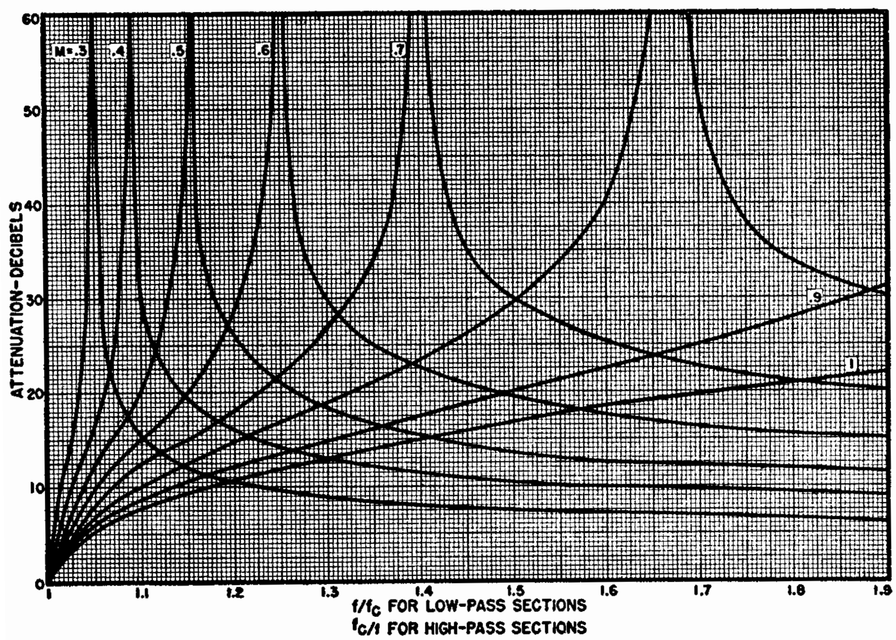


Fig. 16-10. Attenuation per section for low and high pass filters for various values of m.

a definite band of frequencies and attenuate all frequencies outside this band,

(4) band-elimination filters, which attenuate a definite band of frequencies and transmit all frequencies outside this band.

Since band-elimination filters are seldom used in practice, this section will be restricted to the use of the first three types of wave filters.

A wave filter is composed of series and shunt two-terminal impedances so chosen as to give the desired pass band and attenuation characteristics. Unlike attenuation equalizers, the two-terminal impedances in filters must be purely reactive in order that there be no appreciable attenuation within the pass band.

Most of the practical problems which require the use of wave filters can be solved by the use of ladder type filters composed of symmetrical T or π sections connected in tandem in sufficient numbers and types to secure the desired attenuation characteristic. The basic circuit configurations, impedance relations, and the relationships between the different types of ladder filter sections are given in Fig. 16-12. Filter sections are expressed in terms of the two inverse impedances Z_1 and Z_2, which may be any pair of inverse two-terminal reactive networks. The equations given in Fig. 16-12 show that the attenuation characteristics and image impedance of the two various sections depend upon the impedance ratio $Z_1/4Z_2$. The filter has negligible insertion loss for all frequencies that make this ratio lie between 0 and -1, while all frequencies outside this range are attenuated. For example, when Z_1 is an inductance and Z_2 a capacity, a low-pass filter is obtained. When Z_1 is a capacity and Z_2 an inductance, a high-pass filter results. When Z_1 consists of a series-resonant circuit of inductance and capacity in series, and Z_2 the corresponding inverse circuit band-pass filter sections are obtained. Thus, by assigning the proper circuit configurations and values to Z_1 and Z_2 the desired types of filter sections and pass-band characteristics can readily be obtained.

The simplest type of filter section is the constant-k filter, in which only Z_1 appears in the series arm, and Z_2 in the shunt arm. In the m-derived sections, Z_1 and Z_2 appear together in either the series or the shunt arm. The m-derived sections in general give a sharper cutoff and a more constant image-impedance in the pass band, while the constant-k section has a greater attenuation at frequencies far beyond cutoff. From the equations in Fig. 16-12 it can be seen that the constant-k section may be considered as an m-derived section in which $m = 1$. An important point which must be remembered in connection with the design of filters is that when sections are combined into a multi-section filter, they must always be combined on a matched-impedance basis as indicated in Fig. 16-12.

The various types of filter sections, together with the information and formulae necessary for the design of band-pass, low-pass, and high-pass filters are summarized in convenient form for practical filter design in Tables 1, 2, and 3. These tables give the circuit configurations of the various filter sections, together with their attenuation characteristics, image impedance, and formulae for calculation of the values of the various circuit elements from the design data

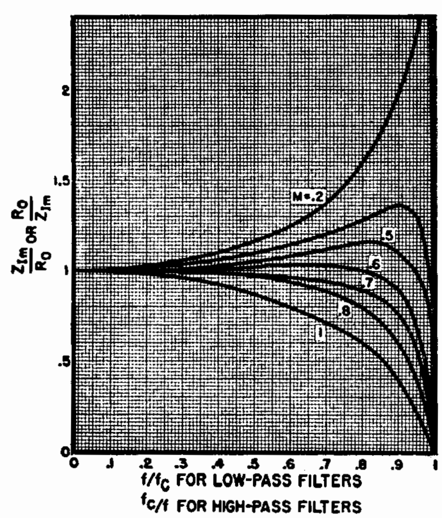


Fig. 16-11. Effect of parameter m on image impedance characteristics.

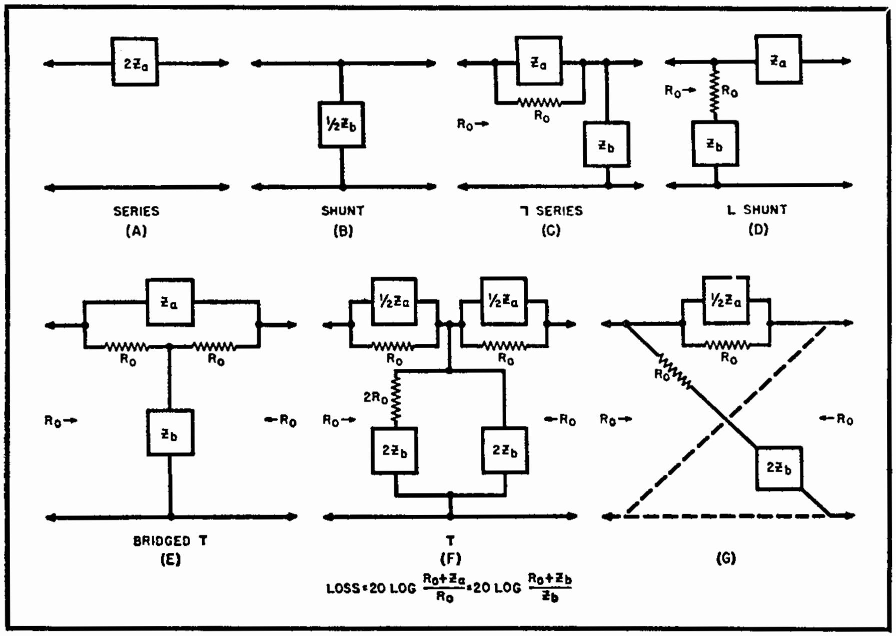
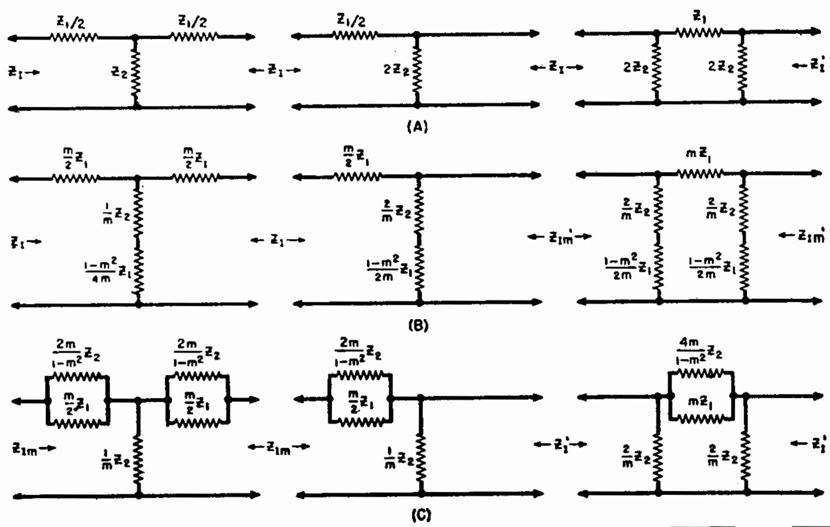


Table 4. Types of equalizer networks. In types (A) and (B), both input and output impedance vary. Types (C) and (D) have a constant input impedance. Types (E), (F) and (G) are symmetrical and have constant input and output impedances.

and the desired attenuation characteristic which is to be achieved by the completed filter.

An accurate set of curves for the attenuation loss of high- and low-pass sections is given in Fig. 16-10. This chart gives curves for the attenuation of filter sections for different values of the parameter m. (For half-sections, the attenuation values should be divided by two.) It must be noted in the use of these curves that they are the ideal curves for dissipationless circuit elements, and that the actual curves will be somewhat different due to dissipation in the network. When the desired attenuation curves of a filter designed to solve some specific practical problem cannot be attained by the use of a single section filter, several sections, connected on a matched image impedance basis, may be used to give the desired characteristic. When such sections are combined in this manner, the resulting insertion loss is obtained by adding together the losses of each section as given in Fig. 16-10. By suitable choice of the parameter m, and hence the frequency of infinite attenuation and the sharpness of cutoff, a wide variety of attenuation characteristics can be obtained.

Since the equations for the image impedance of the various filter sections contain terms which vary with frequency, there can be no exact impedance match between a wave filter and resistive terminations at the generator and at the load. However, the mismatch can be minimized by selection of the proper value of the parameter m. The curves in Fig. 16-11 show the image impedance characteristics of filter sections within the transmission band for various values of m. They show that a value of $m = 0.6$ provides the closest impedance match to a constant resistance R_o over the greatest part of the frequency band. For this reason, practical filters are generally designed with terminal half-sections having $m = 0.6$ to match resistive input and output impedances, and with the intermediate sections chosen to give whatever special attenuation characteristic may be desired.



Equations:

$$Z_1 Z_2 = R_0^2$$

$$Z_I = R_0\sqrt{1+\frac{Z_1}{4Z_2}}\,;\quad Z_I' = \frac{R_0}{\sqrt{1+\frac{Z_1}{4Z_2}}}$$

$$Z_{Im}' = Z_I'\left[1+(1-m^2)\frac{Z_1}{4Z_2}\right]$$

$$Z_{Im} = \frac{Z_I}{1+(1-m^2)\frac{Z_1}{4Z_2}}$$

$$\text{Attenuation for constant-}k\text{ sections} = 20\log\frac{\sqrt{1+\frac{Z_1}{4Z_2}}+\sqrt{\frac{Z_1}{4Z_2}}}{\sqrt{1+\frac{Z_1}{4Z_2}}-\sqrt{\frac{Z_1}{4Z_2}}}$$

$$\text{Attenuation for } m\text{-derived sections} = 20\log\frac{\sqrt{1+\frac{Z_1}{4Z_2}}+m\sqrt{\frac{Z_1}{4Z_2}}}{\sqrt{1+\frac{Z_1}{4Z_2}}-m\sqrt{\frac{Z_1}{4Z_2}}}$$

Fig. 16-12. Circuit configurations and equations for basic ladder filter sections. (A) Constant-k, (B) series m-derived, and (C) shunt m-derived.

The procedure is usually the same in the design of any type of wave filter and is generally done in a number of steps:

1. Determine the cutoff frequencies which mark the edge of the pass-band to give the required characteristics, and the load resistance of the circuit into which the filter is to be inserted.

2. Decide whether T or π intermediate sections are to be used. This decision is based primarily upon considerations of convenience or economy, since the electrical performance of the two types is identical. (Thus, the use of shunt-derived high-pass sections will result in a saving in the required number of inductances, which are generally more expensive than condensers.)

3. Design the terminating half-sections according to the tables, for the chosen cut-off frequencies and terminating impedance.

4. Decide on the number and type of intermeidate sections to be used. The more sections used, the greater is the attenuation in the stop band. More than two intermediate sections are not generally required.

5. Select the frequencies at which the different intermediate sections are to have their maximum attenuation and design the sections according to the formulae given in the appropriate tables.

By following this procedure and using the various tables and charts given in this section, practical wave filters may readily be designed to meet the different filter problems which arise in general electronic practice.

It is sometimes necessary in electrical

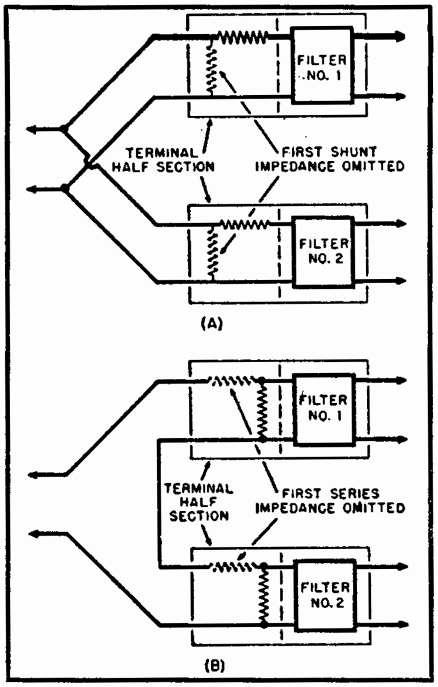


Fig. 16-13. Method of connecting complementary filters for operation in parallel and series.

systems to operate filter sections in parallel at their input or output terminals. In such systems, the image impedance of each filter is across the common terminals and, since the filter impedance may be approximately constant resistance in their pass-band but are reactive and vary with frequency in the stop-band, the parallel impedance and insertion loss are unfavorably affected. Proper design methods must be used to eliminate this factor. Arrangements of this type are used widely in audio work in frequency-dividing networks, such as for loudspeaker systems, and the manner in which proper impedance matching is obtained will be described in this section.

In using dividing networks for most types of loudspeaker systems, the frequency band is divided into two parts by means of a low-pass and a high-pass filter operated in series or in parallel. When filters are connected in this manner, it is found that it is necessary to employ special design methods only in the first terminal half-section of each of the filters. The changes in design which have been found necessary in filters which are to be used in this type of service are as follows:

(a) When parallel operation is required, the low- and high-pass filters must have T intermediate sections, and therefore L input half-sections. If the two input half-sections are designed for $m = 0.6$ and the two filters have equal load resistances, then normal impedance matching is obtained throughout, by omitting the shunting impedances at the inputs of both filters.

(b) When series operation is required, the filters must have π intermediate sections, and therefore π input half-sections. Then if the input half-sections are designed for $m = 0.6$ and the two filters have equal load resistances, normal impedance matching is obtained by omitting the series impedances at the inputs of both filters.

These relations are indicated in Fig. 16-13. They hold true not only for loud-

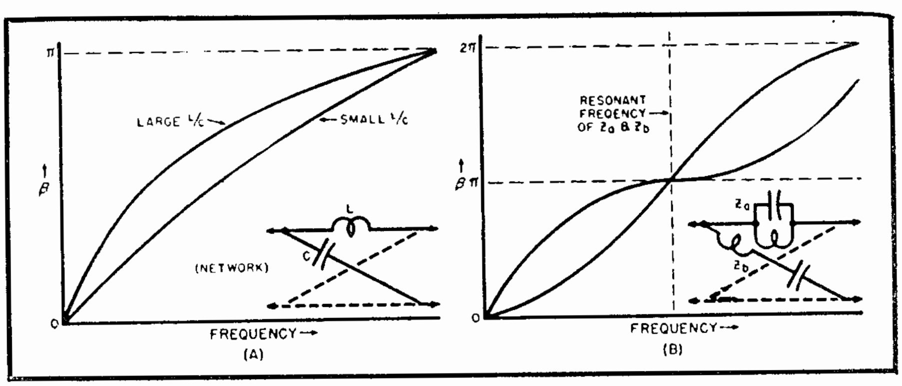
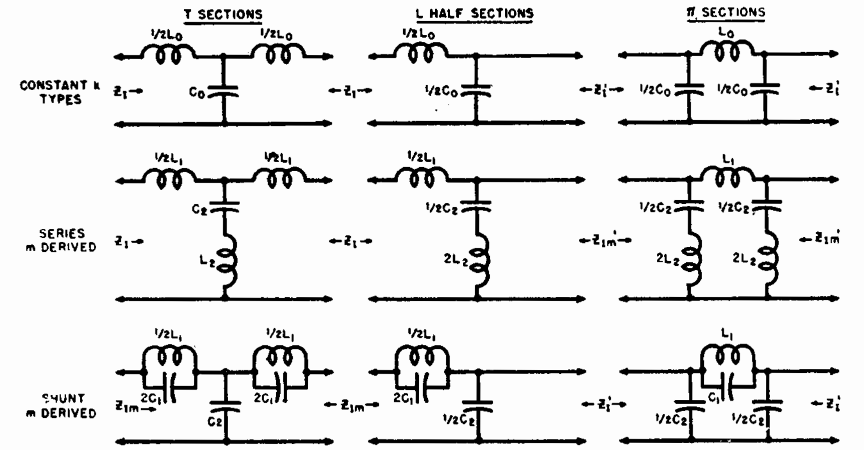


Fig. 16-14. Phase-shift characteristics of two simple types of all-pass sections.



R_0=load resistance $\quad f_R$=a frequency of high attenuation
f_c=cut-off frequency $\quad$ =resonant freq. of tuned circuits

$$L_0 = \frac{R_0}{\pi f_c},\ L_1 = mL_0,\ L_2 = \frac{1-m^2}{4m} L_0$$

$$C_0 = \frac{1}{\pi f_c R_0},\ C_1 = \frac{1-m^2}{4m} C_0,\ C_2 = mC_0$$

$$f_c = \frac{1}{\pi\sqrt{L_0C_0}},\ f_R = \frac{1}{2\pi\sqrt{L_1C_1}} = \frac{1}{2\pi\sqrt{L_2C_2}}$$

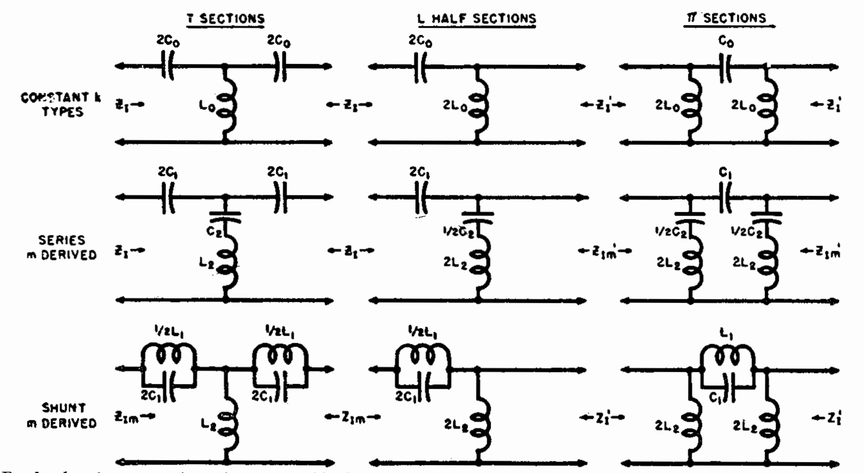
$$\frac{f_R}{f_c} = \frac{1}{\sqrt{1-m^2}},\ m = \sqrt{1-\left(\frac{f_c}{f_R}\right)^2}$$

$$Z_I = R_0\sqrt{1-(f/f_c)^2},\ Z_I' = \frac{R_0}{\sqrt{1-(f/f_c)^2}}$$

$$Z_{I\,m} = \frac{Z_I}{1-(1-m^2)\,(f/f_c)^2}$$

$$Z'_{I\,m} = Z'_1\,[1-(1-m^2)\,(f/f_c)^2]$$

Table 5. Information for the design of low-pass filter sections.



R_0=load resistance $\quad f_R$=a frequency of high attenuation
f_c=cut-off frequency $\quad$ =resonant freq. of tuned circuits

$$L_0 = \frac{R_0}{4\pi f_c},\ L_1 = \frac{4m}{1-m^2} L_0,\ L_2 = \frac{L_0}{m}$$

$$C_0 = \frac{1}{4\pi f_c R_0},\ C_1 = \frac{C_0}{m},\ C_2 = \frac{4m}{1-m^2} C_0$$

$$f_c = \frac{1}{4\pi\sqrt{L_0C_0}},\ f_R = \frac{1}{2\pi\sqrt{L_1C_1}} = \frac{1}{2\pi\sqrt{L_2C_2}}$$

$$\frac{f_c}{f_R} = \frac{1}{\sqrt{1-m^2}},\ m = \sqrt{1-(f_R/f_c)^2}$$

$$Z_I = R_0\sqrt{1-(f_c/f)^2},\ Z_I' = \frac{R_0}{\sqrt{1-(f_c/f)^2}}$$

$$Z_{I\,m} = \frac{Z_I}{1-(1-m^2)\,(f_c/f)^2}$$

$$Z'_{I\,m} = Z'_I\,[1-(1-m^2)\,(f_c/f)^2]$$

Table 6. Information for the design of high-pass filter sections.

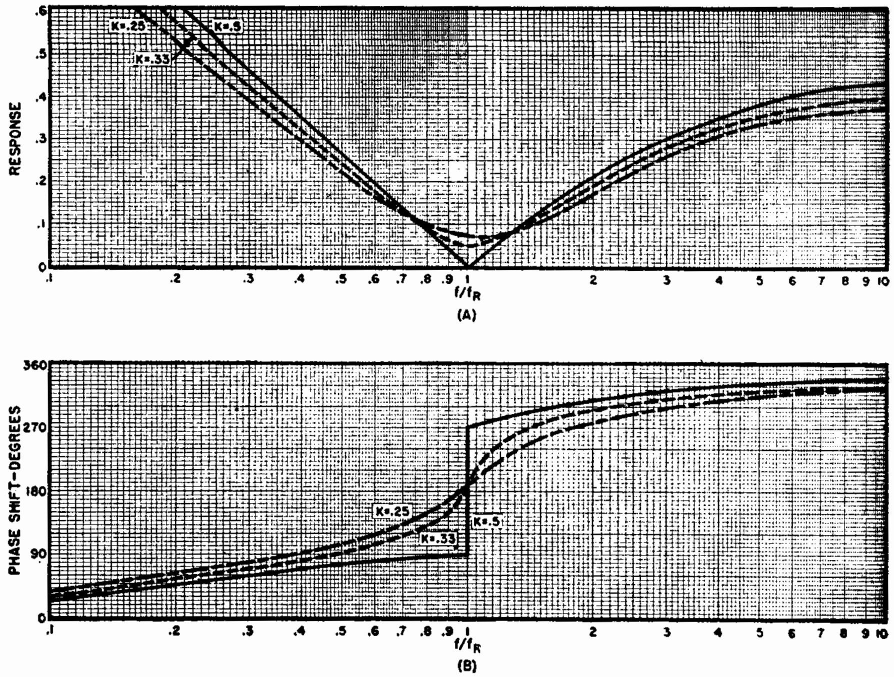


Fig. 16-15. (A) Amplitude response and (B) phase characteristics of parallel-T circuit shown in Fig. 16-20, for various values of the parameter K.

speaker dividing networks, but for all cases of two complementary filters, i.e., where those frequencies that lie in the stop-band of one filter are in the pass-band of the other.

Dividing networks for loudspeaker systems are not of the sharp cut-off type, and are generally designed to have from 12 to 18 db attenuation one octave away from the cross-over frequency. The circuits of the most common types of networks for this service are given in Table 9, which includes also the necessary design formulae. The upper two networks consist of input half-sections designed as indicated above, followed by constant-k full-sections, and give an attenuation of approximately 20 db for the first octave beyond cross-over. The lower two networks consist of the same input half-sections followed by constant-k half-sections, and give an attenuation of about 15 db for the first octave away from cross-over. The attenuation curves for these four networks are given in Fig. 16-16.

Since loudspeaker frequency-dividing networks carry the full output power of the amplifier, they must be designed for low transmission loss. When high-Q coils having low resistance are used, the loss may be kept down to the order of 0.5 db in systems of the type described above.

A type of circuit which has been found extremely useful in audio-frequency network design is the resistance-capacity single-frequency rejection circuit. Its usefulness is derived mainly because of the difficulties which attend the use of the standard L-C resonant circuit for many audio-frequency applications. Because audio frequencies are low, the inductances tend to become quite large physically, and high Q is generally difficult to attain, especially at the lower audio frequencies.

The R-C circuit can be made to exhibit the characteristics of a series-resonant L-C circuit, or to have a sharp cusp-shaped null at the desired frequency. By proper use of this circuit many desirable results can be obtained

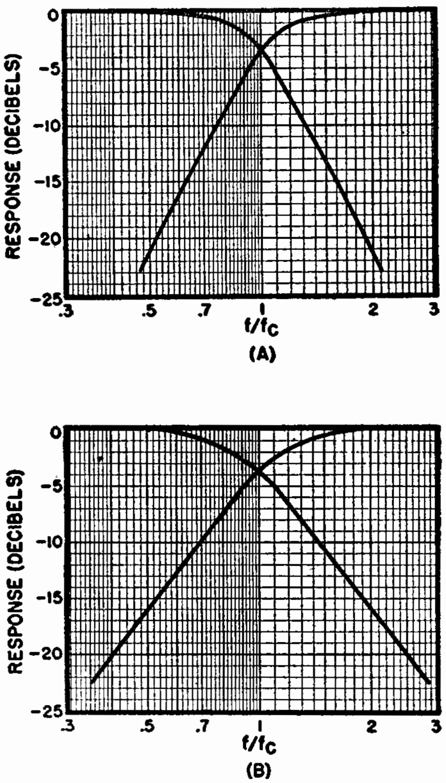
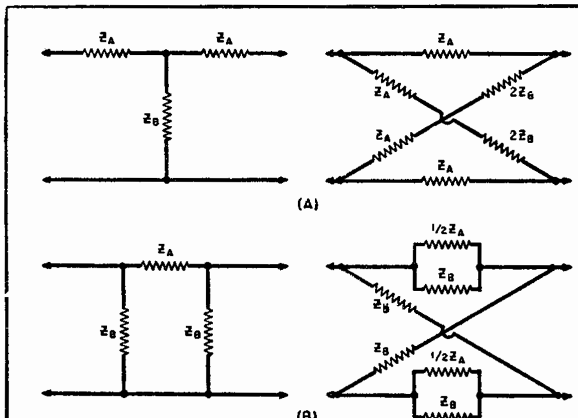


Fig. 16-16. Response curves of networks shown in Table 9. (A) is for the upper two networks, and (B) is for the lower two.

economically and in a small space. The circuit which gives these characteristics is the resistance-capacitance parallel-T network which is shown in Fig. 16-20, together with the design equations which determine its performance. When the parameter K in this circuit is assigned the value ½, a sharply selective cusp-shaped curve is obtained and there is a null for the desired frequency. When other values are assigned to K the curve approaches that of an L-C tuned circuit. The amplitude and phase characteristic of the circuit for various values of K are shown in the sets of curves in Fig. 16-15.

The transfer characteristic of this type of network, exhibiting high attenuation at the one resonant frequency, may be made to serve many functions where low cost or space requirements are an important factor. Wherever a dip may be desired in the response at some particular frequency, a parallel-T network may be inserted directly in the circuit to give the required attenuation and frequency response characteristic. When a peak in the response is desired at some particular frequency, this network may be used in a feedback circuit in an amplifier to give reduced feedback (and hence an increase in response) at the frequency to which it is tuned.

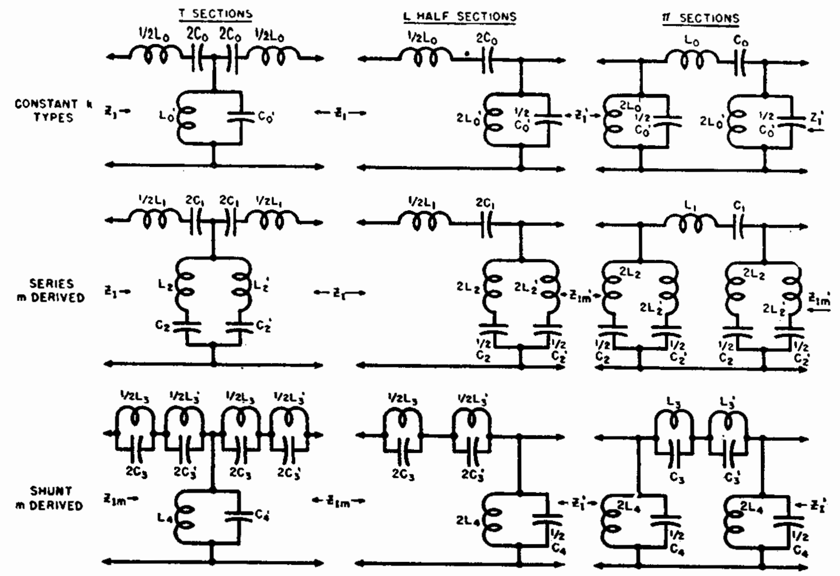


$Z_0 = \sqrt{Z_a Z_b}$ = image impedance

$$\tanh\left(\frac{\theta}{2}\right) = \sqrt{\frac{Z_a}{Z_b}}$$

θ = Image transfer constant

Fig. 16-17. (A) Lattice equivalent of T section. (B) Lattice equivalent of π section. At right, performance equations for general symmetrical lattice networks.



$$g = \sqrt{\left(1 - \frac{f_{1R}^2}{f_1^2}\right)\left(1 - \frac{f_{1R}^2}{f_2^2}\right)}$$

$$h = \sqrt{\left(1 - \frac{f_1^2}{f_{2R}^2}\right)\left(1 - \frac{f_2^2}{f_{2R}^2}\right)}$$

$$a = \frac{(1 - m_1^2) f_{2R}^2}{4 g f_1 f_2}\left(1 - \frac{f_{1R}^2}{f_{2R}^2}\right)$$

$$= \frac{(1 - m_2^2) f_1 f_2}{4 g f_{1R}^2}\left(1 - \frac{f_{1R}^2}{f_{2R}^2}\right)$$

$$b = \frac{1 - m_2^2}{4g}\left(1 - \frac{f_{1R}^2}{f_{2R}^2}\right)$$

$$c = \frac{(1 - m_1^2)}{4h}\left(1 - \frac{f_{1R}^2}{f_{2R}^2}\right)$$

$$d = \frac{(1 - m_1^2) f_{2R}^2}{4 h f_1 f_2}\left(1 - \frac{f_{1R}^2}{f_{2R}^2}\right)$$

$$= \frac{(1 - m_2^2) f_1 f_2}{4 h f_{1R}^2}\left(1 - \frac{f_{1R}^2}{f_{2R}^2}\right)$$

When $m_1 = m_2$:

$g = h$, $a = d$, $b = c$,

$$f_{1R} = \frac{f_1 f_2}{f_{2R}}, \quad m_1 = m_2 = \frac{h}{1 - (f_1 f_2 / f_{2R}^2)}$$

$$f_{2R}^2 = \frac{f_1^2 + f_2^2 - 2m^2 f_1 f_2}{2(1 - m^2)} + \sqrt{\left(\frac{f_1^2 + f_2^2 - 2m^2 f_1 f_2}{2(1 - m^2)}\right)^2 - f_1^2 f_2^2}$$

R_0 = load resistance

f_1 = lower freq. limit of pass band

f_2 = higher freq. limit of pass band

f_{1R} = a freq. of high attentuation in the low freq. attentuating band

f_{2R} = a freq. of high attentuation in the high freq. attenuating band

$$L_0 = \frac{R_0}{\pi(f_2 - f_1)}, \quad L_0' = \frac{(f_2 - f_1) R_0}{4\pi f_1 f_2}$$

$$L_1 = m_1 L_0, \; L_2 = a L_0, \; L_2' = c L_0$$

$$L_3 = \frac{1}{b} L_0', \; L_3' = \frac{1}{d} L_0', \; L_4 = \frac{1}{m_2} L_0'$$

$$C_0 = \frac{f_2 - f_1}{4\pi f_1 f_2 R_0}, \quad C_0' = \frac{1}{\pi(f_2 - f_1) R_0}$$

$$C_1 = \frac{1}{m_2} C_0, \; C_2 = \frac{1}{b} C_0, \; C_2' = \frac{1}{d} C_0$$

$$C_3 = a C_0', \; C_3' = c C_0', \; C_4 = m_1 C_0'$$

$$m_1 = \frac{(f_1 f_2 / f_{2R}^2)\, g + h}{1 - f_{1R}^2 / f_{2R}^2}, \quad m_2 = \frac{g + (f_{1R}^2 / f_1 f_2) h}{1 - f_{1R}^2 / f_{2R}^2}$$

Table 7. Information for the design of band-pass filter sections. Additional equations to go with this table are given in the text.

By using a considerable amount of negative feedback applied through a parallel-T network, so that there is high degeneration in the amplifier at all frequencies except those near the null frequency of the network, a band-pass amplification curve will be attained which closely approximates the usual *L-C* tuned-circuit characteristic. Values of K between 0.2 and 0.5 are found to give the best results in this type of application. Oscillators can be designed without the necessity of using tuned circuits by making use of the feedback principle. By using the parallel-T network (with a value of $K = 0.5$ to give a complete null at the resonant frequency) for negative feedback and a certain amount of positive feedback to cause oscillation, audio oscillators can readily be constructed which are stable and have very low distortion.

The manner in which the parallel-T network may be used in a circuit to give the various frequency response characteristics which have been described is indicated in the block diagram in Fig. 16-19. The network is used in general in a high impedance circuit, whether it is in the feedback circuit or inserted directly in the signal circuit. By proper choice of the resonant frequency and the amount of attenuation at resonance, the use of parallel-T networks can give the wide variety of frequency response characteristics which have been described. This type of circuit can be used to give low-frequency or high-frequency lift or attenuation, and is also capable of giving high-pass, low-pass or band-pass frequency characteristics.

When the parallel-T network is used in the manner described, it gives frequency characteristics which are often more desirable for many applications than the standard *R-C* lift and attenuation characteristics. There are also certain applications, particularly in high-impedance circuits at low frequencies, where the use of *L-C* resonant circuits is impractical. Normally, however, the parallel-T network is not capable of attaining the results which are possible with good *L-C* equalizers and wave filters, but is widely used for reasons of economy and space requirements.

There are certain audio-frequency network applications where it is necessary to alter the phase or time relations of the various frequency components of the signal without altering the amplitude relations. Circuits which perform this function are known as all-pass networks. An all-pass filter has zero attenuation for all frequencies within the range which is of interest, and can be designed to have the desired phase shift characteristics so that it may be used as a phase equalizer or to correct phase distortion introduced by other parts of a system. It can also be used to introduce a time delay in the transmission of a signal.

In general, phase distortion in a network occurs when the relative phases of various components of the signal are not the same in the output as in the input. For any individual frequency component, the difference in phase between the output and input voltage can be expressed in terms of the time of transmission through the circuit according to the relation

$$\text{Phase Shift} = \mathrm{r}\,\omega + n\,\pi \text{ radians}$$

where n is an integer, $\omega = 2\pi f$, and τ is termed the delay time of the network. In order for the network to have zero phase distortion, τ must remain constant for all frequencies. In other words, the curve of phase shift as a function of frequency must be a straight line passing through an integral multiple of π at zero frequency. If this condition is satisfied, then any voltage component

$$e = E_1 \sin\,(\omega t + \beta)$$

of the input signal will be represented by $e = E_2 \sin\,[\omega\ (t\ \tau) + \beta]$ in the output. Thus, when β is constant the relative phases of all components of the signal are undistorted and the signal remains unchanged except for a time delay τ, a possible phase reversal of 180° (depending upon the value of the integer n), and whatever amplitude relations it was desired to introduce by the transmission network.

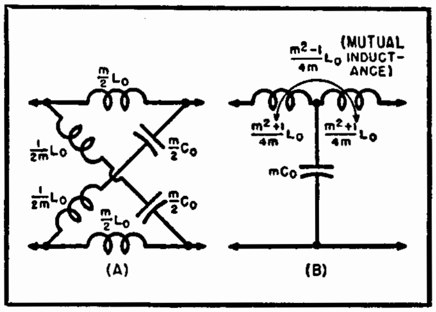


Fig. 16-18. Equivalent lattice (A) and T (B) m-derived low-pass sections when m is greater than unity which may be used for time delay.

The characteristics of the all-pass phase-correction network cannot in general be realized by the ladder-type networks which have been described, but must instead make use of the properties of the more general lattice network. The lattice structure is the most general type of symmetrical filter section that can be devised, and includes the T and π ladder sections as special cases. Almost all of the network problems encountered in general engineering practice can be solved by the use of ladder sections, and lattice networks usually need not be considered. However, the exception is in the case of phase-compensation and time-delay networks, and here the properties of lattices must be used.

The relationship between ladder and lattice networks may be seen from Fig. 16-17, which shows the transformations necessary to convert any symmetrical T or π section to an equivalent lattice and the performance equations for the general symmetrical lattice network. It can be seen that any symmetrical T or π section can be converted to a lattice, but that the reverse transformation of a lattice to a T or π is not necessarily possible. Thus, the transformation from a lattice to a symmetrical T will result in a physically realizable structure only when it is possible to subtract the series arm Z_a of the lattice from the diagonal and have a physically realizable remainder; and the conversion from a lattice to a symmetrical π can be performed only when it is physically possible to subtract the admittance $1/Z_b$ of the diagonal lattice impedance from the admittance of the series impedance arm and have a physically realizable remainder.

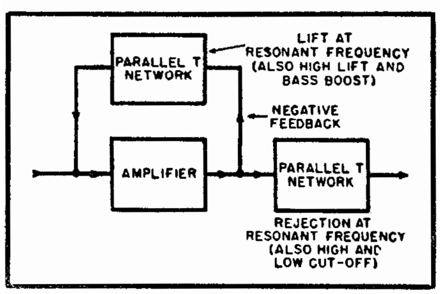


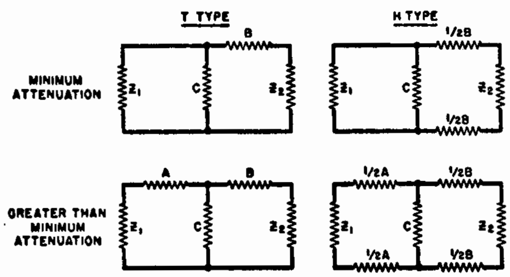
Fig. 16-19. Method of using parallel-T networks with an amplifier to give various frequency-response characteristics.

An all-pass network can be obtained by making use of a lattice network in which Z_a and Z_b are reactances which are reciprocal with respect to the desired image impedance R_o. This gives a pass band in which the attenuation constant α is zero, and an image impedance that is a constant resistance R_o for all frequencies. Under these conditions the phase shift β is given by:

$$\tan\frac{\beta}{2} = -j\sqrt{\frac{Z_a}{Z_b}} = \pm j\sqrt{\frac{Z_a^2}{R_o^2}}$$

The variation of the phase shift β with frequency is determined by the number and location of the internal zeros and poles in the impedance Z_a, and by the value of the reactances. Typical phase-shift characteristics for two different networks of this type are given in Fig. 16-14. It may be noted by comparison with Fig. 16-17 that these networks have no T or π section equivalents.

A type of network which is widely used for providing time delay is the m-derived low-pass filter in which the value of m is made greater than unity. In this network, it happens that although the equivalent T section requires a negative inductance in series with a capacitance, this can physically be realized by the use of mutual inductance. The manner in which this is accomplished is seen in Fig. 16-18, which shows the m-derived lattice and its equivalent T section where m is greater than unity. The low-pass π



$$R = \frac{I_2}{I_1} = \sqrt{\frac{Z_2}{Z_1}} = \sqrt{\frac{Z_2 - Z_1}{Z_1}}$$

$$A = \frac{2Z_1 - 2R\sqrt{Z_1Z_2}}{1 - R^2} - Z_1$$

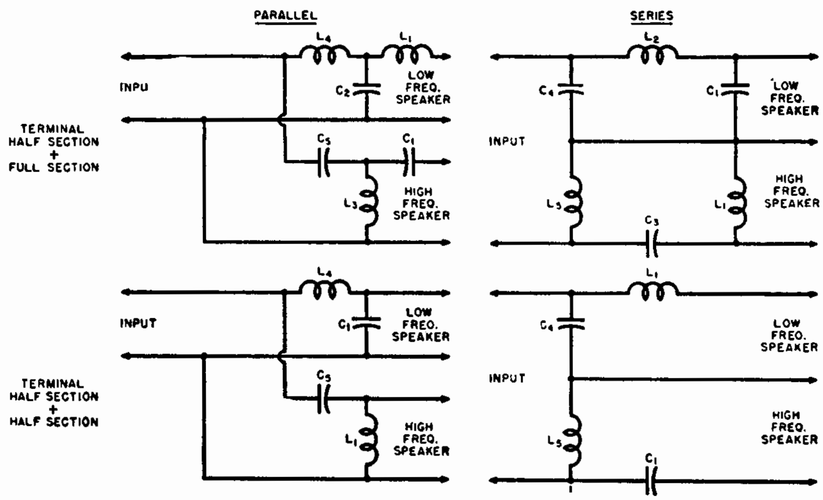
$$B = \frac{2Z_2 - 2R\sqrt{Z_1Z_2}}{1 - R^2} - Z_2$$

$$C = \frac{2R\sqrt{Z_1Z_2}}{1 - R^2}$$

For top circuits, $Z_2 > Z_1$

For bottom circuits, $Z_1 \gtrless Z_2$

Table 8. Information for the design of impedance-matching networks.



R_0 = speaker resistance

f_c = cross-over frequency

m = design constant = 0.6

$$L_1 = \frac{R_0}{2\pi f_c},\ L_2 = 2L_1,\ L_3 = \tfrac{1}{2}L_1$$

$$L_4 = (1 + m)\,L_1,\ L_5 = \frac{1}{1 + m}\,L_1$$

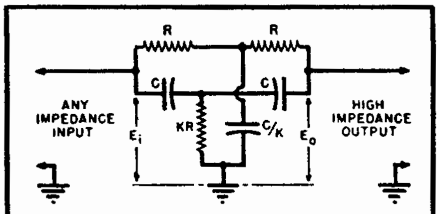
$$C_1 = \frac{1}{2\pi f_c R_0},\ C_2 = 2C_1,\ C_3 = \tfrac{1}{2}C_1$$

$$C_4 = (1 + m)\,C_1,\ C_5 = \frac{1}{1 + m}\,C_1$$

Table 9. Information for the design of frequency-dividing networks.

section is not physically realizable when m is greater than unity, since it requires always a negative capacitance and is unrealizable without the aid of negative resistance.

The characteristics of this type of time-delay network are given in Table 10 in a form which is convenient for practical calculation. The curves for the delay characteristics for different values of m show that the curve which has been drawn for $m = 1.27$ offers the flattest possible delay characteristic up to about $0.55\omega_o$. Therefore, the procedure in designing a delay line to have a given time delay over some definite frequency band is to determine the time delay for a single low-pass section having a value of $m = 1.27$ and a cut-off frequency ω_o equal to 1/0.55 times the highest frequency which is of interest, and then to design a sufficient number of such sections in tandem to give the required delay. This network may then be matched to other filter networks in the conventional manner, and in particular may be terminated in half-sections having a value of $m = 0.6$ to obtain better matching to a resistive load. The choice of whether to use a lattice or the equivalent T section as indicated in Fig. 16-18 will generally depend upon considerations of economy and facility of construction, since both types yield identical results.



$$E_0/E_i = \beta,\ R = \frac{1}{2\pi f_R C}$$

$$\beta = \frac{2K^2 - K}{2K^2 + K + 1} \text{ at freq. } f_R$$

f_R = resonant frequency

Fig. 16-20. Basic parallel—T R-C filter.

Impedance Matching

Because of the importance of the terminating impedances in determining the performance of electrical networks, it is often necessary to be able to change the impedance of a network in order to match it properly to another network. A simple and convenient method of matching unequal impedances is by means of resistive impedance matching networks.

The circuits which are most generally used for this type of service are the T and H type networks, where the T network is used principally in unbalanced circuits and the H in balanced circuits. The basic circuit configurations of these networks, and the formulae for their design, are given in the chart in Table 8. From the design equations it can be seen that whenever resistive networks are used for impedance transformation the network introduces a certain amount of attenuation, while there is no such loss when a transformer is used. However, resistance networks have the advantages of lower cost and better frequency response, and the loss in level is not generally a major consideration since additional gain in amplifier circuits is relatively easy to obtain.

The matching networks shown in the table may be used both to match input and output impedances, and at the same time to introduce any desired amount of attenuation. When the networks are used only to match two unequal impedances, with minimum insertion loss, the U or L type networks should be used. These are the limiting cases of the T and H networks operating between unequal impedances, with the series elements on the side of the smaller impedance made zero to give minimum loss. The minimum loss is given by the equation:

$$R = \frac{I_2}{I_1} = \sqrt{\frac{Z_2}{Z_1}} - \sqrt{\frac{Z_2 - Z_1}{Z_1}}$$

where $Z_2 > Z_1$, and is seen to depend only upon the relative values of the terminating impedances. When greater loss is desired, the T or H type will give the required attenuation with matched impedances.

Networks of this type may also be of considerable value in many filter appli-

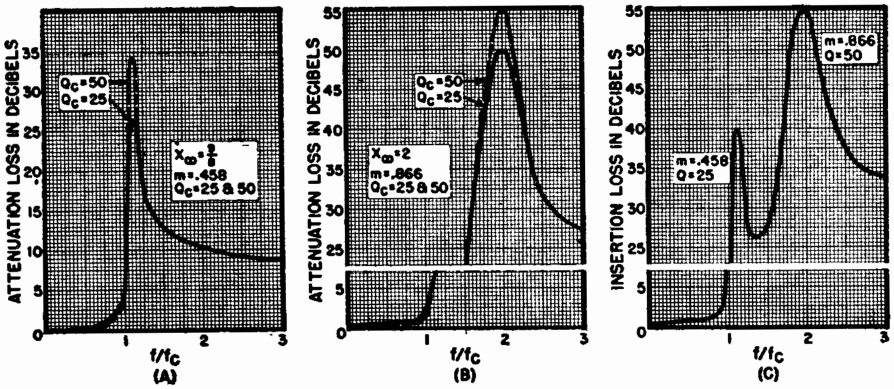
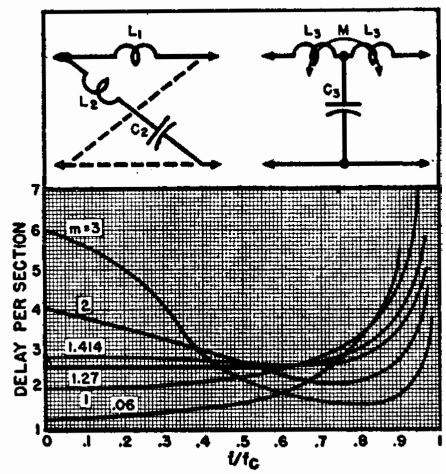


Fig. 16-21. (A) and (B) Attenuation curves of two different networks with different values of m, for two values of Q. (C) Attenuation characteristic of a composite filter made up of two of the sections shown in (A) and (B).

cations where there may be undesirable interactions between tandem filter sections. In such cases a resistive T or H attenuator with $Z_1 = Z_2$ (and having for instance 6 db attenuation) will generally be found to provide sufficient isolation between the two filters.

The theoretical curves and data which are customarily used in the design of electrical networks are always given in terms of dissipationless circuit elements. However, in practice it is found that circuit elements have appreciable losses, which prevent the theoretical curves from ever being attained. The customary procedure in network design is to base the design upon the lossless structure, modified according to the knowledge of what will be the effects of circuit dissipation upon the theoretical design.

In audio circuits the major losses are generally found to be in the inductances, which cannot easily be built to have high Q at these frequencies. By winding inductances in a toroidal form, using molybdenum permalloy powder cores, it is possible to attain Q's of the order of 100 to 200 at frequencies of about 3000 cycles. However, these coils are expensive and difficult to wind, and their Q is considerably lower at the low audio frequencies. These high-Q coils are used for networks which must be constructed to critical specifications and close tolerances, but in most practical work lower



R_0 = load resistance = $\sqrt{L_0/C_0}$

f_c = cut-off frequency = $1/\pi\sqrt{L_0C_0}$

$$L_0 = \frac{R_0}{\pi f_c},\ L_1 = \frac{m}{2}L_0,\ L_2 = \frac{1}{2m}L_0$$

$$L_3 = \frac{m^2+1}{4m}L_0,\ M = \frac{m^2-1}{4m}L_0$$

$$C_0 = \frac{1}{\pi f_c R_0},\ C_2 = \frac{m}{2}C_0,\ C_3 = mC_0$$

$$T = \frac{2m}{\sqrt{1-(f/f_c)^2}\,[1-(1-m^2)(f/f_c)^2]}$$

Table 10. Design of time-delay networks. To obtain time delay per section in seconds, divide by $2\pi f_c$.

Q's of the order of 25 to 50 are generally quite acceptable.

The general effect of dissipation in a two-terminal network is to raise the minimum impedance attained, and to lower the maximum value of impedance. Thus, an inductance at low frequencies will approach a definite resistive value instead of zero, and at high frequencies will remain finite instead of approaching infinity. A series resonant circuit does not have zero impedance at its resonant frequency when dissipation is present, and parallel-resonant circuits have finite impedances at resonance. By taking these factors into account, the deviations from the theoretical dissipationless case, which are encountered when coils (or condensers) having low Q are used in practical designs, can be estimated without too much difficulty.

The effects of dissipation in attenuation equalizers are:

(1) the equalization of the actual network is less than the theoretical design value because of the finite minimum and maximum impedances,

(2) there is some mis-match between the equalizer and its termination, since the dissipative resistance tends to alter somewhat the reciprocal relationship between the series and shunt impedance arms.

The effects of incidental dissipation losses in wave filters are somewhat more complex, due to the increased complexity of the attenuation characteristics. These effects may be summarized briefly as follows:

(1) The theoretically infinite peaks in the attenuation loss and the reflection loss remain finite, depending upon Q and upon the closeness of the peak to the cut-off frequency.

(2) The attenuation is not zero in the passband, and phase shift is different from the dissipationless value.

(3) At the cut-off frequency, the attenuation is not zero, as in the dissipationless case, and the curve is rounded at the cut-off frequency.

Curves showing the effects of dissipation on the frequency characteristics of wave filters are given in Figs. 16-21A, B, and C. Figs. 16-21A and B show the attenuation curves of two networks with different values of m for two different values of Q. In Fig. 16-21 is given the characteristic of a composite filter made up of two of the sections shown in Figs. 16-21A and B.

From the information in these curves, it may be seen that with Q's of the order of 25 to 50, audio frequency networks may be designed according to the information given in this data with very little modification to correct for circuit dissipation.

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CHAPTER 17

Tone Control (Equalizers)

A discussion of R-C and Resonant systems, Record Equalizers and Special Program Equalizers.

Resistance-Capacitance (R-C) Networks

● Nearly all of the simpler means for tone control or shaping the response curve are based on simple high and low pass filter circuits. For example, the circuit of Fig. 17-1A is a simple high pass filter while that of Fig. 17-1B is a low pass system.

In operation, the output from the circuit of Fig. 17-1A will increase with increasing frequencies applied to the input, while that from Fig. 17-1B will decrease as the frequency is increased. This is caused by the fact that the reactance of *C* decreases with increasing frequency and therefore, a smaller portion of the applied voltage appears across the condenser as the frequency increases. Theoretically the above action is true from zero frequency (dc) to infinitely high frequency, but for all practical purposes the effect is restricted to a small section of the spectrum. The change in response is negligible beyond the frequency limits at which the condenser reactance is equal to 1/10 and 10 times the value of resistance in the circuit. There are many variations of these two simple circuits which are used for treble or bass attenuation.

Typical Tone Compensation Systems

To evaluate any form of tone control used in connection with an amplifier it is necessary to establish a reference point for both gain and frequency. For this discussion we will consider the gain of the amplifier at 400 cycles-per-second as the reference point, and all circuits mentioned will be classified accordingly as low or high frequency boost or attenuation systems.

The circuit of Fig. 17-2 is perhaps the most commonly used form of tone control but is seldom recognized as such. It consists of nothing more than a coupling condenser and the following grid resistor. By proper selection of the values of *C* and *R* this circuit may be adjusted to give negligible attenuation at the reference frequency and increasing attenuation as the frequency is reduced.

Due to other design considerations in the amplifier, this form of tone control is made variable only on rare occasions.

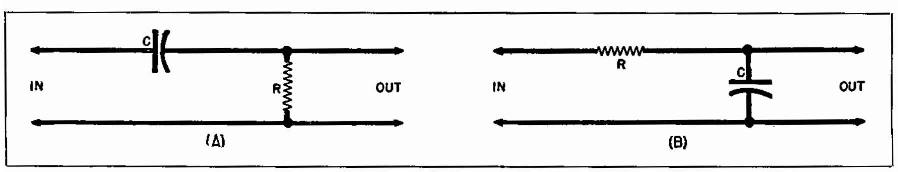


Fig. 17-1. (A) A simple high pass filter. (B) A commonly used low pass system.

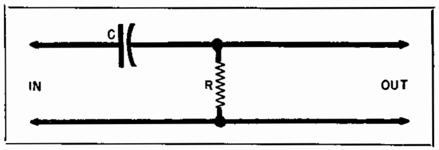


Fig. 17-2. Commonly used form of tone control.

The opposite of the action secured from the circuit of Fig. 17-2 may be obtained by the use of either Fig. 17-3A or 17-3B. Both of the circuits operate on the principle that the total impedance in the circuit increases as the frequency is reduced, thus forming a load impedance which varies inversely with the frequency. The circuit of Fig. 17-3A is most commonly used as the plate load for voltage amplifier tubes whose operating conditions are so adjusted that the stage gain increases as the plate load impedance rises. The circuit of Fig. 17-3B is commonly used with a tapped volume control and may be inserted as a second detector load impedance or as the grid return circuit of an amplifier stage. When the circuit of Fig. 17-3A is used as a microphone or pick-up load impedance the dotted line shown is used as the high output lead. When connected in this manner the condenser shunts the output terminals and therefore increases the ratio of output to input impedance as the frequency decreases. This circuit could be considered high frequency attenuation just as well as low frequency boost since the operation is the same.

Both circuits of Fig. 17-3 may be made variable in response by using different values of C selected by a switch or other means.

Fig. 17-4A and 17-4B show the most commonly used means of providing tone control on the upper portion of the audio spectrum. The most frequently used position for these circuits in an amplifier is the grid circuit of one or more of the amplifier tubes.

The circuit of Fig. 17-4A is used quite frequently by radio receiver manufacturers to give a smooth tone control which operates on the high frequency end of the audio passband.

Condenser C is usually selected so that when R_2 is zero the high frequency attenuation starts at a frequency near the reference point. As R_2 is increased C becomes increasingly less effective in bypassing the high frequencies, thus providing smooth control of the high frequency response of the amplifier. In the circuit of Fig. 17-4B both R_1 and R_2 are usually relatively high values of resistance and C a small capacitance. In this circuit the output at and below the reference frequency is approximately equal to $E_{in}R_2/(R_1 + R_2)$ and increases to a value approaching E_{in} as the frequency increases. When properly designed, all the foregoing circuits attenuate certain frequencies while maintaining normal response from the reference frequency to the opposite end of the band.

Fig. 17-5A and 17-5B show circuits similar to the above but applied in a manner which gives a stage gain variation with frequency. This is accomplished by connecting the filter section into a feedback circuit. Fig. 17-5A will

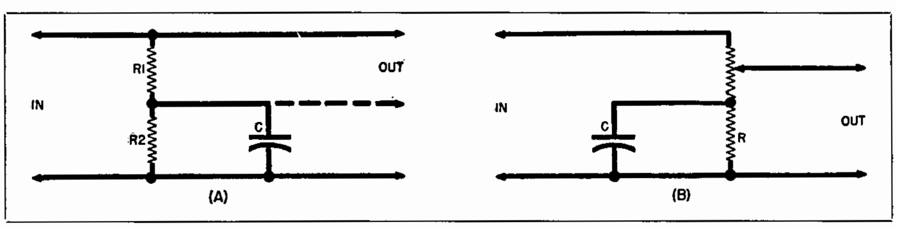


Fig. 17-3. (A) Circuit commonly used as the plate load for voltage amplifier tubes. (B) Circuit often used with a tapped volume control. This circuit may be inserted as a second detector load impedance in amplifier.

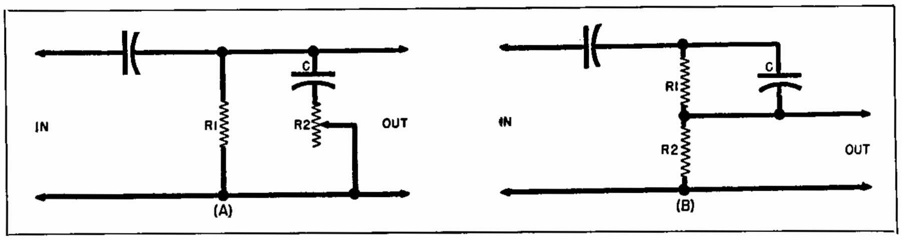


Fig. 17-4. (A) Circuit used in radio receivers to give smooth tone control on the high frequency end of the audio passband. (B) Another frequently used circuit which provides tone compensation for commercially-built receivers.

boost the higher frequencies when condenser C is chosen to have a value insufficient to provide an adequate bypass for cathode resistor R at the reference and lower frequencies. At the higher frequencies C becomes an effective bypass and thus removes degeneration from the tube circuit.

When properly chosen values of R_1, R_2 and C are used in the circuit of Fig. 17-5B, the stage may be made to have considerable inverse feedback at the high frequencies and relatively little at the reference and low frequencies. Condenser C should be so chosen that its reactance at the low frequencies is high with respect to the total impedance in the series-parallel combination of R_1, R_2 and R. Resistor R_1 is used to limit the amount of feedback at the high frequencies and R represents the impedance in the driving source or, in practice, the output impedance of the preceding stage (i.e., R_b and R_p of the preceding stage in parallel). The circuit of Fig. 17-5B must be carefully designed since the R-C network may give sufficient phase shift to the feedback signal for the stage to become regenerative at an unwanted frequency.

A word about the position of tone control systems in an amplifier may be in order at this time. Since nearly all of the simple tone control circuits described may be classified as the opposite of the name applied, care must be taken that a low frequency boost circuit is not followed by a high frequency boost circuit which operates by reducing the reference and low frequencies. For example, the circuit of Fig. 17-3A or 17-3B should not be followed by the circuit of Fig. 17-2A since these two circuits will tend to cancel out and produce flat response. In most phono and microphone amplifiers some bass boosting and, for high fidelity, some treble boost is desirable. Bass boost may be obtained by using the circuit of Fig. 17-3A in the

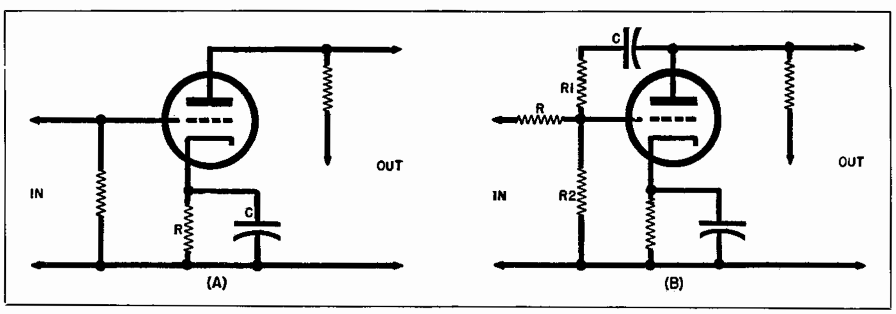
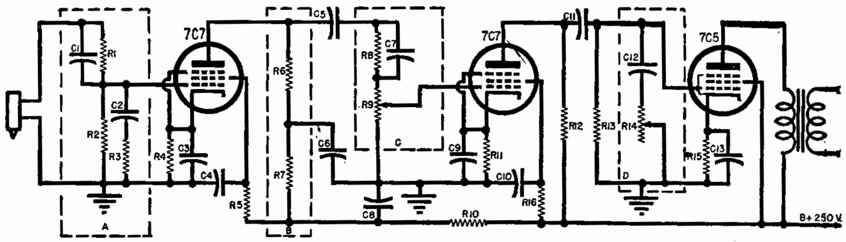


Fig. 17-5. (A) This circuit will boost the higher frequencies when condenser C is of too low a value to provide an adequate bypass for the cathode resistor at the reference and lower frequencies. (B) By choosing component values, this circuit can provide inverse feedback on the highs.



R_1—1.5 megohm, ½ w res.
R_2—3.3 megohm, ½ w. res.
R_3, R_6—100,000 ohm, ½ w. res.
R_4—1000 ohm, 1 w. res.
R_5—1.2 megohm, ½ w. res.
R_7—150,000 ohm, ½ w. res.
R_8, R_{13}—250,000 ohm, ½ w. res.
R_9—250,000 ohm, var. res.
R_{10}—47,000 ohm, ½ w. res.
R_{11}—470 ohm, 2 w wirewound res.
R_{12}—68,000 ohm, ½ w. res.
R_{14}—1 megohm var. res.
R_{15}—360 ohm, 2 w. wirewound res.
R_{16}—330,000 ohm, ½ w. res.
C_1—100 μμfd. mica cond.
C_2, C_7—.001 μfd., 400 v. cond.
C_3—25 μfd., 25 v. elec. cond.
C_4—.1 μfd., 400 v. cond.
C_5, C_{11}—.05 μfd., 400 v. cond.
C_6—.01 μfd., 400 v. cond.
C_8—8 μfd., 350 v. elec. cond.
C_9, C_{12}—50 μfd., 25 v. elec. cond.
C_{10}—.25 μfd., 400 v. cond.
C_{13}—.005 μfd., 400 v. cond.

Fig. 17-6. A pentode input amplifier which provides bass boost action. See Table.

input grid circuit (with the dotted line connected to the input grid) or the same circuit in the plate load for a pentode input amplifier as shown in Fig. 17-6. Treble boost may be obtained by using the circuit of Fig. 17-4B in the following grid circuit or the circuit of Fig. 17-5A in the cathode circuit of the following tube. If treble attenuation is desired, the circuit of Fig. 17-4A may be used in the grid circuit of the output tube. With push-pull output stages the condenser C and resistor R_2 of Fig. 17-4A may be connected from grid to grid. These circuits are also shown in Fig. 17-6.

In designing an amplifier having any of the bass or treble boost circuits described it must be remembered that the over-all gain of the amplifier is reduced at the reference frequency. Therefore, sufficient additional gain at the reference frequency must be included in the design to make up for that lost in the tone control circuits. The design of the amplifier must also provide for the additional gain which occurs at the boosted frequencies so that overloading, with resulting distortion at these frequencies, does not occur.

In the circuit of Fig. 17-6, a slight modification has been added in Section *A* which was not covered in the description of Fig. 17-3A. In Fig. 17-3A the low frequency boost is obtained at the cost of high frequency attenuation which becomes quite extensive as the frequency approaches the high end of the audio band. The addition of the 100 uufd condenser and the 100,000 ohm resistor in the circuit limits the high frequency attenuation of this circuit to approximately the attenuation at the reference frequency. The resistor and condenser values used in this circuit are found by calculation of admittances and impedances at each of several frequencies in the band desired, including the reference frequency.

These calculations, using the values assigned, indicate that the signal applied to the grid of the input tube, assuming constant voltage from the pickup, will be approximately 61% at 50 cps, 33% at 100 cps, 21.5% at 400 cps and 21.8% at 4000 cps.

Sections *B* and *C* in Fig. 17-6 both affect the impedance of the plate load and therefore the gain of the input Type 7C7 tube. The gain at each of several frequencies in the desired band was obtained by calculating the plate load impedance at each frequency and interpolating the gain from the *R-C* data given in a late *Sylvania* Technical Manual. At frequencies of 50, 100, 400, and 4000 cycles-per-second the gain of the input stage (assuming constant grid signal) will be approximately 164, 159, 145, and 120 respectively.

Section D has been used in the circuit of Fig. 17-6 to illustrate the most common form of high frequency attenuation circuit. The condenser value has been chosen to give a negligible effect at the reference frequency when the resistance, R_{14}, is set at zero. The plate resistor, R_{12}, for the second amplifier has been assigned a relatively low value so that the tube operation approximates a constant current generator. Under these conditions the voltage across the total plate impedance consisting of C_{12}, R_{12}, R_{13}, and the tube R_p in parallel is equal to IZ with Z variable with frequency. Using the values assigned, the signal reaching the output tube grid will approximate 100%, 100%, 95%, 88%, and 35% of the low frequency value, at the frequencies of 50, 100, 400, 800 and 4000 cycles-per-second when the 1 megohm control is set for maximum high frequency attenuation. When the 1 megohm control resistance is all in the circuit no appreciable high frequency attenuation will be introduced and the response will be that caused by sections A, B, and C.

By summarizing the results described above it is possible to plot the approximate response curve of the circuit in Fig. 17-6 from the pickup to the grid of the output tube from calculations of the over-all gain at the several frequencies. See table below.

Attenuation of Section D in Fig. 17-6.

Frequency (in c.p.s)	Over-all Gain	Attenuation of Sec. D. Max.	Min.
50	.61 x 164 x .5 x 80 x 1		4000
100	.33 x 159 x .503 x 80 x 1	2115	2115
400	.215 x 145 x .542 x 80 x 1		1352
400	.215 x 145 x .542 x 80 x .95	1287	
4000	.218 x 120 x .865 x 80 x 1		1810
4000	.218 x 120 x .865 x 80 x .35	633	

With 400 cps as the reference the response with minimum attenuation in section D would be +9.4 db, +3.9 db, 0, and +2.54 db at frequencies of 50, 100, 400, and 4000 cycles per second. With section D set for maximum high frequency attenuation the response at the same frequencies would be +9.83 db, 4.32 db, 0, and —6.127 db, respectively.

In all of the above calculations and results reported, all effects of tube and stray capacities have been neglected and it has been assumed that the output from the pickup is constant over the frequency range. Since these conditions do not exist in actual practice the circuit shown in Fig. 17-6 should be accepted as a basic illustration of the circuits used for tone compensation and control systems, and may require some modification if these exact characteristics are desired in a practical design.

Reference: Sylvania News, Vol. 15, No. 1.

Resonant Equalizers

In previous paragraphs we discussed various types of (R-C) tone controls (equalizers) employing resistance and capacitance. More flexibility is had with the "resonant" type of equalizer, as it has been found from actual listener tests (see Fig. 17-7) that to obtain a balance between the low and high frequency response is highly desirable. For best performance, the product of

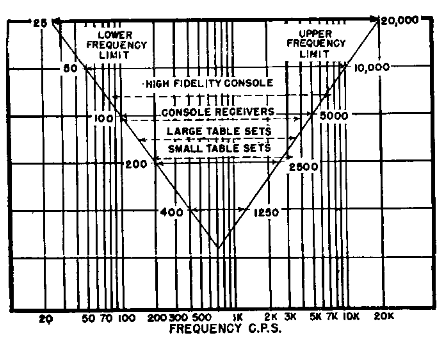


Fig. 17-7.

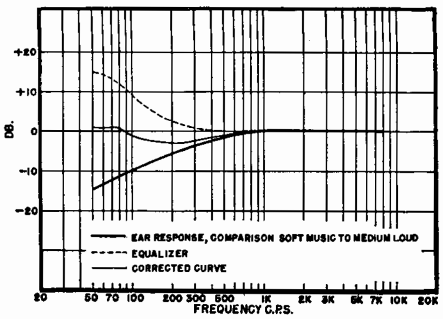


Fig. 17-8.

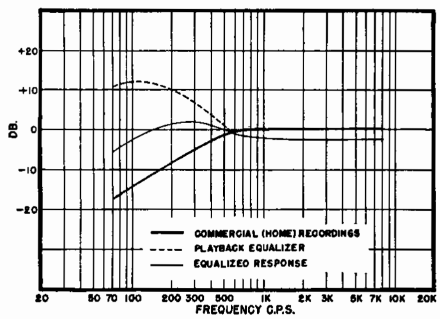


Fig. 17-9.

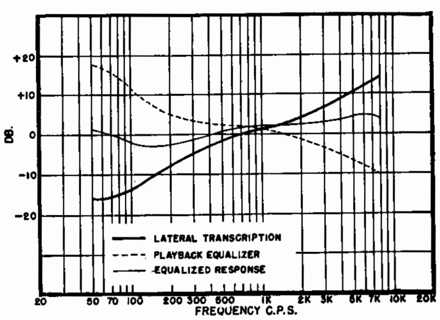


Fig. 17-10.

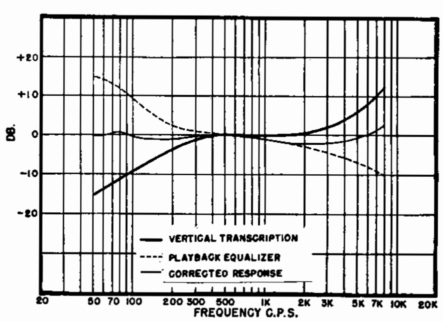


Fig. 17-11.

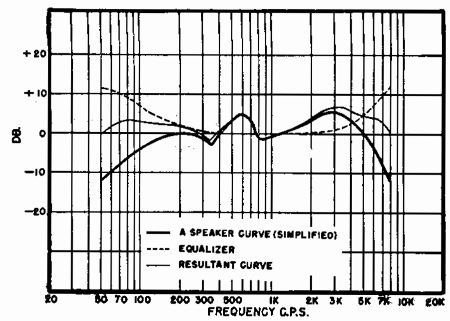


Fig. 17-12.

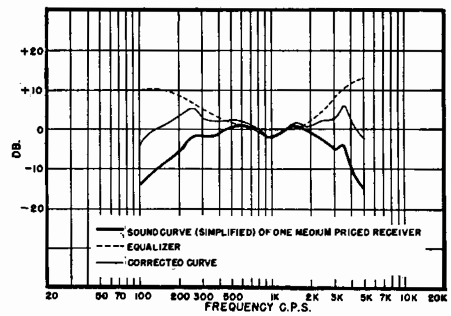


Fig. 17-13.

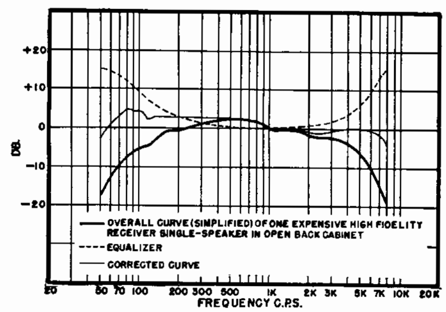


Fig. 17-14.

the low frequency and high frequency limits of the equipment should equal approximately 500,000.

It has been established that where an audio system in a receiver was good to only 200 cycles at the low frequency end a tone control was necessary and when it limited the high frequency end to 2500 cycles, the best balance of response was obtained. In the reproduction of phonograph records a similar condition exists, especially those records having a sharp drop-off at the low end. Accordingly, to obtain balance the high end frequencies are often removed unnecessarily. While the sound becomes balanced to the ear the fidelity is, unfortunately, greatly decreased. Low notes which were not reproduced originally are still not reproduced, and the high notes are missed as well. The obvious answer then is to bring back the low notes through the use of an equalizer, thus increasing the over-all fidelity rather than reducing it.

Virtually all recording systems, whether discs, sound on film, sound on wire, embossed film, etc., require playback equalization as well as equalization when the recording is being made. The necessity for equalization is apparent.

Most microphones, pickups and loudspeakers can be effectively equalized. Low frequency droop used for dialogue equalization will greatly improve the intelligibility of speech. It will also permit higher power levels from speakers used in pa equipment. A portion of the power normally going to the speakers and not required for intelligibility is removed.

Mid-frequency equalization (low end and high end droop) is frequently used by radio amateurs to effect a maximum signal level in the frequencies most necessary for intelligibility. Various acoustic conditions will frequently lend themselves to equalization. For example, the absorption of high frequencies may easily be 15 db depending upon the drapes in the room, the number of people, as well as the presence of any other sound absorbing material. Another point of equalizer use comes from the realization that at low sound levels the ear is less responsive to low frequencies. This type of equalization is commonly called "bass boost."

In order to obtain smooth and positive control and not introduce hum into the audio circuits it is advisable to employ commercially made equalizers such as the CGE-1 developed by *U.T.C.* engineers. Reference to the curves will show how well such units are suited to many applications. The electrical components are stable, free from hum pickup, and dependable in operation.

This type of equalizer is readily adapted for use with commonly used audio amplifier equipment. As mentioned, the frequency correction curve with the *RC* type of tone control is a gradual slope and does not accomplish what is required in the boost condition. For example, if the circuit to be equalized is down 15 db at 5000 cycles, an equalizer which brings back this 15 db but also boosts 6 db at 1000 cycles is not desirable. Accordingly, the CGE-1 unit employs resonant circuits for both the low and high frequency boost. As will be noted from the curves with 15 db. boost at 8000 cycles the response curve is flat at 2000 cycles. A similar condition exists at the low end. Two controls are required, one either boosts (accenuates) or drops (attenuates) the low frequencies and the other either boosts or drops the high frequencies. This type of equalizer is of high impedance and is designed for insertion into an audio amplifier between a triode plate and subsequent grid, or from a high impedance source (5000 to 30,000 ohms) other than a crystal microphone, to a subsequent grid loaded with 10,000 ohms.

Some insertion loss is effected by this equalizer, particularly at the maximum setting. If the amplifier system does not have excessive gain an additional audio stage may be required. As the filament and plate drain of the added tube is small, it can normally be taken from the power supply of the original equipment.

Figs. 17-8 through 17-19 illustrate the various typical curves obtained with this equalizer with the setting at maximum value.

To permit equalization for a wide variety of applications, the high frequency and low frequency boost sections are arranged for two frequencies each. In most applications the frequency desired is predetermined, and when the unit is wired into the equipment, the appropriate connections are employed.

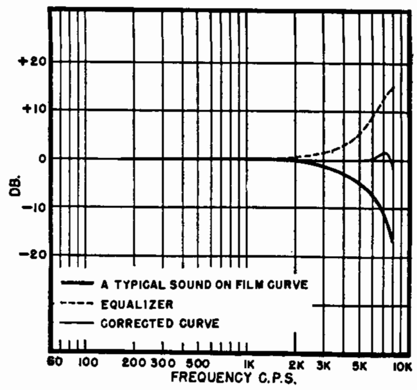


Fig. 17-15.

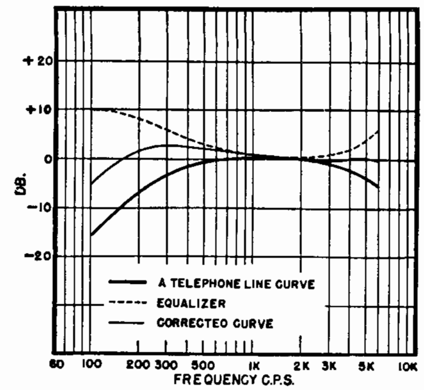


Fig. 17-18.

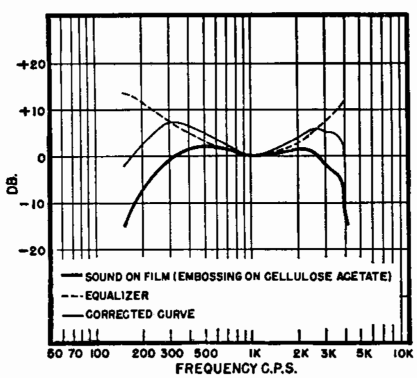


Fig. 17-16.

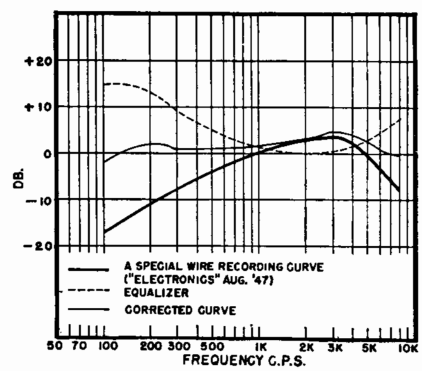


Fig. 17-19.

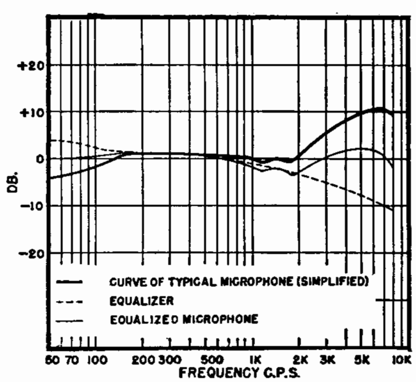


Fig. 17-17.

If a wide range of use is anticipated, the 50 cycle and 100 cycle terminals can be brought out to a single-pole, double-throw switch, and in like manner, the 5 kc terminals to a similar switch, thus permitting instantaneous changeover to the desired resonant frequency.

The following considerations should be observed in order to take full advantage of the possibilities of the equalizer:

1. The unit is designed to work between two impedances of 10,000 ohms, and this value of termination must be used to maintain accuracy of calibration.

2. No distortion is introduced if the maximum level at the equalizer input is held below 2 volts, with negligible distortion at several times this value. The unit should not be used at signal levels above 10 volts.

3. When adding this type of resonant equalizer to an amplifier with very little reserve voltage gain, a preampli-

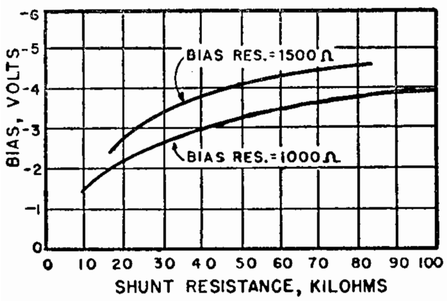


Fig. 17-20. Variation of bias with shunt resistance for phase-inverter of Fig. 17-23.

fier such as cascaded triodes will result in an output voltage essentially equal to the input voltage when both maximum bass and treble boost are used.

4. When an amplifier incorporating the CGE-1 is run with no boost, the mid-frequency gain is 30 decibels over that with full boost. For best signal-to-noise ratio under these conditions, a master gain control should be used in the circuit after the equalizer itself.

The Degenerative Tone Control

The degenerative type of tone control has enjoyed rather widespread use in audio amplifiers.[1] It has the particular advantage that only a single tube is required to accomplish both bass and treble boost and cut; this results in reduction of total amplifier stages to a minimum and simplifies the power-supply requirements when compared to other more complex controls.

On the other hand, as usually designed, the tone control makes use of an iron-core choke which is considered undesirable by many designers. Furthermore, when utilized in certain ways a parallel-resonant arrangement is introduced into the circuit and this, in the opinion of a large number of engineers, is to be avoided at almost any cost.

The usual objection to the use of an iron core choke coil for tone control is the possibility of hum pickup. With modern, well-shielded chokes available for this purpose, this is not a valid objection, and the hum introduced by the choke is negligible.

The basis of the degenerative tone control is the simple plate-and-cathode loaded phase inverter, incorporated in the circuit of Fig. 17-23. As is generally known, if R_1 and R_2 are equal the signal voltages at points A and B will also be equal but of opposite sign. The output is taken from point A and, in principle, a bass cut is attained by shunting this output with a suitable choke while a bass boost results if the cathode resistor is shunted. By substituting a condenser for the choke, the treble is similarly controlled. The maximum amount of cut in either case is 6 db per octave but the maximum boost depends upon the amplification factor of the tube. With a low-mu triode like the 6J5, which is customarily used, a boost of about 5 db per octave is the most that can be realized.

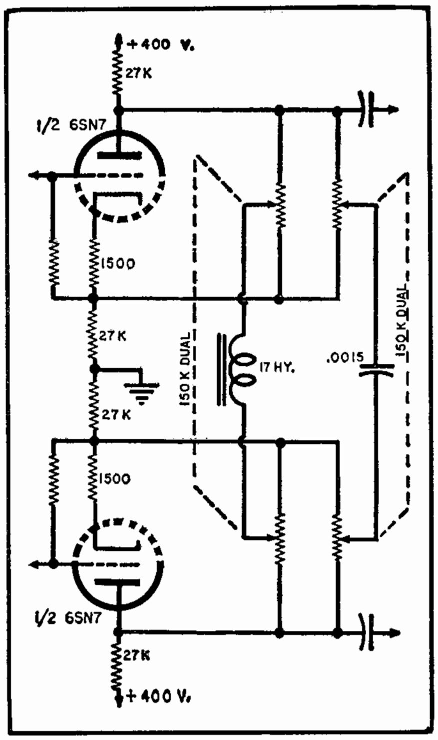


Fig. 17-21. Push-pull tone control circuit.

[1]Boegli, Charles P. "The Degenerative Tone Control," Radio and Television News, June 1951.

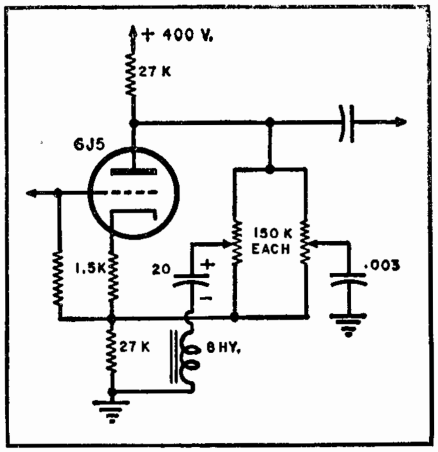


Fig. 17-22. Single-ended tone control circuit.

The first step in the design of the stage is thus simply the choice of component values for a suitable phase inverter. By way of example, a 6J5 will be assumed with a plate-supply voltage of 400, under which circumstances R_1 and R_2 may be 27,000 ohms each and the bias resistor may be 1000 ohms. The bypass is omitted from the cathode bias resistor with little effect.

The bass turnover frequency (the frequency at which bass boost or cut begins to become effective) and the treble turnover frequency may be independently specified. In the case of bass and treble cuts, the choke and condenser are shunted across the output resistance of the stage, which is substantially equal to R_1. Bass cut becomes effective when the reactance of the choke equals R_1; hence, if a 500 cps turnover is chosen with $R_1 = 27{,}000$ ohms, an 8.5 henry choke will be required. In a similar manner if treble cut is to begin at 2000 cps, an .003 μfd condenser will be needed. Boosts become effective at approximately the same frequencies because the cathode resistor is the same size as the plate resistor. The condenser and choke together in the above case resonate at 1000 cps, so that when maximum bass and treble cut are both employed, a 1000 cps parallel-resonant circuit is shunting the output.

In order to introduce each of these shunts independently to either the plate or cathode portion of the circuit, two controls must be used. There is more than one way to connect each control into the circuit, but the simplest method seems to be to attach one end directly to point A and the other to point B (Fig. 17-23). Since two controls are used, the tube then operates with a dc shunt equal to the parallel resistance of the two controls. This shunt naturally affects the bias voltage and some adjustment in the size of the bias resistor is necessary to

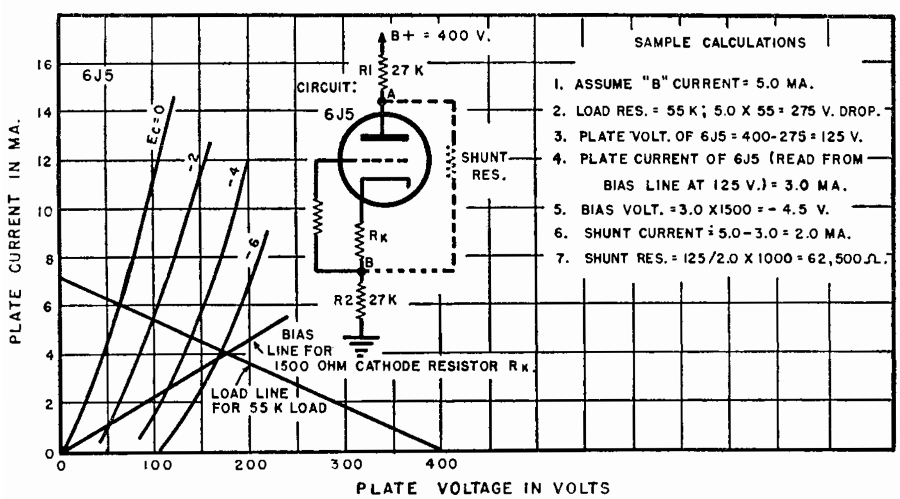


Fig. 17-23. Calculations used in the design of the shunted phase-inverter.

permit the shunted tube to handle the same signal voltage as an unshunted one. It would, of course, be possible to use a control of very high resistance, say, five megohms, in which case the effect on the dc voltages would be negligible. As the resistance of the control is increased, however, the region in which the control action takes effect becomes confined more and more to the ends of the rotation of the knob; with a five megohm control the entire boost or cut action occurs within a few degrees of the ends of this rotation, which effect is decidedly undesirable. It has for this reason been found preferable to choose a value equal to about five times the plate resistor of the inverter—in this case, around 100,000 or 150,000 ohms.

The dc voltages in a shunted-triode plate-and-cathode loaded phase inverter can easily be calculated, and the simplest method for finding the required bias resistor seems to be to assume a series of values of "B" supply currents, from which the voltage drop through the load resistors and hence the effective plate-cathode voltage across the tube can be found. With this voltage and a bias line drawn for a given bias resistor on the tube characteristics chart, the tube plate current and effective bias can be located. By subtracting this plate current from the assumed "B" supply current, the current flowing through the shunt is immediately found, and the tube plate-cathode voltage divided by this current equals the size of the shunt required to bring about the assumed operating conditions. Fig. 17-23 illustrates the method of calculation just described.

This procedure must be repeated for several assumed values of bias resistor, and the results plotted as shown in Fig. 17-20, which applies to the circuit used as an example in this problem. From this chart, it is evident that with 150,000 ohm controls, which impose a 75,000 ohm shunt across the tube, a bias resistor of 1500 ohms results in a grid bias approximately the same as that for an unshunted tube with a 1000 ohm bias resistor.

Since the signal voltages occurring at each end of the controls are equal in magnitude but opposite in sign, the center point of each control is effectively at ground ac potential even though no grounded center tap is provided. If the center point of the knob rotation is to correspond to flat response equal resistance must be provided each side of this center, which usually indicates the use of linear-taper potentiometers. Fader types have been tried but found to be unsatisfactory for this circuit. To obtain the control action the slider of the bass control is grounded through the choke whose size was previously calculated, and the other slider is connected to ground by means of the condenser.

In a single-ended stage (Fig. 17-22) the dc must be prevented from flowing through the choke to ground. This requires a very large blocking condenser because the series resonance of the choke and blocking condenser must occur below the lowest frequency to be amplified. Electrolytic condensers are usually used in consideration of space requirements. A push-pull stage (Fig. 17-21) has the advantage of eliminating this blocking condenser and in addition, as usual, leads to reduced distortion in the amplifier output and permits some simplification in the power supply.

The last step in the design is to assign values to the tube grid resistor and the input coupling condenser. This is complicated by the fact that the input resistance at low frequencies decreases when bass boost is employed, and this decreasing input resistance acts in combination with the coupling condenser to reduce the bass response. For example, if a grid resistor of 100,000 ohms is used with a .01 μfd coupling condenser the bass response will be down 3 db at 16 cps with the bass control set for flat response, but at maximum boost the 3 db point will be at 160 cps. This undesirable effect can be eliminated only by making the bass response extend to the proper frequency at maximum boost; in other words, the combination of grid resistor and coupling condenser should be chosen for the desired bass response under the assumption that the stage input resistance is equal in magnitude to the grid resistor. Since at flat response the gain of the stage is slightly less than unity, motorboating will not occur, but at an intermediate bass-boost setting low-frequency oscillation does sometimes take place. It can be avoided by careful decoupling and control of the bass response of the preceding and succeeding amplifier stages. No such difficulty with oscillation is experienced with the push-pull circuit.

The cathode of the tone-control tube is at a high dc potential above ground. This makes a separate heater supply essential in some cases, but this arrangement is at any rate always desirable because hum is considerably reduced.

Because the push-pull arrangement is least likely to introduce distortion and since it also eliminates the problem of the large condenser required to prevent current flow through the choke, it was selected for the tests. The choke was a *U.T.C.* VIC-17 reactor, which could be adjusted to the desired inductance, and when this was done the frequency response was substantially as expected, Fig. 17-24. The use of 150,000 ohm controls resulted in a smooth control action over most of the rotation of the knob.

When connected to an oscilloscope, the output appeared to be distortionless as long as the input was maintained at less than that permitted by the bias voltage of the control tubes. When the controls are in flat position the circuit will handle at least ten times this permissible input without distortion, because of degenerative action. Distortion due to overloading is, therefore, most apt to occur at low or high frequencies when bass and treble boost are on full.

The apparent fact that the resonant circuit introduces no transient distortion at mid-frequencies (1000 cps) may perhaps be explained by noting that at these frequencies its impedance is higher than the plate or cathode resistance of the tube and the resonant circuit is consequently of little effect there. With the controls in "flat" position, of course, the resonant combination is effectively removed from the circuit and has no effect whatsoever.

The measured gain of the stage is .88. The hum level of the output is very low, less than that of the power supply used for the experimental work. It is nevertheless desirable to keep the signal voltages in such stages at reasonably high levels to override the various sources of noise. With an input of two or three volts peak to the stage, the noise is negligible.

The degenerative tone control serves the quadruple purpose of boosting or

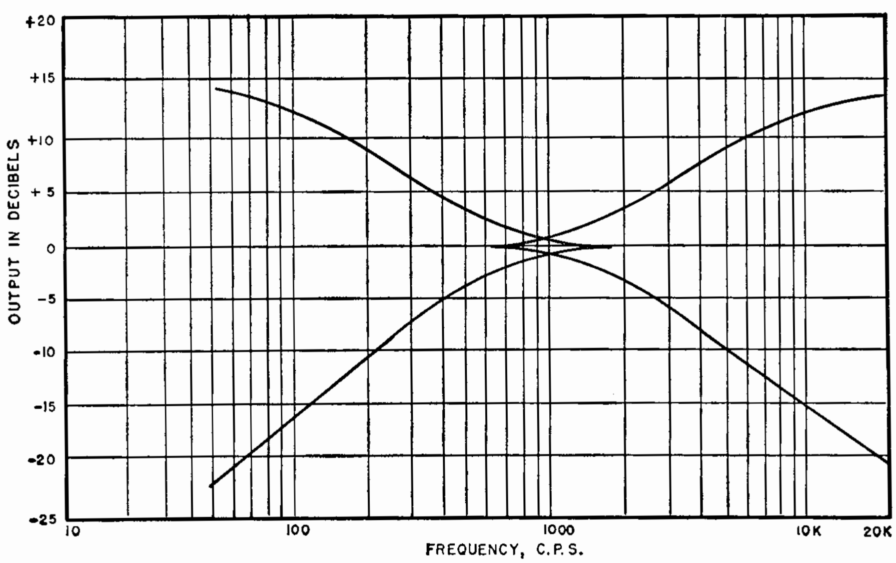


Fig. 17-24. Frequency response curves for the degenerative tone control discussed in text.

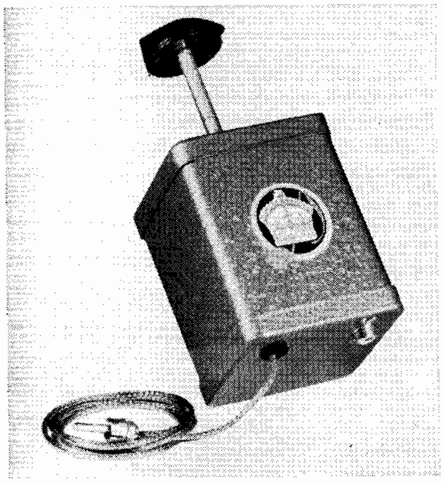


Fig. 17-25. *Goodell* NSF-2 noise suppression filter. (Courtesy Minnesota Electronics Corp.)

cutting bass or treble frequencies with a simplicity unmatched by any other arrangement. In the past the most serious objection raised against its use has been the possibility of introduction of distortion by the iron-core choke and the resonant circuit. This distortion, however, appears to be extremely small and since most other tone-control arrangements use two cascaded stages it seems that at best there is little difference from the distortion standpoint. On the other hand there are two more valid objections to the degenerative tone control; the first is that the high cathode voltage necessitates a separate heater winding and the second is the possibility of noise in the output.

The first objection is hardly a basis for rejection of the circuit since a separate heater transformer is readily supplied. With regard to the second, however, it appears that the tone control shares the same shortcomings as the plate-and-cathode loaded phase inverter, namely, the noise level is higher than that of an ordinary triode stage. For this reason it is desirable that the circuit be incorporated into an amplifier where the peak signal voltages are moderately high—on the order of two or three volts.

The gain of the stage is so near to unity that the tone control is easily incorporated into existing amplifiers without extensive changes in other parts of the circuit. Thorough decoupling of the

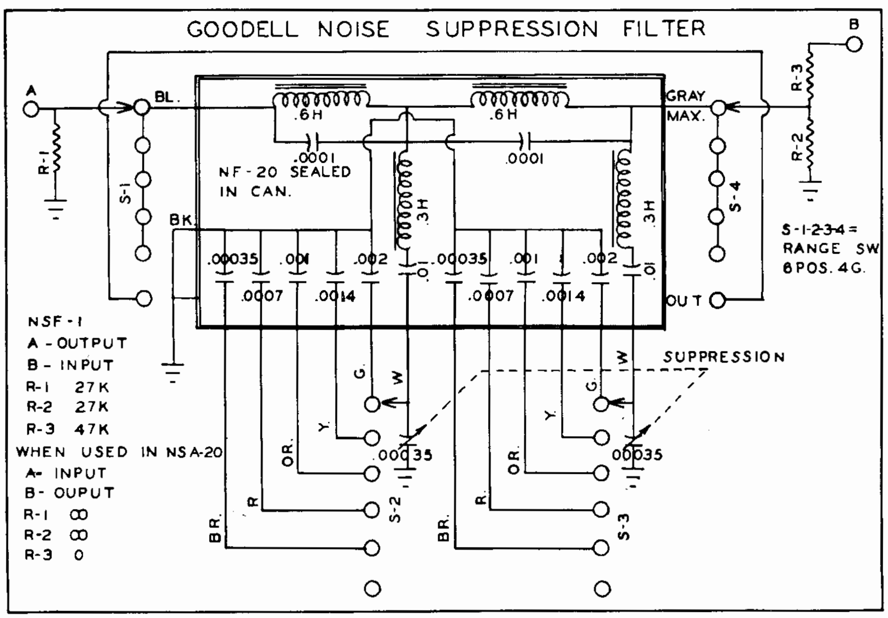


Fig. 17-26. Schematic of the NSF-2 noise suppression filter.

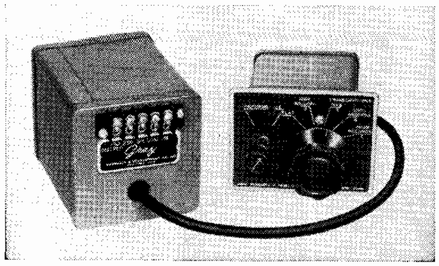


Fig. 17-27. The *Gray* Model 603 equalizer. (Courtesy Gray Research)

"B" supply is about the only stringent requirement.

Finally, not the least of the advantages of this circuit is the ease with which it may be designed for any required bass and treble turnover points and the fact that the controls themselves are commercially available from a number of sources. It appears, therefore, that this circuit is an excellent choice for incorporation into new or existing equipment, including that intended for high-fidelity use.

Noise Suppression Filter

This *Goodell* NSF-2 (Fig. 17-25) unit is very simply installed by plugging the output of the pickup cartridge directly into the small jack on the filter. The output of the filter is a cable fitted with a plug which should be connected to the input of the preamplifier.

When the switch is in the extreme counter clockwise position, the filter is switched entirely out of the circuit and there is no suppression. As the switch is rotated successively through its several positions (Fig. 17-26) clockwise, there will be increased suppression of background noise in the high frequency region. For average shellac commercial records, position three will usually be found most satisfactory and will attenuate very little, if any, of the music signal while effecting a great reduction in surface noise. For good plastic records and the best quality of shellac records, position two will be satisfactory in most installations. Where the background noise is serious or where there is excessive high frequency distortion on the record, it will be desirable to use positions four or five, even though these positions may result in a slight reduction of brilliance at the very high frequency end.

The shielded cable from the pickup cartridge to the input of the filter should be kept as short as is conveniently possible. If the filter is placed in a location such that the length of lead supplied with the unit for connection to the preamplifier input can be appreciably shortened, it is desirable to do so rather than let it hang loosely in order to minimize possibilities of hum pickup from associated wiring. The components in this unit are mounted in a double shield can and there is very little possibility of hum pickup within the filter structures. Consequently the unit may be placed in the most convenient physical location in most installations without regard for electrical and magnetic field considerations. However, it is always best not to run input devices any closer to power transformers, motors, etc., than is necessary for convenient installation.

Gray Record Equalizer

The *Gray* Model No. 603 Equalizer, Fig. 17-27, has been designed to complement recording characteristics of instantaneous lacquers, commercial wide groove records, transcriptions and micro-groove records when played back with *G.E.* RPX-046 Professional series or *Pickering* No. 120 and No. 140 type S or M Cartridges. A choice of 5 conditions of equalization is provided.

The equalizer is housed in two metal cans containing switching and components interconnected by 18″ shielded cable.

Specifications

Circuit—Four terminal network isolated from ground, used with balanced or unbalanced line. One input terminal common to one output terminal, marked "COM," isolated from can and cable shield "SH" terminal.

Compensation Means—Reactive and resistance elements of close tolerances

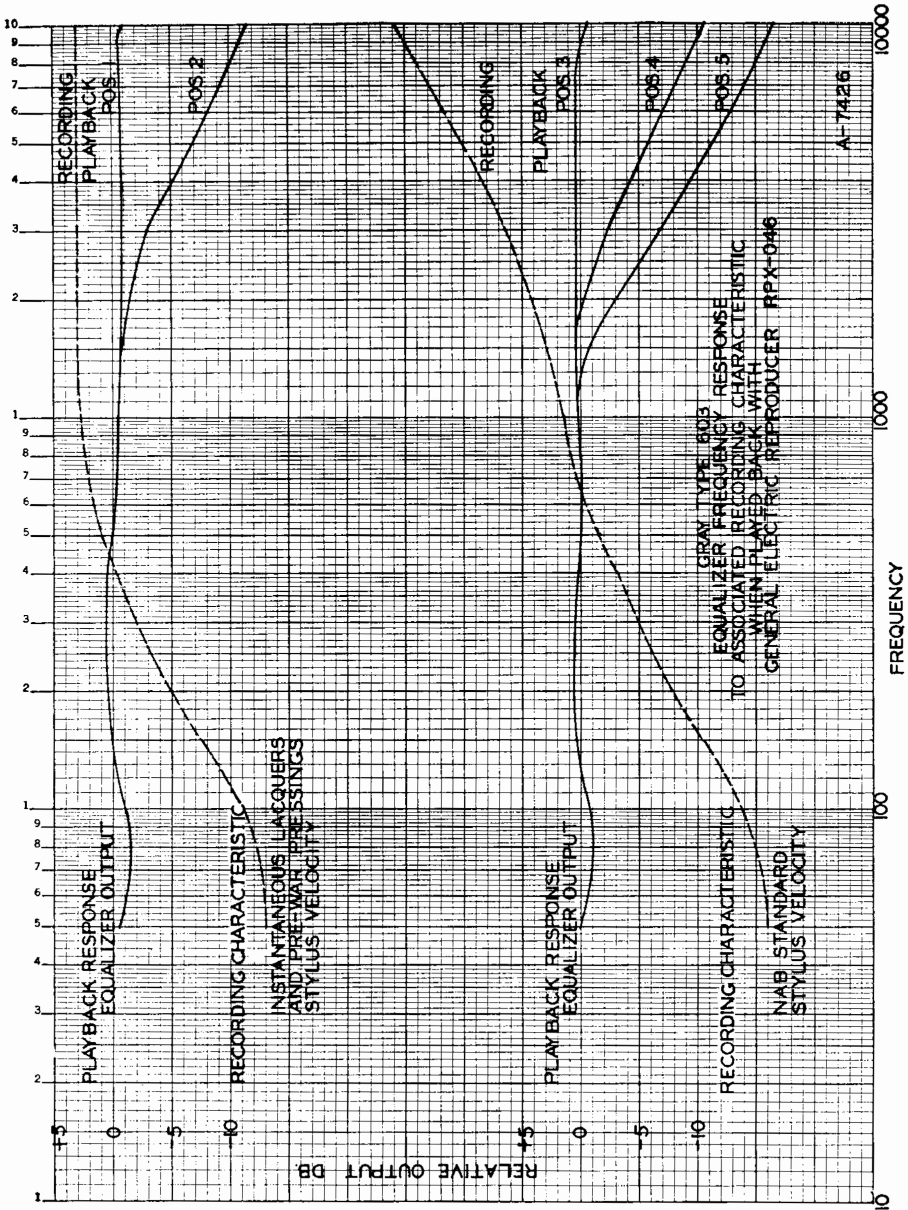


Fig. 17-28. Frequency response curves when used with *G.E.* RPX-046 cartridge.

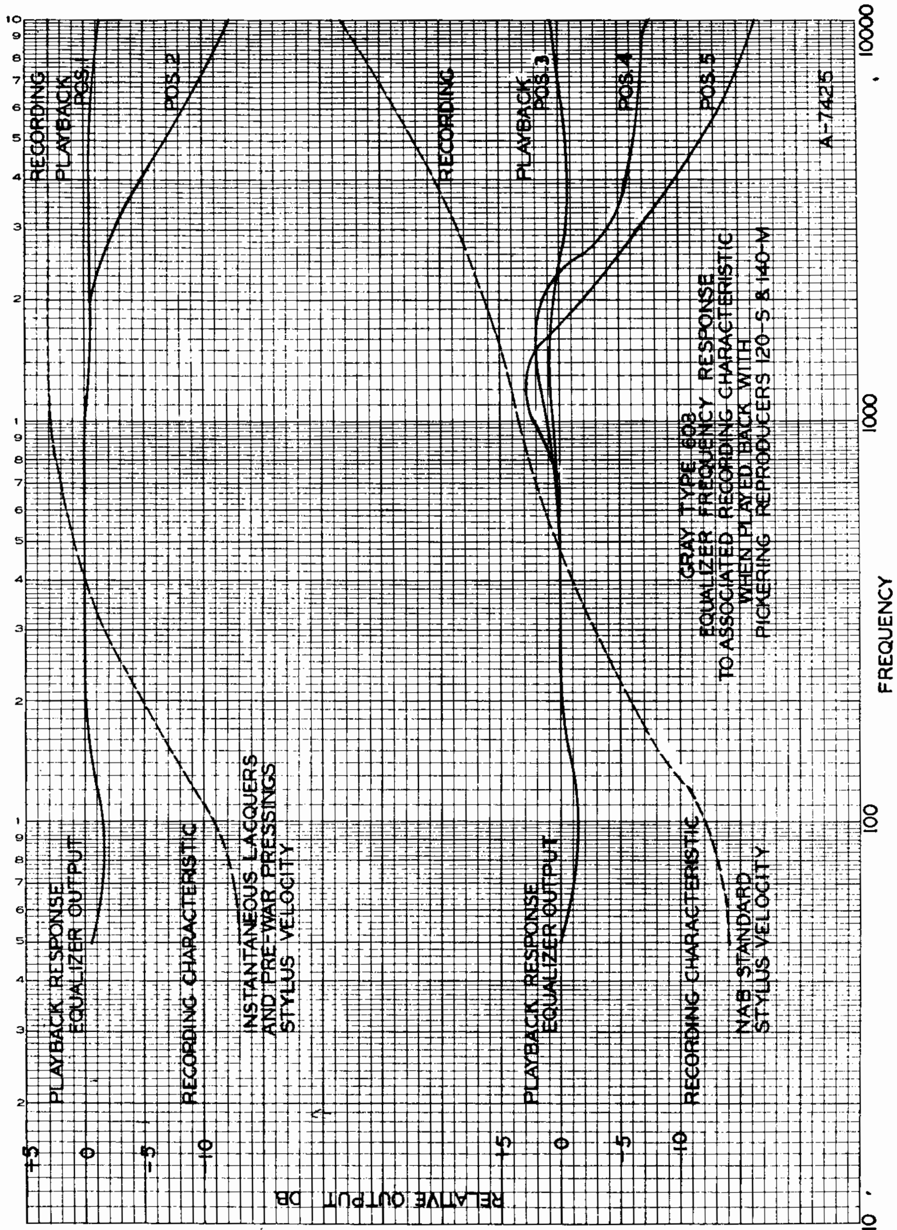


Fig. 17-29. Frequency response when played back with *Pickering* 120-S and 140-M cartridges.

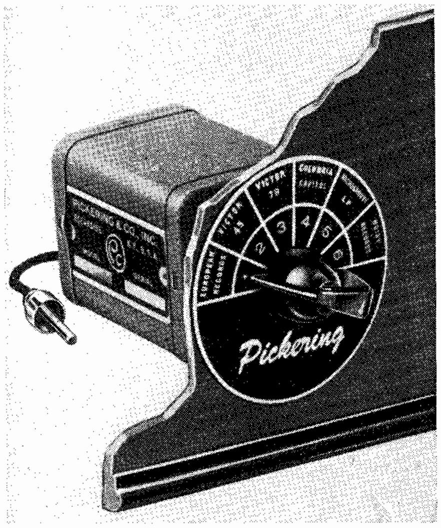


Fig. 17-30. The *Pickering* 132E record compensator. (Courtesy Pickering & Co.)

connected across cartridge. Equalizing networks utilize cartridge impedance as part of compensating networks.

Input Impedance—Function of frequency as well as switch setting due to type of network used. Example: On *G.E.* Position No. 1, about 600 ohms at 50 cps, falls to about 160 ohms at 400 cps, rises to about 2000 ohms at 10 kc. See Fig. 17-28.

Output Impedance—Designed to work into pre-amplifier intended to be operated from 150 or 250 ohm mike source. In generally accepted broadcast practice, such pre-amplifier is essentially unloaded on tube side. Working equalizer into higher than nominal impedance, little effect on frequency response or output level. Somewhat lower impedance reduces general level

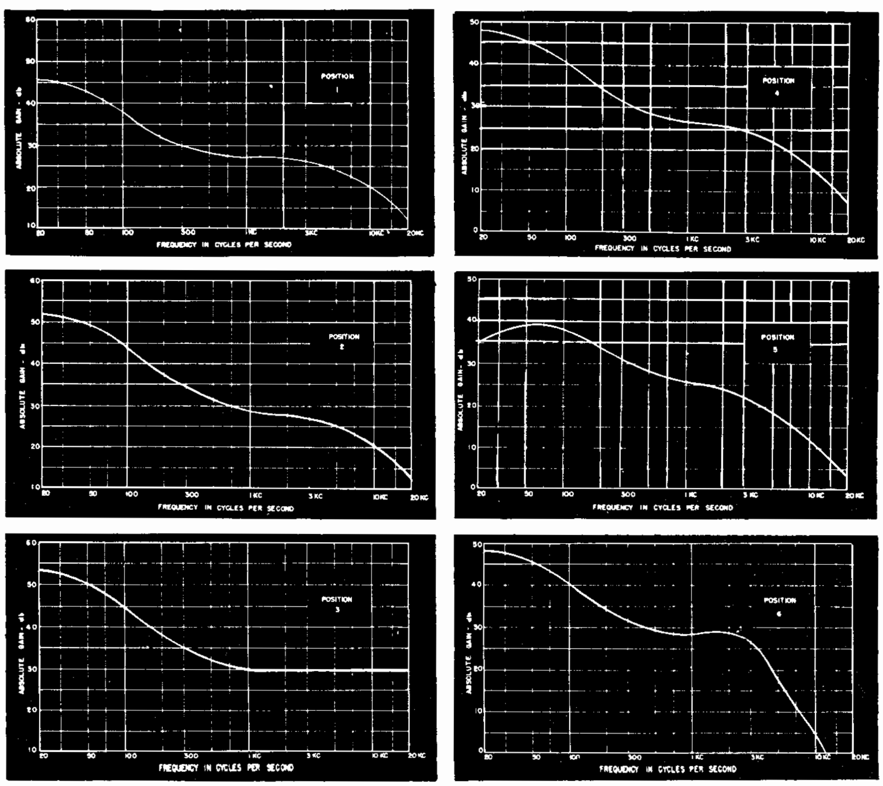


Fig. 17-31. Frequency response curves Model 132E, record compensator with *Pickering* Model 130H preamplifier or equivalent.

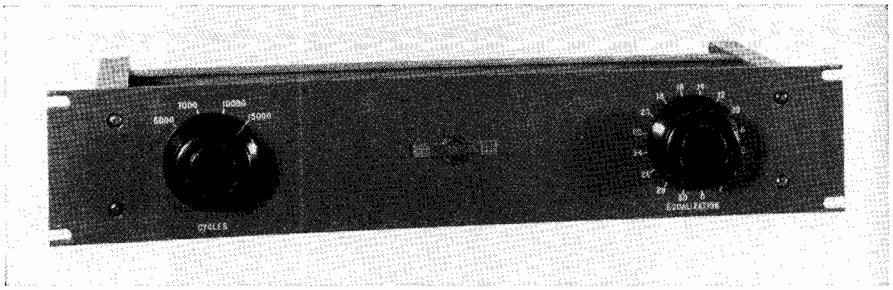


Fig. 17-32. *Collins* 116E-3 program equalizer. (Courtesy Collins)

slightly, several db more at higher frequencies.

Insertion Loss—Defined as function of source, input, output and load impedances. Since first two vary with frequency and input impedance varies with switch setting, a single figure cannot be given. See *Output Level.*

Output Level—With switch on Position No. 1, measured across 250 ohm equalizer output terminals, average *G.E.* RPX-046 cartridge driven at 4.7 cm per sec., level approximately –64 dbm. With *Pickering* 120 or 140 type S or M, level approximately –50 dbm. Columbia 10003-M record 1000 cycle band is approximately this velocity at 78 rpm. Above 1000 cps, setting of compensation switch affects these values.

Cartridge Selection—Two position rotary switch selects proper equalization circuits for cartridge type indicated. Single input termination for one cartridge at a time.

Frequency Control—5 position selector switch for choice of 5 different compensations:

Position No. 1—Flat playback to recordings made without high frequency pre-emphasis, such as instantaneous lacquers or pre-war pressings.

Position No. 2—Moderate roll-off for such records when worn.

Position No. 3—Complements NARTB recording characteristics.

May also be used for additional roll-off for worn lacquers.

Position No. 4 and No. 5—Successively increased roll-off for NARTB recordings to reduce abnormal surface noise.

For curves—See Figs. 17-28 and 17-29.

Cartridge Characteristics — Equalizer design based on *G.E.* cartridge inductance of 250 mh and dc resistance 220 ohms and *Pickering* inductance of 200 mh dc resistance 580 ohms. Variation of ± 20% of inductance and/or resist-

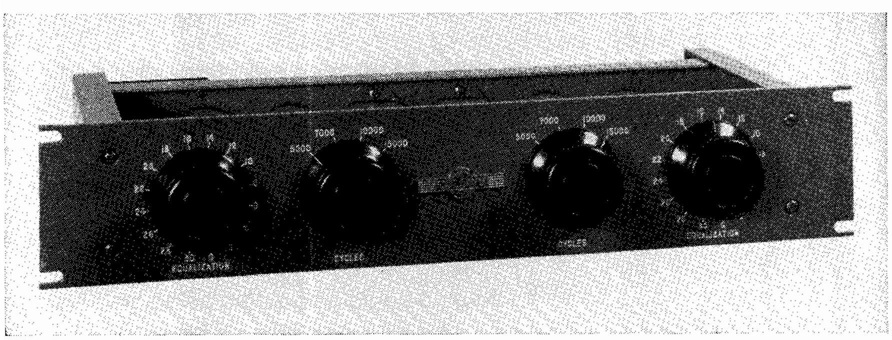


Fig. 17-33. *Collins* 116E-4 program equalizer. (Courtesy Collins)

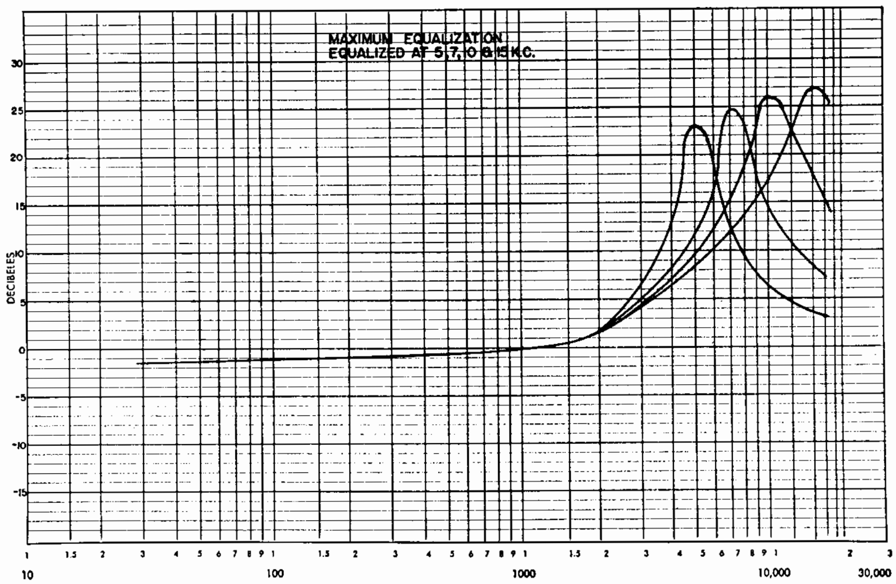


Fig. 17-34. Equalized characteristics of the high-frequency equalizer.

ance from these values has negligible effect. Increase in cartridge inductance of the order of 100% will reduce output above about four hundred cycles by several db. Increase in cartridge resistance of the order of 100% reduces output in region of 100 to 400 cycles several db with negligible effect above 1500 cycles.

Pickering Record Compensator

The *Pickering* 132E Record Compensator, Fig. 17-30, permits proper equalization of the amplifier system to produce optimum reproduction of individual records; because all linear circuit elements are used it has no inherent distortion. Its six positions correctly equalize for all of the established recording characteristics including micro-groove and standard records, domestic and foreign. See Fig. 17-31.

Specifications

Output Level—To feed into high-gain amplifier which has 6 db per octave rise below 500 cycles per second.

Installation—Unit can be mounted in any position (on panels up to ½″ thick) by means of threaded bushing. Switch shaft is 1⅝″ long and can be cut to any desired length. Since no power is required to operate the Record Compensator only a single connection has to be made to a suitable preamplifier. Input connection—standard socket. Matching plug furnished with unit. Maximum distance between record compensator and preamplifier input 20 inches, cable supplied.

Dimensions and Weight—Size of unit: 1⅞ x 2″ by 2½″ overall, less switch shaft. Weight: 6½ oz.

Program Equalizers

The *Collins* 116E-3 (Fig. 17-32) and 116E-4 (Fig. 17-33) equalizers are especially suited for stations having a variety of remote programs coming from different lines. The 116E-3 and -4 offer equalization in the high frequency ranges only. A calibrated attenuator selects the amount of equalization at the required frequency which is selected by a panel switch. Such calibration reduces line equalization time to a single run to find the line characteristics, and adjustment of the equalizer to the conjugate frequency characteristic.

Fig. 17-35. *Collins* 116F-1 program equolizer. (Courtesy Collins)

The 116E-3 is a single high frequency equalizer while the 116E-4 has two identical high frequency equalizers mounted on the same panel with separate input and output terminals.

116E-3 Equalizer Specifications

Input and output impedance—600 ohms unbalanced.
Equalization frequencies—5, 7, 10 and 15 kc.
Maximum boost—Approximately 30 db.
Insertion loss—Approximately equal to amount of equalization used.
Frequency range—30 to 15,000 cps.
Dimensions—19″ w, 3½″ h, 7¼″ d.
Weight—6 pounds, 7 ounces.
Finish—Metallic gray panel; flat gray back.
Collins Part No.—520 3577 00.

116E-4 Equalizer Specifications

Input and output impedance—600 ohms unbalanced.

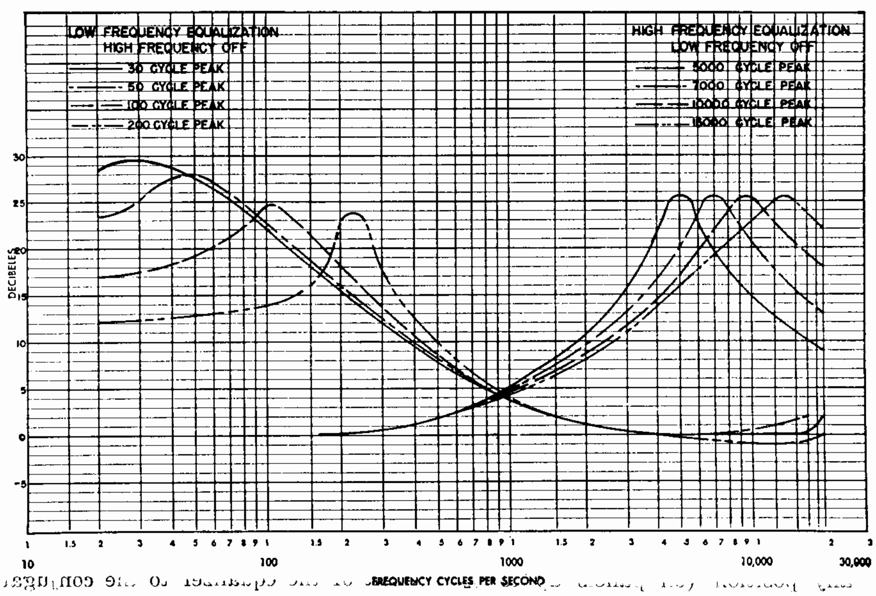


Fig. 17-36. Performance curves of the *Collins* 116F-1 program equalizer.

Equalization frequencies—5, 7, 10 and 15 kc. (Fig. 17-34).
Maximum boost—Approximately 30 db each channel.
Insertion loss—Approximately equal to amount of equalization used.
Frequency range—30 to 15,000 cps.
Dimensions—19″ w, 3½″ h, 8¼″ d.
Weight—9 pounds, 7 ounces.
Finish—Metallic gray.
Collins Part No.—520 3578 00.

Low-High Program Equalizers

The *Collins* 116F-1 Equalizer (Fig. 17-35) provides complete facilities for controlling the frequency response of program and communication circuits. As these units have an insertion loss of approximately 30 db, the *Collins* 6R Isolation Amplifier used in conjunction with the equalizers will provide a means of bringing the level back to normal, plus a little gain if desired.

Specifications

Input and output impedance—600 ohms, unbalanced.
Equalization frequencies—30, 50, 100 or 200 cps at low frequency. 5, 7, 10 or 15 kc at high frequency. (See Fig. 17-36.)
Maximum boost—26 db, in steps of 2 db each. High and low frequency equalization independently adjustable.
Insertion loss—30 db at unequalized frequency.
Frequency range—30-15,000 cps.
Dimensions—19″ w, 5½″ h, 7½″ d.
Weight—15 pounds.
Finish—Metallic gray.
Collins Part No.—520 2893 00.

CHAPTER 18

Attenuators and Mixers

An Analysis of various controls used in audio amplifiers, recording equipment and in broadcast amplifiers.

Input Coupling Methods

● Many amplifiers on the market today, particularly the "run of the mill" variety, employ for their input systems attenuators of simplified design. Several of these are illustrated in Fig. 18-1. Reference to Fig. 18-1A shows the simplest type of coupling to a single grid from a single input source. This is a conventional voltage divider circuit.

In Fig. 18-1B we note a series resistor between the volume control slider and the grid. This affords a bit of isolation, as we shall see becomes necessary as in E.

Fig. 18-1C illustrates a simple mixing circuit for two inputs when it is desired to feed into a single grid. The potentiometer winding has a center tap which is grounded.

Fig. 18-1D shows two input sources, connected in parallel, across two potentiometers, feeding one grid. This circuit is not to be recommended as one control can short out the other input.

Fig. 18-1E is a dual version of that shown in B, isolating resistors are required in order to prevent the shorting action as would occur in D.

Fig. 18-1F illustrates a preferred technique whereby two input sources are fed to two grids of two different tubes, the plates of which are connected in parallel.

Fig. 18-1G is the same only here we find that a dual tube, such as a 6SN7 is employed using independent grids and the two plates are connected in parallel.

The above are simplified forms of mixer controls, found in conventional equipment. When designing circuits, attenuators and mixers for broadcast applications special types of controls or attenuators are required in order to maintain a constant impedance load throughout the range of the control.

Attenuators for Recording and Broadcast

Studio installations of speech input equipment almost invariably use ladders (Fig. 18-2C) or bridged T. (Fig. 18-3B) attenuators as the mixing and volume level controls in their equipment. A potentiometer 18-2A is sometimes used where the cost factor must be considered. Usually this is used as a master gain control, but only where the entire speech equipment is built as a unit incorporating amplifiers and mixing equipment in one housing. This is done to avoid long high impedance leads and the attendant danger of hum pickup. It should be noted that changing the characteristic impedance of a sound channel is often necessary, particularly in test installations where a

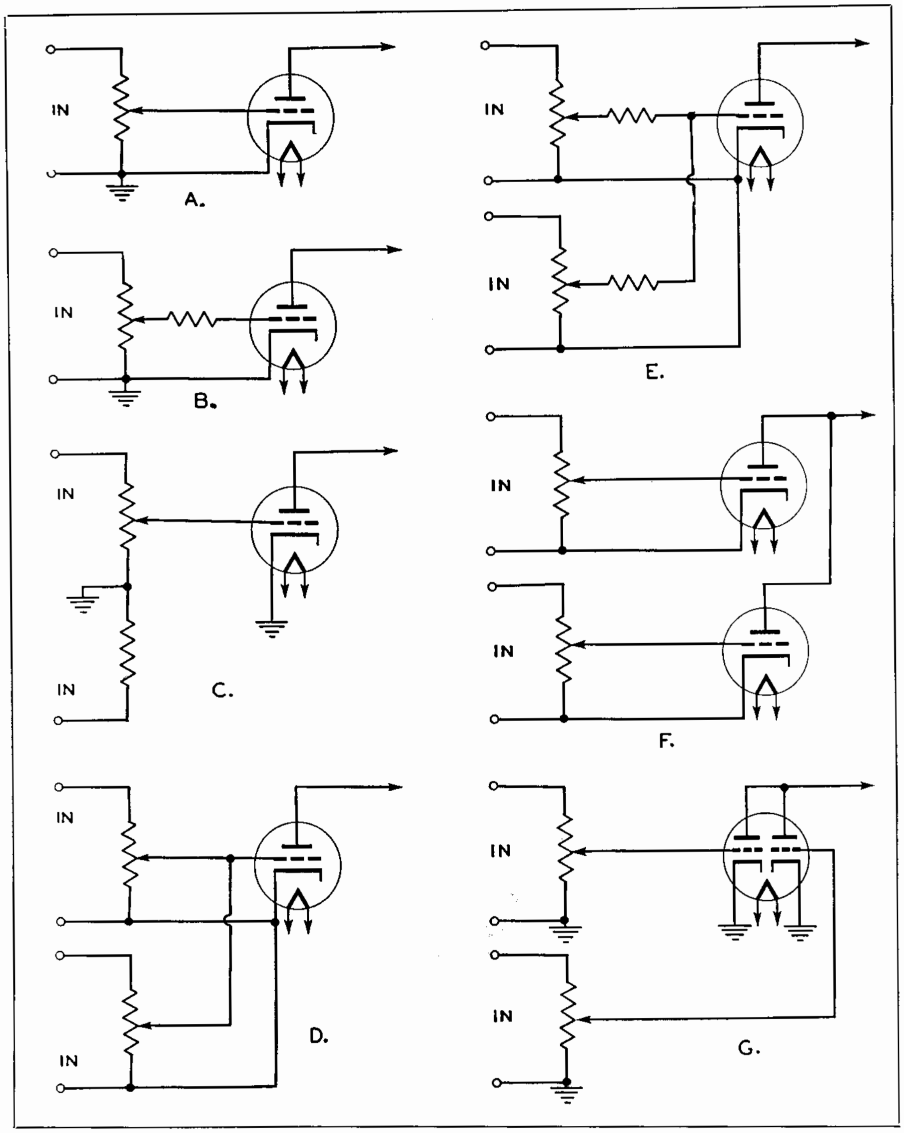


Fig. 18-1. High impedance circuits for general uses. Representative controls of the potentiometer variety.

special type of attenuator is available for the purpose. This offers either the minimum loss for the ratio of impedances matched, or a constant loss for any ratio selected.

Public Address Attenuators

Installations of any form of public address equipment (PA) for sound reinforcement, present the problem of control of level from individual or groups of loudspeakers where a single amplifier feeds more than one transducer.

The power attenuator can be used to control the sound level delivered without altering the volume from the remaining speakers connected to the same power amplifier source. As a control must dissipate power, its size and se-

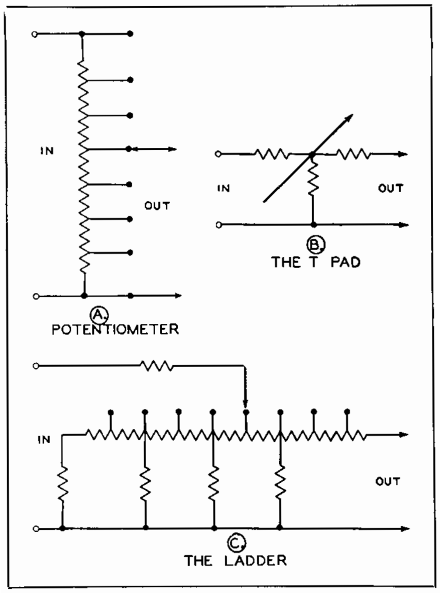


Fig. 18-2.

lection depends upon the level at which the loudspeaker is designed to work. Frequently these devices for group control assume large physical dimensions, and in such cases it is well to observe that the power rating of such devices is based upon sine waveform for speech and music. This will allow a slight overloading of the control with safety. However, not more than 25% overload can be tolerated unless it is for very short periods of time, otherwise the life of the attenuator will be endangered.

Attenuators can generally be used to achieve tone compensation where they are merely elements of a complex correcting circuit as in the familiar equalizing units offered in the trade and discussed in other chapters. A tone-compensated ladder is shown in Fig. 18-3A. In the recording process, particularly at 33⅓ rpm used in making transcriptions, there must be considerable equalization as the recording radius is reduced. Attenuators lend themselves nicely to this function and can be ac-

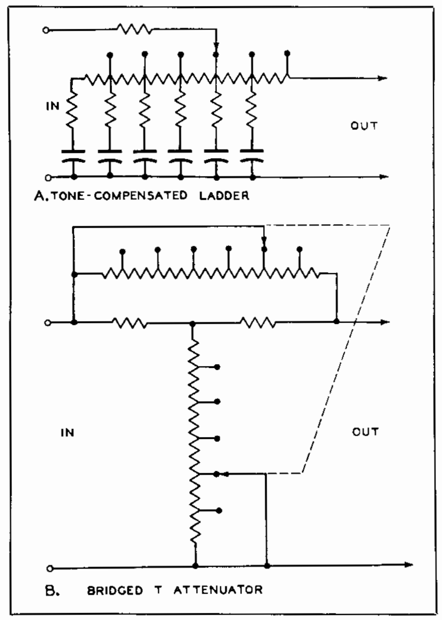


Fig. 18-3A, B.

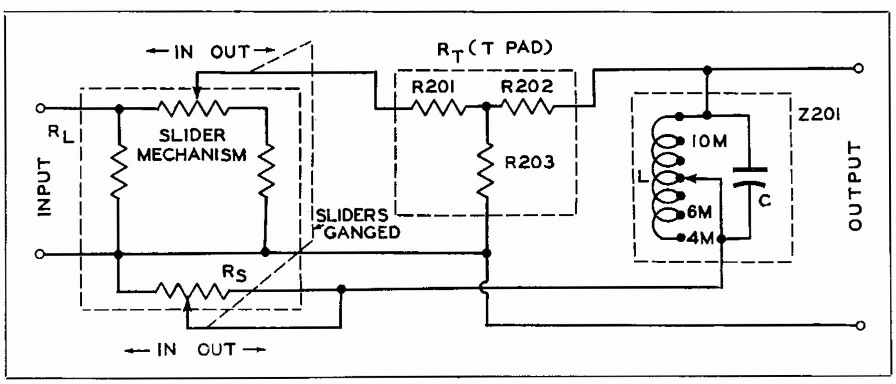


Fig. 18-3C. Simplified diagram of the Presto automatic equalizer.

tuated by the recording lead screw mechanism or manipulated separately at the discretion of the recording engineer. When attached to the recorder, they take the form of sliding units or those with an arc movement which are operated through a mechanical linkage to the traveling recording head. See Fig. 18-3C.

Another use for attenuators is the application in circuits measuring volume level, as discussed in the chapter under "Decibels." The familiar decibel meter which is a rectifier type of volt meter has long been used in all sound installations and is still widely used in laboratory and recording studios. Where it is necessary to terminate the measured circuit in its characteristic impedance, the meter can be connected to a T pad across the line, thus serving to adjust the range of the meter. The accuracy of measurement is determined by the accuracy of calibration of the resistance built into these meter multipliers. Representative values for such resistances are included in the appendix.

Several years ago the broadcasters cooperated in developing a standard measuring instrument "the VU meter" which today is universally used in that industry. These meters possess specific desirable ballistic characteristics and they read an audio level of plus 4 dbm which is the reference level at 1 mw at 600 ohms impedance (Z). This is equivalent to the zero reading on the VU meter scale. This type of resistance or multiplier circuit takes the form of a pure T pad so that the characteristics of the meter are not changed nor any inordinate loss imposed upon the audio line being measured. A small attenuator in the form of a rheostat is incorporated in these circuits wherever it is necessary to standardize many instruments throughout a broadcasting network to read alike. See Chapter 8.

Fixed Attenuators and Pads

There is a vast field of applications for fixed attenuators or fixed pads. In mixing circuits several incoming signals are frequently not of identical level, and any great discrepancy can be compensated for by a "fixed pad" of suitable circuit configuration to match the general schematic of the mixer. This may be balanced or unbalanced to ground. When two amplifiers are coupled together at low impedance, good engineering practice dictates that at least 6 decibels of loss in the form of a fixed attenuator should be included in the circuit between the two connected transformer windings. This will eliminate circulating audio currents. Multiple end use of a single source of sound energy (feeding both an AM and FM transmitter, or driving many power amplifiers in the sound system) calls for the use of a dividing network. Fixed "loss pads" for 500 Ω lines (10 db) are shown in Fig. 18-4.

This is a form of fixed attenuator offering a small loss between the source and its respective loads, but simultaneously presenting considerable separation between the several loads. Inverted, the same pad becomes a combining network for assembling several sources into one load. Usually these pads are designed to present a uniform impedance to all terminals and assume several complicated circuit configurations.

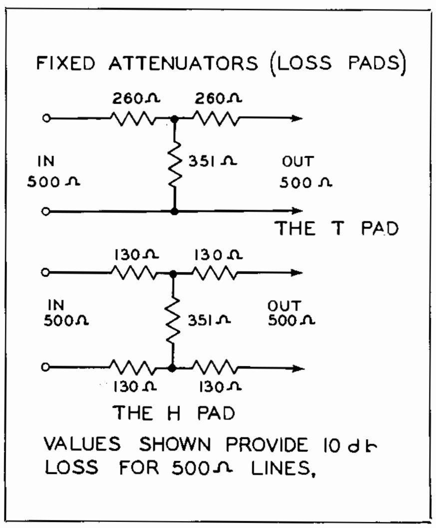


Fig. 18-4.

However, they are all examples of attenuators. When connecting a bridging amplifier across a properly terminated junction of source and load, a bridging pad can be used, if the second amplifier does not possess a bridging transformer.

Classifications

Modern attenuators can be grouped into four main classes. Of these, two classes are variations of each other. These are the ladder, T, bridged T and potentiometer. Figs. 18-2 and 18-3B show the electrical differences of these units. The ladder, Fig. 18-2C has but one row of contacts and a moving arm. It is easy to operate and possesses smooth motion. Furthermore it is electrically quiet and relatively low in cost.

This device offers an insertion loss of about 6 db and this means cutting in half the gain of the source signal and additional amplification is therefore required. If the impedance ratio of the ladder can be made 1:2, then the insertion loss becomes 2 db. This change in impedance ratio is also an undesirable factor. A special case of zero insertion loss ladder has been made, but it is costly and mechanically complex therefore it is not met with wide commercial acceptance.

The T Pad

The T pad of Fig. 18-2B requires two or three sets of contacts. The ladder is for the pure T, and no collector rings are used. While it is possibly a bit harder to rotate, the advantages of a zero insertion loss and smoother impedance characteristics favor the selection of this form of control. The bridged T, offers smoother over-all impedance characteristics when the brush shunts a pair of contacts and requires fewer resistors in its construction.

Potentiometers

A potentiometer, as its name implies, is a voltage measuring device and is usually found in high impedance installations, such as are illustrated in Fig. 18-1. Similar to the ladder in physical appearance Fig. 18-2A, these units can be made economically. The load side of the potentiometer offers a varying impedance and unless this can be accepted, its compensated form must be used. However, any type of compensation presents serious frequency discrimination, even at audio frequencies, due to the series resistors and the attendant "Miller effect" in the grid circuit to which it is connected, see Fig. 18-5.

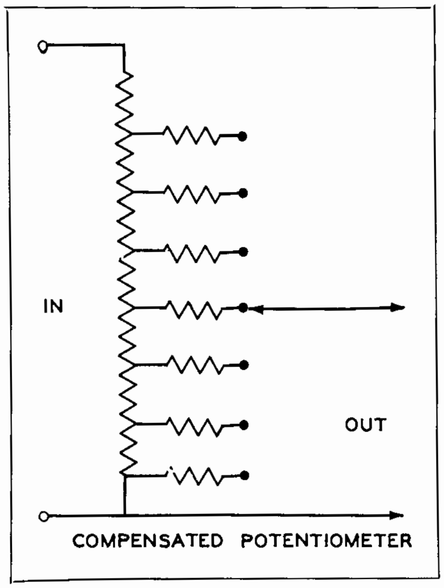


Fig. 18-5.

Mixer Circuits

Mixer circuits, at least up to a few years back, for broadcast were usually of the "low level" variety. Mixing took place right after the microphone and before any amplification had been inserted into the system. This was practical then, as most microphones were not of the highest fidelity. They possessed higher output than they do now and the noises introduced by the mixer were not so noticeable. With the acceptance and universal use of the wide range, high fidelity, low output microphone and transcription recorders, some form of preamplification became essential. Mixer circuits then were moved to a higher level position in the speech input installation and this is the generally accepted form used today.

Ladder Mixers

The ladder is almost universally used for low level mixing circuits because it tends to be quieter. An attenuator's noise along with the signal receives the effect of the full amplification of the system so consideration must be given to noise characteristics. The ladder is also used in "high level" mixing, where cost of components is a consideration and where ease of rotation is desirable. If the mixing installation is a large one, and the inclusion of ladder attenuators will tend to increase the loss of the mixer then bridged T pads are to be recommended. They are the better of the two for this service and frequently are used as master gain controls. Noise voltages introduced by the controls are smaller in high level mixing and with a suitable form of internal wiring, the noise can be made to decrease with the signal, as added loss is inserted in the circuit.

Linear Control

A linear control is just what its name implies. The loss it introduces is a linear function. It increases an identical amount for each step of rotation, thus the amount of loss inserted is uniform for each increment of motion of the knob and the total loss is as stated on the name plate of the unit by the manufacturer. Smooth opening and closing of a program is impossible if the control used has a loss of 40 decibels and the circuit has a sound level of about 60 decibels above the threshold of hearing. The first step will be up 20 decibels from silence, which is an undesirable condition. The tapered control on the other hand offers a uniformed amount of attenuation per step for approximately 70-75 percent of its travel and then increasing amounts for each extra step of attenuation.

Electrical smoothness of operation of the control circuit is thus afforded without any sudden jump from no signal to one of appreciable level. While tapered controls of only 20 steps are quite common, smoother control, especially on tones of long sustained duration is afforded by controls of 30 or more steps.

Master Controls

A master gain control has usually been selected from among the linear controls offered. In all recording installations where a most careful control of the sound level must be maintained to prevent overcutting into adjacent grooves, a master gain of 0.5 db per step is a prerequisite. For monitoring amplifiers and those used in PA work, a control as coarse as 3 db per step is allowable. For audiometric and laboratory work, the decade attenuator is selected with two bridged T pads connected in tandem to afford control of 1 db per step for a total of 110 db.

Mixers for Recording and Broadcast

Every source of signal must eventually reach a mixer if it is to be combined into an outgoing signal. It is not necessary to provide more mixing positions than is dictated by the number of signal sources generally used. Selector key switches, patch panels or some other means of switching is now incorporated into the speech input system, to facilitate selection of those programs which occur only occasionally if an inordinate number of programs sources must be handled. For a small broadcasting station a six channel mixer is adequate. About the most that could be handled would be about 7 channels. In such layouts they will handle two studios, two transcription machines, an announcer's booth (without calling upon the control operator to do any switching) and other inputs. In such small installations, four premixing amplifiers are generally used.

A small studio can get along well with a four channel mixer especially if its use is confined to live programs or to a disc jockey show where the turntables can be wired into the mixer instead of requiring additional microphones. A single microphone will suffice for this type of program, leaving the remaining position for the use of the announcer's mike or a third turntable, reserved for transcribed spot announcement. Large studios are frequently equipped with as many as a dozen microphone channels. In such cases provision for two or three sub-master gain controls is also made. Wiring these into the master gain control is becoming standard practice and gives the mixing engineer added facility in the control of the open microphones.

Remote equipment of the portable type has been standardized at two and four channels. Nevertheless a single channel remote amplifier is useful in small stations for special events programs, especially when they cannot be rehearsed in advance or for emergency pickups. Here a single microphone is used by the announcer who frequently manipulates the gain as well.

Mixers for Recording Studios

Two general categories and their mixer system requirements divide similarly. Those studios making "off the air" checks of a radio program, proof of performance and reference can get along with only four channels. Frequently only two are used. However, when making master records the most flexible arrangement is necessary and there must be facilities comparable to those of large broadcasting studios. Exercising the finest possible control of sound is essential.

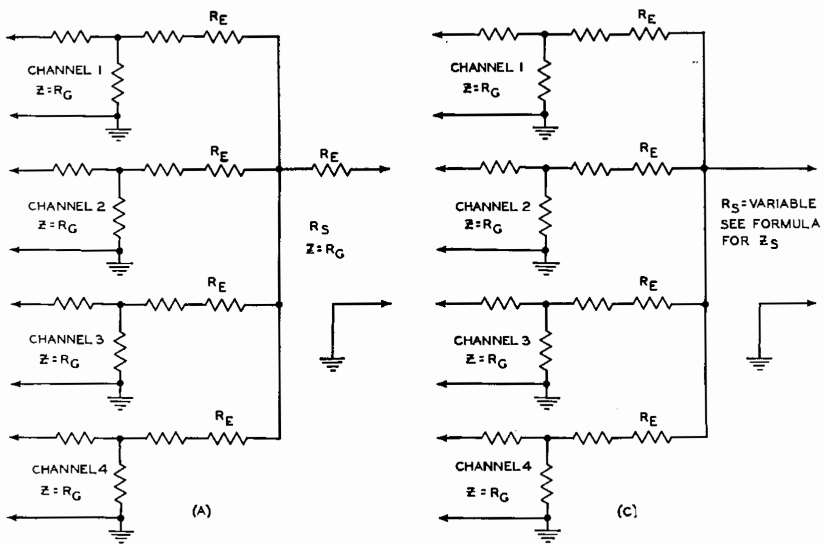
The loss per step of attenuation is usually never more than one decibel per mixer, and one-half decibel per step on the master gain control. There are two essential locations in the speech input system where a mixer can be employed. One is the installation in which the mixer immediately follows the signal sources, which is known as low level mixing and the other where some form of pre-amplification is used ahead of the attenuators, or known as high level mixing. No matter where the mixer is located, the insertion loss it presents to the over-all system remains the same. Modern attenuators are so well constructed that their noise level, due to thermal agitation, is well below that of the microphone and its associated premixing amplifier. In high level mixing, the noise of this combination also decreases with that of the signal as more attenuation is inserted into the system through manipulation of the attenuators. The dangers of hum pickup by the wiring and other random noise sources are added reasons which dictate that low level mixing should not be continued.

Constant Impedance Mixing

Here the impedance of the mixing system remains constant both with respect to the source and load between which it is connected and to the setting of the controls themselves. One exception will be discussed; the parallel type mixing circuit of reduced insertion loss. In this the output impedance is a function of the number of channels but is not adversely affected by the operation of the controls. This condition of constant impedance is an easy one to maintain in today's designs because the sources of signals are the output of pre-mixing amplifiers. All of these can be engineered for a standard impedance of 600 ohms.

Parallel Mixers

These are of two general types, those with constant input and output impedance and those where the output impedance changes when the number of channel changes, but whose insertion loss is less than the former type for the same facilities. In selecting a definite type of mixer it is well to bear in mind that possibly more channels may be added later and consequently one whose impedance remains constant is preferable. Fig. 18-6A shows the typi-



INSERTION LOSS db. $= 20 \log_{10} N$

BUILD-OUT RESISTORS $R_E = \dfrac{R_G\,(N-1)}{(N+1)}$

EXAMPLES

NO. OF CHANNELS N	INSERTION LOSS, db.	BUILD-OUT RESISTOR R_E
2	6.0	$R_G/3$
3	9.5	$R_G/2$
4	12.0	$3R_G/5$
5	14.0	$2R_G/3$
6	15.6	$5R_G/7$

(B)

OUTPUT Z (OHMS) $= R_G(2N-1)/N^2$

INSERTION LOSS db. $= 10 \log_{10}(2N-1)$

BUILD-OUT RESISTORS $R_E = R_G\,(N-1)/N$

EXAMPLES

NO. OF CHANNELS N	INSERTION LOSS, db.	BUILD-OUT RESISTOR R_E	OUTPUT IMPEDANCE
2	4.8	$R_G/2$	$3R_G/4$
3	7.0	$2R_G/3$	$5R_G/9$
4	8.4	$3R_G/4$	$7R_G/16$
5	9.5	$4R_G/5$	$9R_G/25$
6	10.4	$5R_G/6$	$11R_G/36$

(D)

Fig. 18-6. (A) Four channel constant Z mixer. (B) Formulas for build-out resistor and insertion loss, with sample calculations. (C) Four channel mixer with varying output Z. (D) Formulas and sample calculations. Figures for insertion loss cover the mixer when assembled with pads of 0 minimum loss. Use of ladders will increase this loss by 6.0 db. for the mixer, and an added amount if the ladder is used for master gain control.

cal four channel parallel mixer of the constant impedance type, and Fig. 18-6B gives both the formulae and typical values. Formulae for the necessary "build-out" resistors are included where the circuit parameters require them. Fig. 18-6C shows the schematic for the four channel variable impedance mixer and Fig. 18-6D covers the formulae and typical examples of the calculations.

To match this system to a standard master amplifier or even a master gain control requires a repeat coil with a number of input impedances available, and of such design that its transmission characteristics are equal to or above the over-all characteristics of the system. It is well to isolate the secondary of the repeat coil from the master amplifier input transformer primary with some sort of a pad. A potentiometer gain control can be used in the master amplifier in this type of system and this is frequently found in moderate priced ready-built studio equipment offered to the broadcasters.

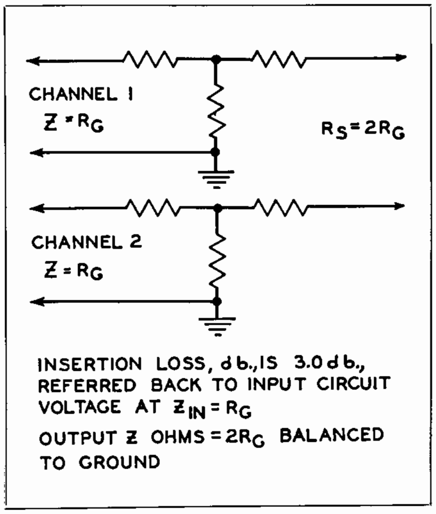


Fig. 18-7. The Two Channel series mixer.

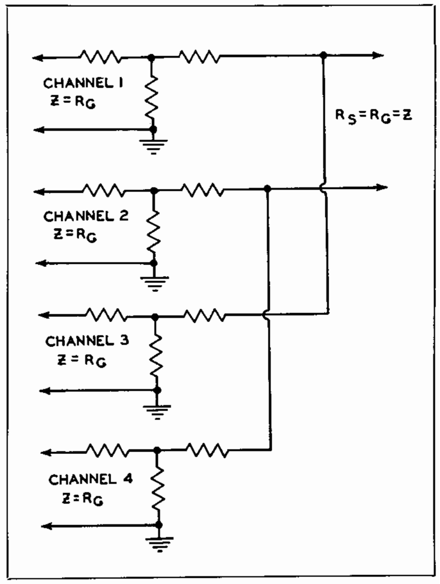


Fig. 18-8. Four-channel parallel-series mixer whose loss is 6 db and whose impedance ratio is unity.

Series Mixers

These mixers as a rule are not desirable because they do not allow grounding of every input channel if more than two such channels are used. However, the two channel mixer is accepted and is used in transmitter control desks between two turntables and for small studios. See Fig. 18-7. Parallel-series and series-parallel mixers are to be avoided because of their varying impedance characteristics and the inability to properly ground several channels. The number of channels effects the loss and impedance, and there is a strong tendency for the settings of the individual controls to be reflected in the other channels.

One exception to this type of circuit is the special design of four channel mixer, Fig. 18-8, which allows for perfect grounding of all channels and has an insertion loss of only 6 db provided T pads are used. One point to observe is that the output is balanced to ground, which would require using a balanced attenuator or repeat coil if a single ended master control is on hand. If the program amplifier has its own gain control shunting the input transformer the four channel mixer will work into its characteristic impedance and obviate an expensive dual attenuator. Special study of this circuit is warranted as it will be the one most frequently used by the audio facilities engineer of small and large broadcasting stations and recording studios.

There is no substitute for quality when selecting an attenuator for use in high quality sound installations. This applies in particular to broadcasting requirements and to professional recording studios. Even in low priced recorders extreme care should be taken to select equipment having low noise level when turning the various volume controls (attenuators) of the circuits. Many commercial amplifiers today are lax in their selection of proper gain controls and some are extremely noisy and result in great distortion through the circuits which they control.

Reference: Keim, L. B., "Attenuators." Radio-Electronic Engineering edition of Radio News, April, May, 1948.

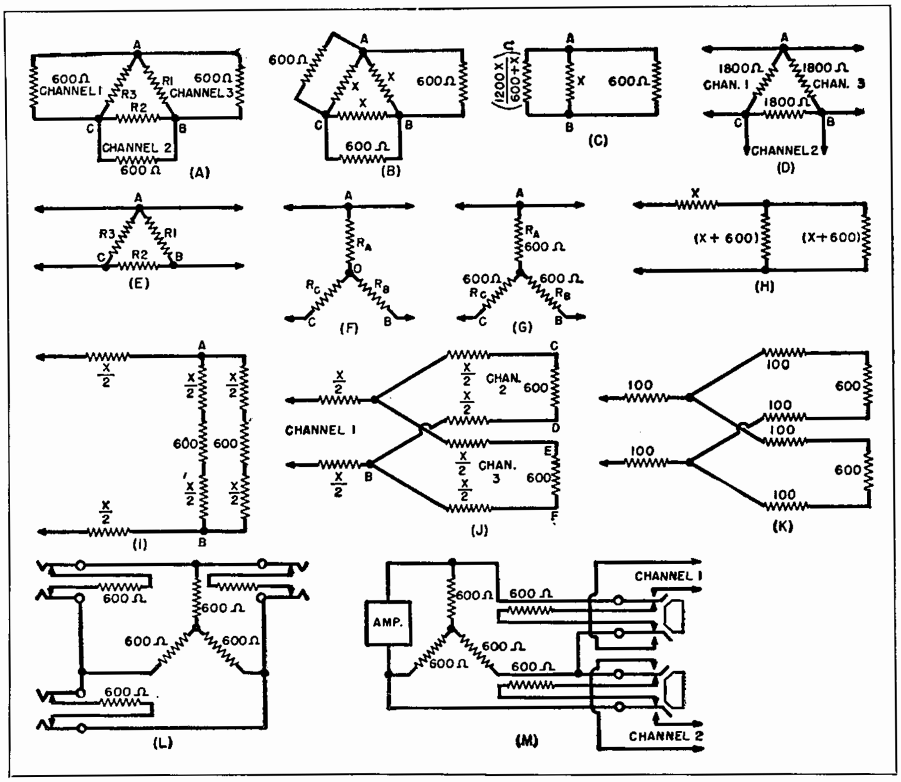


Fig. 18-9. Various forms of pads designed for impedance matching.

Splitting Pads

It is sometimes necessary in audio, communications and television work to apply two alternating currents such as speech, carrier, or music currents to one channel. This may be done by means of a transformer, but owing to the electromagnetic fields and frequency losses from the transformer, this method is not always desirable, and another method using combinations of non-inductive resistances is sometimes used. This method involves the formation of resistance networks, called pads, from non-inductive resistances. In the case of 600 ohm circuits it is usual to make matching pads from 1200 to 600 ohms and to connect the 1200 ohm outputs in parallel. There is a third method, shown in Fig. 18-9A, using a network of standard non-inductive resistors which involves a small power loss but is less expensive than the other methods.[1]

An application of this method is shown in Fig. 18-9A, where it is used to connect two sources of alternating current to one channel or *vice versa.* For purposes of matching, it is necessary to make the values R_1, R_2, R_3, shown in Fig. 18-9A, such that when channels 1 and 2 are terminated in their correct impedances (600 ohms), then channel 3 will also be 600 ohms. Fig. 18-9A may be simplified to the arrangement shown in Fig. 18-9B. Now, when two resistances, R_1 and R_2, are connected in parallel, their joint resistance may be found by the aid of the formula:

$$R_T = \frac{R_1 R_2}{R_1 + R_2} \quad . \quad . \quad . \quad . \quad (1)$$

[1]Carmichael, H. C., "Splitting Pads," Radio & Television News, Vol. 45, No. 4.

If this formula is applied to Fig. 18-9B, then the joint resistance of the two paths (600 ohms and x ohms) from A to C is:

$$\frac{600\,x}{600+x}$$

Similarly, the joint resistance of the two resistances (600 ohms and x ohms) between B and C is:

$$\frac{600\,x}{600+x}$$

These two joint resistance values are in series with respect to the line and thus their total resistance is:

$$\left(\frac{600\,x}{600+x}\right)+\left(\frac{600\,x}{600+x}\right)=\frac{1200\,x}{600+x}$$

The other resistor from A to B (x ohms) is in parallel with this combination as shown in Fig. 18-9C which is a further simplification of Fig. 18-9A. Thus, if formula (1) is applied to this circuit, then the joint resistance of the combination will be:

$$\frac{\left(\frac{1200\,x}{600+x}\right)x}{\left(\frac{1200\,x}{600+x}\right)+x}$$

This, of course, must equal the resistance of channel 3 and thus:

$$\frac{\left(\frac{1200\,x}{600+x}\right)x}{\left(\frac{1200\,x}{600+x}\right)+x}=600$$

This may be simplified to show that $x = 1800$ ohms.

Fig. 18-9D shows the resistance network to satisfy the conditions shown in Fig. 18-9A.

Power Loss of Pad

The power loss from each channel to the mixer channel resulting from the use of this pad can be calculated as follows. Assuming a voltage of 10 volts across a source of 600 ohms in channel 3, then the voltage across points A and B in Fig. 18-9D would be 10 volts and the voltage across points A and C would be 5 volts, as the resistance of the combined resistor AC is equal to the combined resistor CB. The power loss equals:

$$20\log\frac{E_1}{E_2}$$

and if the information supplied in the data is substituted in this formula, then:—

$$\text{Power loss}=20\log\frac{10}{5}=20\log 2=6.02\text{ decibel}$$

Delta to Star Conversion

Another connection called the "star" connection has the advantage over the "delta" connection of being a simpler combination, and the delta circuit shown in Figs. 18-9D and 18-9E may be converted to the star or "Y" connection shown in Fig. 18-9F. Now, if these circuits are to be equivalent, then the resistance between points AB, BC, and CA in Figs. 18-9E and 18-9F must be similar. At a glance it may be clear that in Fig. 18-9F the resistance from A to O will equal 600 ohms, the resistance O to B will also equal 600 ohms, making A to B equal to 1200 ohms as required. It may be mathematically proved, however, that:

$$R_A=\frac{R_1\;R_3}{R_1+R_2+R_3}$$

$$R_B=\frac{R_1\;R_2}{R_1+R_2+R_3}$$

and

$$R_C=\frac{R_2\;R_5}{R_1+R_2+R_3}$$

For example, to convert the delta formation in Fig. 18-9D to an equivalent star or "Y" formation, then:—

$$R_A=\frac{1800\times 1800}{1800+1800+1800}$$
$$=600\text{ ohms}$$

and likewise, R_B and R_C will be equal to 600 ohms respectively. The equivalent star or "Y" connection would be then as shown in Fig. 18-9G.

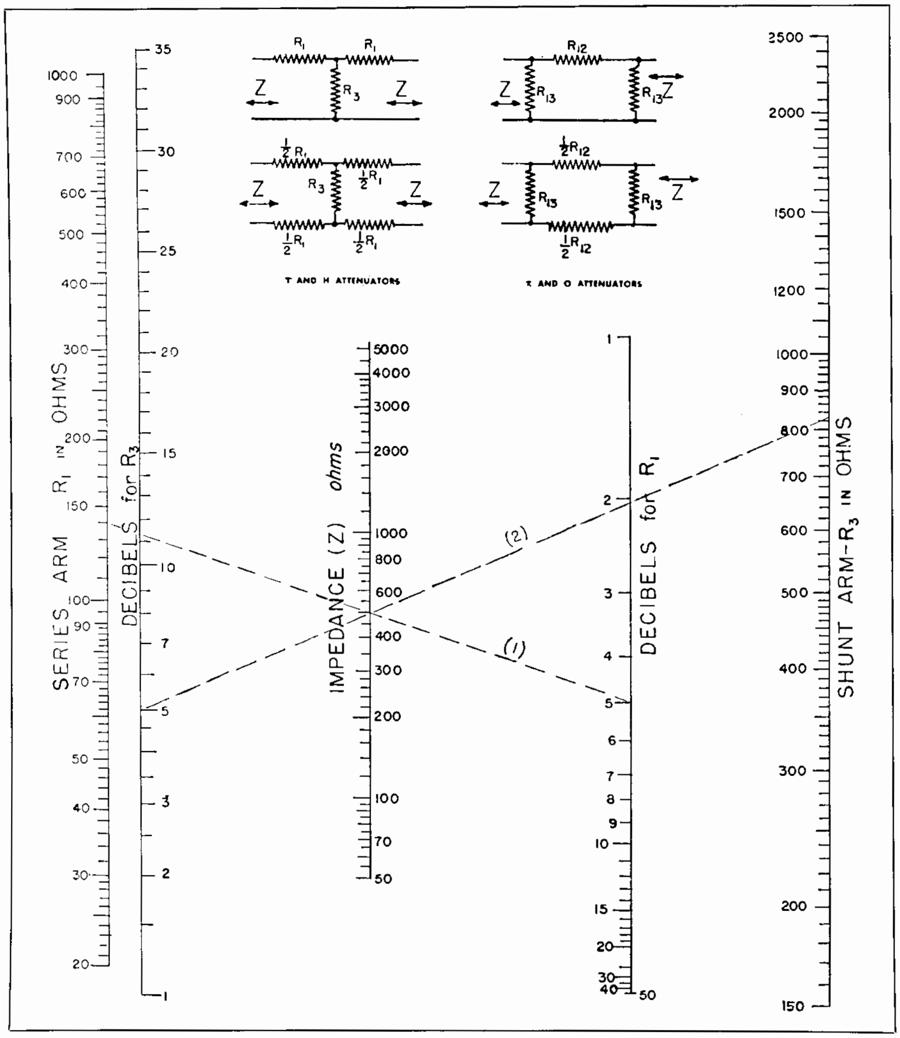


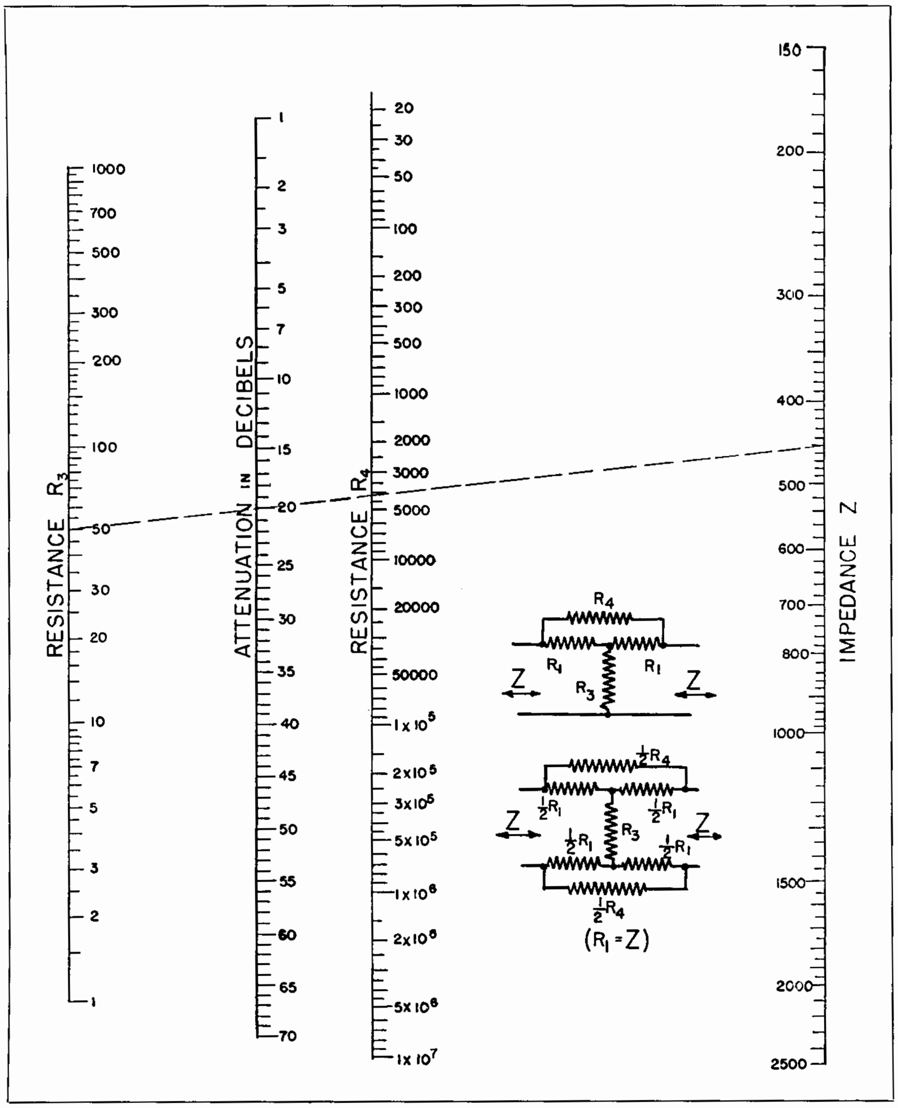
Fig. 18-10. Symmetrical T and H Attenuators. A monograph for designing symmetrical attenuators when the terminal impedance and required loss are known. A straight line through "Decibels for R_1" and "Impedance" gives value of R_1. Another line through "Decibels for R_3" and "Impedance" gives value of R_3 for T and H attenuators. Example shows design of a 500 ohm attenuator with a 5 db loss. The value of R_1 is 140 ohms and that of R_3 is 822 ohms. For symmetrical π and O attenuators, the monograph is used to determine values of R_1 and R_3. The values of R_{12} and R_{13} are then given by the following equations: $R_{12} = Z^2/R_3$ $R_{13} = Z^2/R_1$.

(Courtesy Federal Telephone and Radio Corporation.)

Practical Application

This type of network has practical application in circuits such as the one shown in Fig. 18-9L where three balanced channels are connected together via the star connection. This circuit enables two sources of alternating current to be fed into one channel or one source of alternating current to be split into two channels.

Another practical application of the star pad is shown in Fig. 18-9M where the output from an amplifier may be:—



Fif. 18-11. Bridged T and H Attenuators. The design of bridged T and H attenuators can be readily accomplished with the aid of the monograph. The known factors in bridged T and H attenuator design are usually the terminal impedance ond the attenuotion. A straight line passing through these two points on the monograph shown below gives the values of R_3 and R_4. R_1 equals the terminal impedance. In the example shown, a terminal impedance of 450 ohms and an attenuation of 20 db gives $R_1 =$ 450 ohms, $R_3 =$ 50 ohms, and $R_4 =$ 4050 ohms. (Courtesy Federal Telephone and Radio Corporation.)

(a) Disconnected from both channels,
(b) Connected to both channels, or
(c) Connected to either channel.

This arrangement may be also used to mix the output of two amplifiers into the one channel.

There are many and varied applications of this simple circuit arrangement but it must be remembered that the loss of 6.02 decibel is a disadvantage. The use of standard 600 ohm non-inductively wound bobbins considerably simplifies the construction of the pads and carbon resistors of suitable value may also be used.

Matching with Series Resistors

Another method of matching is by the use of series resistances in each branch of individual circuits as shown in Fig. 18-9J. In this circuit, channels 2 and 3 are assumed to be terminated in their correct impedances, and thus channel 1 must look like 600 ohms. Now the circuit shown in Fig. 18-9J may be simplified to that shown in Fig. 18-9I, and Fig. 18-9I, in turn, simplified to that shown in Fig. 18-9H. The joint resistance of two equal resistances in parallel may be found by dividing the resistance of one by the number of resistances connected in parallel, and in the case of the two equal resistances shown in Fig. 18-9H, the joint resistance will be:—

$$\frac{(x + 600)}{2}$$

The impedance of the network shown in Fig. 18-9H must match the impedance of channel 1 (that is, 600 ohms) and thus:—

$$x + \frac{x + 600}{2} = 600$$

therefore $x = 200$ ohms.

This value of 200 ohms represents the sum of the values of both resistors in each branch of the network and thus each resistor will be:—

$$\frac{x}{2} \text{ or } \frac{200}{2} \text{ or } 100 \text{ ohms}$$

The completed circuit will then be as shown in Fig. 18-9K. This circuit has an approximate loss of 4.44 decibels in each channel which, however, is not a very series disadvantage. The same circuit arrangements may be used with this method as shown in the previous circuits, and, while not as simple as these, it is nevertheless simpler than the standard matching pad. Another feature of the split pad is that the pad does not affect the frequency response, provided that non-inductive resistors are used.

CHAPTER 23

Acoustics

The behavior of sound and its control in auditoriums, studios and in the music room and the calculating of audio power requirements.

● The finest PA sound system in an auditorium or a music system in a living room can never produce optimum results unless the area is acoustically correct for the proper control and behavior of the projected sound waves. In the case of broadcast or recording studios—means must be provided whereby the studio acoustics can be "adjusted" to meet specific needs.

To study the behavior of the dispersion of sound requires measurement techniques that are capable of interpreting sound behavior and to show methods for its control.

Auditorium Acoustics

Acoustics comes first in auditorium design. It is basic to the purpose of the structure. Beauty means nothing if an auditorium fails to provide a meeting place where many people can hear and see a single performance simultaneously and comfortably. The problem is to project speech and music in ample loudness, with distinctness and faithful reproduction to all parts of the room. In part this means separating "wanted sound" from "unwanted sound."

Solving Structure Borne Noise

One type of unwanted sound has its source outside of the auditorium. This is partially a problem of auditorium placement. Some "outside" unwanted noises may originate within the building. They are transmitted through the structure of the walls or framework of the building, or plumbing or heating connections. This calls for insulating the noise source from the structural members by resilient materials, such as *Acoustone,* which absorb vibration.

Sound Insulating Walls

Partitions may be supported with a double set of studs—one set carrying one face, the other set the opposite face of the partition. Thus sound vibrations striking the outside face of the partition will not be as readily transmitted by the wall structure, to set the inside face of the partition in motion, producing unwanted noise in the auditorium.

To bar outside noise, all openings to the room must be effectively sealed. The cost is high and usually unnecessary. If interfering outside noise is kept about 10 to 20 decibels below the ordinary level of wanted sound, the result is usually satisfactory.

"Inside" Noise

With a completely quiet audience there can still be unwanted sound originating within an auditorium. This is reverberation or echo of the speaker's own voice, caused by the sound bouncing from walls and ceilings. It is estimated that 1/15 of a second interval between two sounds is required to enable the ear to recognize them as separate sounds. In a room with high reflection of sound, speech may be unintelligible because new syllables pile up with the reverberation of previous syllables.

Reducing Reverberation

A plaster wall or ceiling absorbs only about 3 per cent of the sound striking it. The remaining 97 per cent will bounce back as a secondary sound and be reflected again and again, losing only the same percentage of energy on each bounce. These reflected sounds pile up on each other to add to the volume of the first sound and delay its ending until seconds after the original sound ceases.

This is a problem that can be controlled with acoustical materials. They are sound absorbent and leave less sound energy to be reflected on each "bounce."

It is not desirable to eliminate all reverberation within an auditorium. It has been found that a reverberation time between 7/10 of a second and two seconds makes for pleasant and easy hearing. Smaller rooms should be near the low figure—larger auditoriums the higher.

Hints on Applying Acoustical Materials

The best method of applying acoustically absorbent material is to distribute it throughout the room as much as possible. This has led to covering an entire ceiling of a room.

Generally, absorbents should be near the rear of the room, while the front of the room near the sound source should be "live." This prevents absorption of sound energy at the source and permits reinforcing reflections to reach the rear of the room.

In wide, low-ceilinged auditoriums, it is often better to treat side walls and rear walls and leave the ceiling bare to aid reflections to the rear of the room. Usually the ceiling under a balcony should not be treated, since sound levels are often low in that area.

The following locations should always be checked in considering the location of acoustical absorbents:

The ceiling—The rear wall—Side walls of long, narrow "shooting gallery" type of rooms or low-ceilinged auditoriums—The face of the balcony in theaters—Ceilings and side walls (down to wainscot) in high-ceilinged corridors.

Sound Distribution and Size Limits

A room corrected to achieve desirable reverberation conditions can be a poor hearing room if the sound is not well distributed through the area. It is difficult for the average speaker to make himself heard in a room with a volume greater than 400,000 cubic feet (length x width x ceiling height). Two hundred thousand cubic feet is a preferable volume. In rooms of greater size, electrical amplification is necessary.

"Dead" Areas

Even in rooms within the above size limits, there are sometimes found "dead" areas where very little sound is received, or spots of confused sound. Usually these result from concave wall surfaces which focus reflected sound, or from room shapes which prevent reinforcing reflections of sound from reaching remote areas.

These few simple rules often eliminate many of these problems:

1. Avoid concave curved surfaces.
2. If curves are necessary, use convex.
3. Substitute flat areas for curves even if it means breaking up a large curve into a series of straight lines.
4. Avoid large unbroken surfaces. Use projections and surface irregularities to aid in diffusing sound.
5. Avoid complicated designs which theoretically focus sound to remote areas. They are usually unsuccessful.

Public address amplifiers are a good investment for sound distribution, but they do not remedy excessive reverberation or act as a substitute for acoustical treatment.

Sound Fidelity

Sound is a component of a gamut of tone frequencies, and the third aim of scientific acoustics is to achieve fidelity or naturalness of sound.

A long narrow tube-like room, or one with large floor area and low ceiling might provide distortion by reducing or increasing volume of certain frequencies. Even some acoustical absorbents have poor absorption at certain useful frequencies, or "peak" in absorbing efficiency at one or two frequencies, to cause distortion.

MEASUREMENT OF STUDIO AND ROOM ACOUSTICS

The field of sound reproduction has at the present time reached the stage of development where it is possible to reproduce sounds from a loudspeaker, after numerous stages of recording and transmission, which will sound almost indistinguishable from the original sounds which entered the microphone. In addition, present electronic and electromechanical techniques are constantly being used to improve sound reproduction equipment toward the ultimate purpose of making the reproduced sound identical with the original.

However, no matter how well the output sound reproduces the input, the entire character of the reproduction can be drastically altered by faulty acoustic design either in the room where the sound originates, or in the room where it is reproduced. Bad acoustic design can result in loss of intelligibility and "presence," increased noise level and reduction of dynamic range, resonances and spacial distribution defects, and generally make good program material unpleasant to the ear. This factor has long been recognized, and considerable work has been done in the design of studios, rooms and auditoriums to determine how to attain the best acoustic qualities. The problem is a complex and difficult one, and has still not been solved to complete satisfaction, although considerable progress has been made toward its solution.

Even under ideal conditions, the design of any room, studio or auditorium is difficult because of the limitations imposed by architectural factors. Therefore compromises usually must be made in designing for optimum acoustic performance. Then, once the room has been completed it is tested to see how well it meets the performance requirements. Methods of testing[1] form an extremely important part of any type of design procedure, and in acoustics this is especially true. The ideal test gives a measure of the performance, indicates what may be wrong and by how much, and gives some indications of what steps may be taken to correct any defects which may exist.

Such tests have only recently been developed for acoustic measurement, and have considerably increased our knowledge about what factors are important in determining the acoustic quality of a room, and what their effects are. We will describe these measurement techniques, and show how they aid in the improvement of acoustic designs.

In order to understand the methods and equipment used for the measurement of studio and room acoustic properties, it is first necessary to understand what factors are involved and their effects upon any sounds which are present in the room.

Specific Factors Which Determine Acoustic Properties of Rooms

When sound is listened to in a room or an auditorium, the room has two important effects: (a) it reverberates and reflects the sound from the walls, ceiling and floor; otherwise, if there were no reverberation, the sound would appear as if it were being heard in a completely open space; (b) it excludes external noises. Therefore measurements of the acoustic properties of rooms must concern themselves primarily with various types of reverberation and noise measurements.

Reflections from the boundary surfaces of the room, which basically determine its acoustic character, can have several different effects depending upon the nature of the sound which is heard. These different effects require different measurement techniques.

In general, a number of measurements must be performed before it can be determined whether the acoustic properties of a room will be acceptable. The factors which should be known include the following:

(1) Reverberation and reverberation time, including
   (a) fluctuation during decay
   (b) echo and flutter echo
(2) Sound diffusion (and sound concentrations)
(3) Transient characteristics
(4) Noise level

and when the room is the one in which the sound is being reproduced, it is also desirable to know the relationship between the reproducing system and the room acoustics, as given by:

(5) Power output of the reproducing system
(6) Frequency response of reproduced sound

Some of these factors have been studied extensively, and standards determined which correlate the measured value with the acoustic performance. In the case of a few of the above factors, standards have not yet been determined, but practical experience has shown what should be the requirements for acceptability.

When a sound is started in a room the intensity does not immediately reach its maximum, because it takes an appreciable time for some of the sound to reach the walls and undergo one or more reflections before it reaches the listening point. The intensity reaches its maximum when the steady-state condition is attained. After the sound source stops, it also takes an appreciable time before the various reflections are no longer heard, having been completely absorbed. This persistence of sound is called *reverberation*, and is different from an echo in that it consists of a large number of reflections which blend evenly with one another and with the original sound. The *reverberation time* has been defined as the time required for the sound intensity to decrease 60 db after the source has been stopped.

When there are large flat surfaces, these may give rise to distinct reflections which are heard as *echoes* if the path difference is too great. When parallel walls are located opposite one another, there may be heard a succession of distinct reflections between them—this effect is known as *flutter echo.*

The presence of concave surfaces tends to focus sounds towards their center of curvature, giving a greater sound intensity than at other points in the room, and creating the impression that the sound originates at the concave surface. If standing-wave patterns are possible at certain frequencies, these frequencies will tend to be over-accentuated at positions where the standing-wave patterns are set up. For best acoustic properties, the sound pattern in the room should be as *diffuse* as possible at all frequencies, with no standing-wave patterns and no points of excessive sound concentration. Under these conditions, the decay of sound (reverberation) when the source stops will be smooth, marked only by small fluctuations.

The reverberation time and the sound diffusion characteristics of a room are essentially steady-state characteristics, since the sound is allowed to reach equilibrium conditions before these factors are measured. However, all natural sounds are essentially transient in nature, therefore the behavior of the room for transient sounds is of great importance. It is necessary also to know whether the steady-state reverberation time and sound diffusion are accurate for transient sounds, and what differences may exist. In many cases the transient characteristics are much more important than the steady-state, and sometimes give much more information about the characteristics of the room.

The ease with which the reproduced sound may be heard and understood,

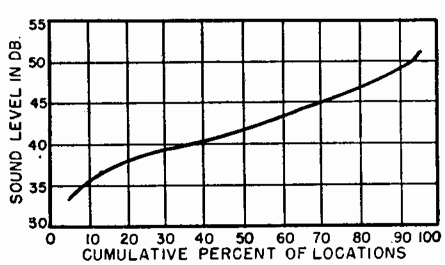


Fig. 23-1. Graph showing the residual noise level in homes.

and the dynamic range which is possible, depend upon the residual noise level in the room. The tolerable noise level in the studio and in the reproducing system depends upon the noise level in the listening room. The average noise level in empty theaters is 25 db (reference level is 10^{-16} watts per square centimeter); with an audience the average will generally be about 42 db. The noise level of most residences is given in Fig. 23-1, which shows that there is a wide variation in noise from one location to another.

In any audio reproduction, it is important also to know the relationship between the reproducing system and the acoustics of the listening room. The power output from the loudspeaker should be capable of producing a minimum sound intensity of 80 db, and for best performance should be capable of producing a level up to 100 db without distortion. Fig. 23-2 shows the acoustical power required, as a function of the room volume, to produce a sound level of 80 db.

The sound output from the loudspeaker should have a flat frequency response characteristic. The function of the reproducing system is to reproduce at the ear of the listener a duplication of the sound which is present at the microphone, and to affect the tonal qualities as little as possible. In certain cases it is necessary to restrict the reproduced frequency range because of acoustic or reproduction difficulties, but in such cases the restriction is a compromise rather than a desirable situation.

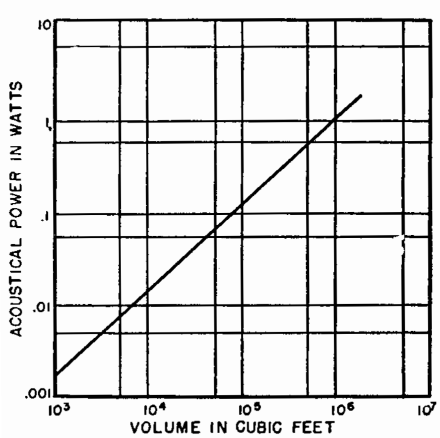


Fig. 23-2. Acoustic power required to produce an intensity level of 80 db as a function of the room volume.

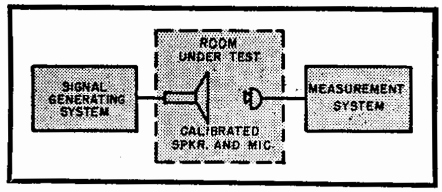


Fig. 23-3. A basic test setup for performing acoustic measurements in rooms.

General Technique of Acoustic Measurements

Acoustic measurements consists essentially of generating a known sound signal in the room which is under test, and determining the resulting sound distribution. The basic setup for performing measurements of this type is shown in Fig. 23-3. A signal generator of the proper type supplies the desired signal which is applied to the calibrated loudspeaker. (The signal generator here is taken to include not only the generator of the low-level signal, but also any auxiliary amplifiers which may be necessary to increase the electrical signal level before applying it to the loudspeaker so that the necessary amount of sound energy may be supplied for testing). When testing a room in which sound is to be reproduced, it is often preferable to use the amplifier and loudspeaker system which is already installed; then the electrical test signal is applied directly to the amplifier input. The sound in the room is picked up by a microphone of known characteristics, whose output is amplified and applied to the measuring device. The specific type of signal which is generated, and the type of measurement system, will be determined by the particular acoustic characteristics under test.

Generally, a calibrated loudspeaker will not be available, whereas standard calibrated microphones are readily available. It is not necessary that both the loudspeaker and the microphone have known characteristics, since if the characteristics of one are known it can be used to calibrate the other. Therefore only a standard microphone is necessary to obtain completely accurate and reliable acoustic measurements. A calibrated microphone which has been widely used for this type of service is the condenser microphone (see Chapter 14). These microphones are calibrated against a primary standard sound source, and may therefore be used as secondary measurement standards. Because of its small physical dimensions this type of microphone is effectively a "point pickup" which does not appreciably disturb the sound field, and it has good frequency response characteristics up to approximately fifteen thousand cycles per second.

The characteristics of the loudspeaker may be calibrated in terms of the microphone characteristics. However, such a calibration must be done in such a manner that the acoustics of the measuring room do not affect the results. The measurement must be performed in what is known as "free-field" room. The requirement of such a room is that all reflected sound and the noise level be so low that they can be neglected in the measurement. The simplest and most direct method of obtaining these conditions would be to perform the calibration out of doors at a great distance from reflecting objects, if favorable weather and noise conditions can be obtained. Free-field conditions are also achieved in special rooms which are carefully designed to have extremely small reflections from the boundary surfaces. In practice the best measuring room available will be a small deadened or partially deadened room. In such cases, the most satisfactory results are obtained by placing the microphone close to the loudspeaker so that the level of the direct sound strik-

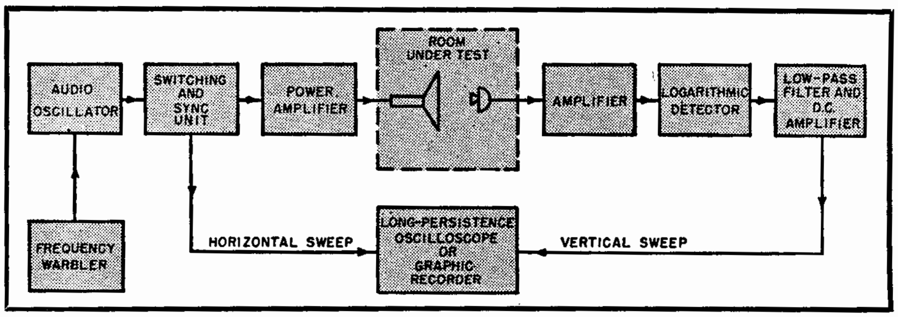


Fig. 23-4A. Measurement setup for determining reverberation time.

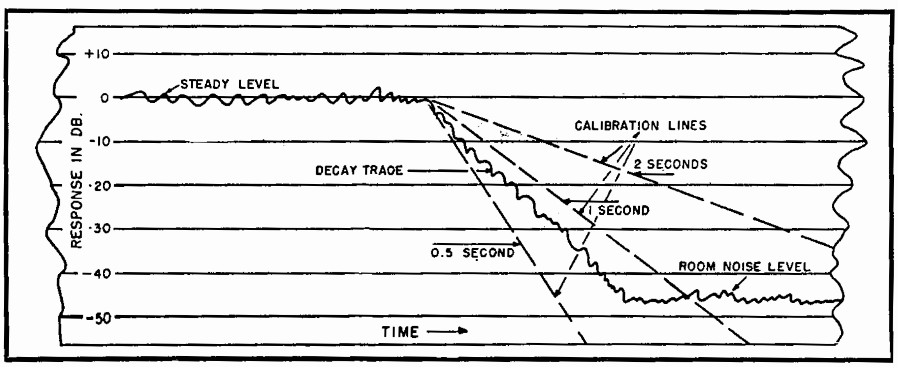


Fig. 23-4B. Typical decay curve obtained in measuring reverberation with a graphic recorder.

ing the microphone is at least 20 db or more above the reflected sound.

In most acoustic room measurements the effects of standing waves are undesirable and should therefore be minimized. This can most readily be done by frequency-modulating the test signal (usually called "warbling") by about $\pm 10\%$ of the mean frequency, at a rate of several times per second. When this method is used there will be continuous small changes in the standing-wave pattern, but resonances will not have a chance to build up.

Measurement of Specific Acoustic Characteristics

The basic setup of Fig. 23-3 is used for measurement of all the various factors that represent the acoustic properties of studios and rooms. Different types of signal generators and measuring devices are used, according to the specific factor being measured.

The method of measuring reverberation time is indicated schematically in Fig. 23-4. The signal is generated by an audio oscillator set to the desired frequency and warbled. The output of the oscillator is applied to the switching and synchronizing unit, which controls the sequence of measurement operations. The signal is then amplified by a power amplifier which drives the loudspeaker that generates the sound signal. The sound in the room is picked up by the microphone and amplified, then detected by a logarithmic detector to give a dc reading on a (decibel) scale, and fed to a low-pass filter. The output of the filter is then amplified by a dc amplifier. The amplified output, which gives the reverberation decay characteristics of the room, may be observed either by means of a graphic pen-and-ink level recorder or upon the screen of a long-persistence oscilloscope.

The switching and synchronizing unit turns on the sound source for a time long enough for steady-state conditions to be reached, then switches off the signal and permits the sound in the room to decay. The microphone picks up the sound intensity in the room at all times, and the sound intensity at the microphone is plotted upon the oscilloscope screen or by the graphic recorder. The decay of sound from the moment the source is switched off is observed, and the slope of the decay curve is measured to give the reverberation time for 60 db decay. There may be fluctuations during decay of the order of 10 or 20 db, but the average slope is the one which is used. In estimating the decay time it is preferable to use the initial slope, since this is the most important to the ear and the remaining portion of the decay is normally masked by subsequent sounds. The presence of large-scale fluctuations and changes in the average slope of the decay curve indicates that the room does not have a completely diffuse sound pattern, and that best acoustic performance has not been achieved.

When the sound decay curve is being measured by a graphic recorder, the measurement setup shown in Fig. 23-5 may be used without the synchronizing and switching unit. In this case, the sound source is turned on and kept on long enough for steady-state conditions to be reached, then the paper is allowed to run in the recorder and the sound source switched off. The sound decay pattern will then be recorded.

Because of the cyclically changing standing-wave pattern due to the frequency-modulated signal, when the graphic recorder type of measurement is used the recorded decay curve will be different depending upon the exact time at which the signal is cut off. With the

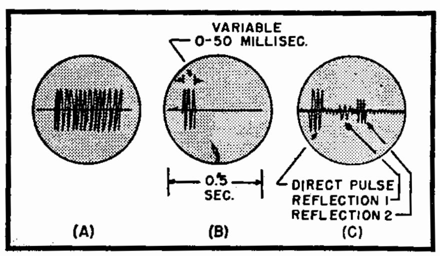
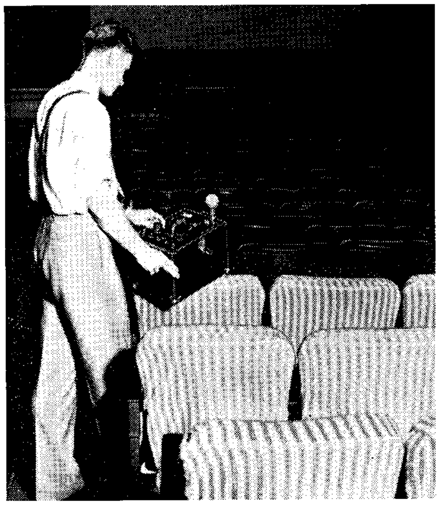


Fig. 23-5. Typical oscillograph pictures of wave shapes in various parts of a test setup (A) audio oscillator output (B) sound output of loudspeakers (C) sound received at microphone.



The General Radio type 759-B sound level meter being used to perform acoustic measurements in a motion picture theater.

oscilloscope method of measurement these fluctuations tend to average out, due to the superposition of a number of different decay curves which in general start at different times in the warble cycle. The effects of standing-wave patterns and interference effects can be further smoothed out by the use of multiple loudspeakers and microphones. In practice, it is desirable to reduce these errors by taking several measurements for each of several locations of loudspeaker and microphone, differing in position by about one yard. If three readings for each of four different positions are taken, accuracy in reverberation time to about 0.1 sec. can be obtained.

The degree of sound diffusion is measured mainly by observing the standing-wave pattern in the room when sound is present. Some indications can be obtained from the reverberation characteristic, but such observations are not too good because in measuring reverberation, steps are taken to eliminate the effects of standing waves. The simplest and most direct method of determining the standing-wave characteristics of a room is to produce a steady sound in the room and survey the room with the microphone to determine the intensity pattern. (In this type of measurement an omnidirectional microphone should be used, and the directional pattern of the loudspeaker radiation taken into account.) With complete diffusion the sound intensity will be uniform throughout the room for all frequencies, or will vary gradually with position according to the directional characteristics of the loudspeaker and the absorption by air of the higher frequencies. The relative intensity of maximum and minimum points will be a measure of the diffusive character of the room, and any sound concentrations will also be detected. Another method of performing this measurement is to keep the microphone fixed and slowly sweep the signal frequency over the entire audio range. Assuming the frequency characteristics of the loudspeaker and the microphone to be reasonably flat, variations in response will indicate standing waves in the room. However, this latter method does not indicate whether there may be any concentrations of sound at various points in the room.

The transient characteristics are measured by applying a test signal which has transient properties similar to those of natural sounds, and observing the resulting sound at the loudspeaker. This method has the advantage that the results can be expected to correspond closely to the actual conditions under which the room will most often be used. The complete test setup for this type of measurement is shown in Fig. 23-6. For a permanent record, the oscilloscope screen may be photographed.

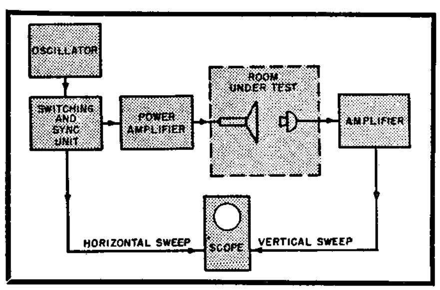


Fig. 23-6. Setup for performing transient acoustic measurements.

Otherwise, a graphic recorder may be used with a low-pass filter and dc amplifier as used in the reverberation-time measurement system shown in Fig. 23-4. In addition to the transient acoustic characteristics of the room, this system also gives considerable information concerning echoes and the location of the various reflecting surfaces which give rise to echoes and large-scale reflections.

The signal wave shapes are shown in more detail in Fig. 23-5, to give a better indication of the type of data obtained with this method of measurement. The output of the audio oscillator is a continuous sine wave which may be set to any frequency at which the acoustic characteristics are desired. The switching and synchronizing circuit contains a gating mechanism (either a motor-driven cam-operated switch or an electronic gating circuit) which permits the signal to pass in pulses as shown in Fig. 23-5B. The signal pulse length is adjustable from 0 to 50 milliseconds duration and is repeated at intervals of about 1 second, so that the reflected sound decays to a negligible value before the next impulse. The horizontal time scale on the oscilloscope screen can be set for a sweep time of 0.5 sec. across the face of the screen. (If a graphic recorder is used, the switching and synchronizing circuit will be set for just one signal pulse, and the recorded response will be the response to this one pulse.) The type of signal received by the microphone consists of the direct sound pulse plus whatever reflections there may be from any parts of the room, as shown in Fig. 23-5C. By measuring the time taken for any reflections to arrive at the microphone after the direct pulse, and by actually laying out and plotting the various possible sound paths between the loudspeaker and the microphone in the room under test, the location of the various reflecting surfaces can readily be determined.

In practice, each measurement should be taken at three different positions at each location in the room, and at several different frequencies over the entire audio-frequency spectrum. A total of at least 10 to 15 different pulse reflection measurements should be averaged for each location. This type of averaging will tend to cancel out any spurious spatial or frequency effects.

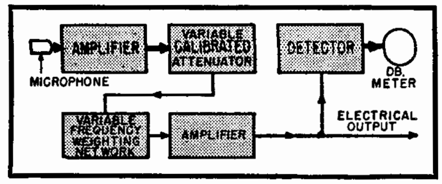


Fig. 23-7. Basic block diagram of a sound-level meter.

Sound Level and Power Measurements

Noise level and sound power output are measured by use of a sound-level meter. The basic block diagram of the standard type of sound-level meter is shown in Fig. 23-7. The sound is picked up by a unidirectional microphone with a known frequency-response characteristic. The output of the microphone is then amplified and passed through a calibrated attenuator which serves to set the meter range. The signal is then passed through a frequency weighting network which can be set for either flat response or for either of the standard noise-measurement response curves. The output of the frequency weighting network is then amplified and measured by a vacuum-tube voltmeter calibrated to read logarithmically in decibels. The output signal is also available before rectification for operation with graphic recorders or with various types of analyzers. The meter reading is accurately calibrated in decibels relative to the standard 1000 cycle/sec. reference level of 10^{-16} watts per square centimeter.

When noise level is being measured, a truly objective measurement is impossible because of the complexity of the human hearing mechanism and because of the wide variety of noises which may be encountered. However, a reliable indication of the noise level is obtained by taking into account the frequency response of the human ear, and making

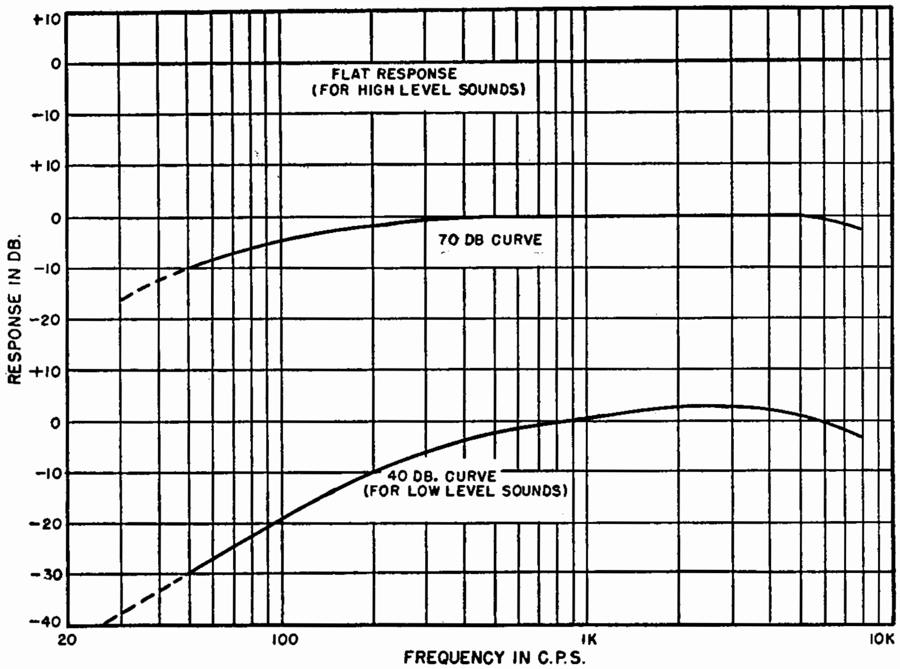


Fig. 23-8. Frequency response characteristics recommended as standard curves for noise level measurements.

the overall response of the noise meter approximately the reciprocal of the ear response characteristic. This condition is approximated by using three different frequency characteristics for the meter for different sound levels. The three response curves which are chosen by the American Standards Association as the standard curves for noise level measurements are shown in Fig. 23-8. Curve A is recommended for measurement of low levels around 40 db; curve B for levels around 70 db; and curve C, which is flat, for very loud sounds around 80 to 100 db. The actual measurement of the noise level is performed simply by having no source of sound in the room and reading the sound level on the meter.

The sound power output of the reproducing system is measured by feeding steady tone (warbled if necessary to reduce standing waves) into the reproducing system and measuring the resulting sound intensity, with the sound-level meter set for flat frequency response. The electrical signal at the auxiliary output of the sound-level meter can also be fed to any of the standard instruments for measuring the various characteristics of audio-frequency electrical signals—harmonic analyzers, intermodulation analyzers, etc. Measurements of this type performed at various frequencies will give the characteristics over the entire audio frequency range.

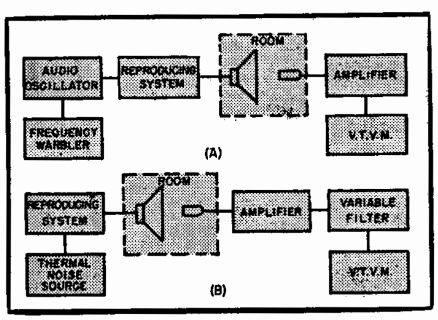


Fig. 23-9. Measurement of frequency response by means of (A) single frequency method (B) a thermal noise generator.

The frequency response of the complete system including the loudspeaker can be measured by using the basic measurement system in the manner shown in Fig. 23-9A. The method is the same as for measuring frequency response of any electrical circuit, except for the warbled frequency. The electrical signal is applied to the input of the system under test. The sound output of the loudspeaker is measured by means of a microphone, amplifier and meter whose frequency characteristics are accurately known. The frequency of the test signal is then set as desired, and the meter read, to give the response characteristic over the entire audio frequency range. The microphone can also be placed in various locations throughout the room to give the spatial radiation pattern as well.

Another method of measuring frequency response is by means of a thermal noise generator and a tunable filter in the microphone amplifier circuit, as shown in Fig. 23-9B. The signal is supplied by a source of thermal noise, such as a diode, and is applied to the input of the reproducing system. The output of the loudspeaker is then picked up by the standard microphone, amplified and passed through a narrow band pass filter, whose band width should be independent of frequency. The output of the filter is then measured by the meter. At the present time, suitable apparatus for the generation of thermal noise, and band pass filters of the type mentioned, are commercially available and this type of measurement will in the future become very important for acoustic measurements.

Results of Acoustic Measurements in Practice

The methods which have been described have been used to determine the acoustic characteristics of rooms and auditoriums in order to obtain a measure of their performance, to aid in their redesign and improvement when they do not give optimum performance, and to obtain information to aid in new constructions.

Many measurements of reverberation time have been made in the past, and much data has been accumulated on this subject. There is no theoretical basis for the choice of desirable reverberation times, but experience has shown what is most pleasing to the ear, and standards have thus been determined subjectively. Early experience with broadcast studios has shown that when there is no reverberation the room gives a dull, lifeless effect to sounds. However, when there is too much reverberation, the energy from successive sounds tends to overlap and reduce intelligibility. The optimum reverberation time is a function of the volume of the room, and rooms for listening to reproduced music should have shorter reverberation times than those for live production of the same type of music because the reproduced music will already contain some reverberation from the production studio.

The optimum reverberation times for a 1000 cycle test signal, are shown in the graph in Fig. 23-10A. The optimum reverberation time as a function of frequency relative to the 100 cycle value is shown in the graph of Fig. 23-10B. The values shown in these curves do not, of course, take into account the possibilities of microphone placement and synthetic reverberation systems which are

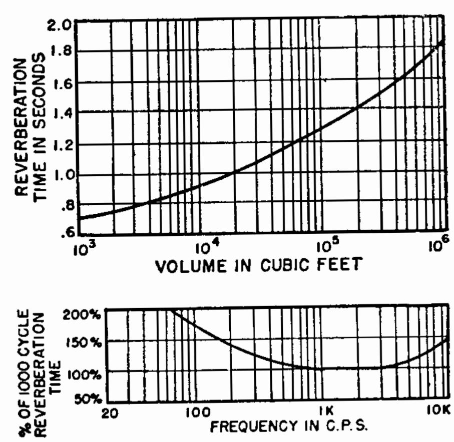


Fig. 23-10. Optimum reverberation time as a function of (top) room volume for a 1000 cycle test signal and (bottom) frequency.

used to increase the apparent reverberation time and "presence" in the reproduction of speech and music.

For a long time the acoustic qualities of room and auditoriums were judged primarily on the basis of reverberation times. However, experience began to show that it was possible for rooms to have the same reverberation time and still to have quite different acoustic properties. Measurements of the diffusive and transient characteristics show that at times these facts are considerably more important than the reverberation time, and at the present time these are being given increasing importance in acoustic measurements.

The pulse method of measuring transient characteristics is an extremely important method, and often gives much more valuable data than the reverberation time and other methods. In many cases it is the only method of correlating measured data with the results observed by the listener, when other methods fail. The results of such measurements upon a number of typical auditoriums show the type of information that can be obtained. The pulse patterns shown in Fig. 23-11 show the results of measurements on a number of moving-picture houses whose acoustic qualities had received different degrees of acceptance by listeners over a period of several years.

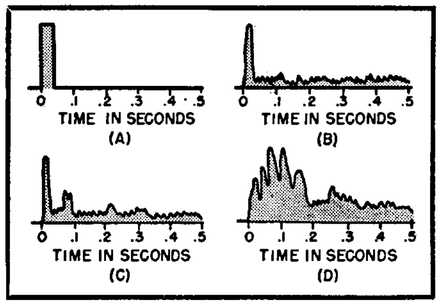


Fig. 23-11. Pulse patterns showing results of transient characteristic measurement on several different theaters.

An investigation was undertaken to determine the causes of the acoustic differences, since the theaters had identical sound reproducer installations, and in all cases the measured frequency characteristics and the reverberation time were found to be satisfactory. The pattern (A) (Fig. 23-11) shows the pulse output of the loudspeaker, which is what the microphone would pick up in a room with no reverberation. Pattern (B) is the sound picked up by the microphone in a theater with uniformly good acoustics; the physical structure of the theater is shown in Fig. 23-12A, showing that there are no undesirable reflections. The pulse pattern represents a bad spot in an otherwise good theater whose layout is shown in Fig. 23-12B. The measurement shows a reflection from the back wall at 80 milliseconds delay, and a further reflection at 220 milliseconds delay which seems to be due to a multiple reflection as shown. Pulse pattern (D) was taken in an auditorium of inferior quality, whose layout is shown in Fig. 23-12C. Large reflections are found at both short and long time intervals, and are the reason for the bad quality.

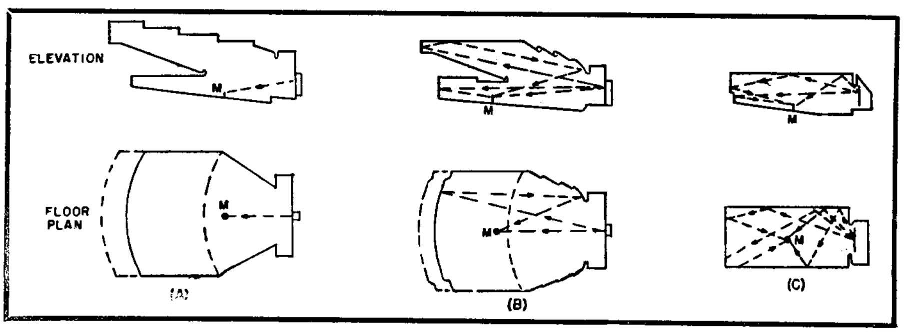


Fig. 23-12. Physical layout of theaters showing reflection paths for the various pulse echos.

In general, reflections with less than 45 milliseconds delay can be tolerated, but reflections with more than 50 milliseconds delay lead to a deterioration in sound quality due to lack of intelligibility. When there are large reflections at short time delays which arrive to the listener at large angles from the path of the direct sound, the directional effects of the sound are lost, resulting in a loss of "presence." In auditoriums where acoustic conditions are not optimum, the pulse technique also gives good indications of the possible locations of the reflections, and thus aids in correcting any defects in the acoustic design.

These measurements have indicated what the basic points in good acoustic design are, and what rules should be followed in the design of rooms, studios and auditoriums. Some of these rules are:

(a) Maximum sound diffusion should be aimed for in all acoustic designs.

(b) The room should be as unsymmetrical as possible (with no lines or planes of symmetry), and if possible there should be no walls parallel to one another, and no concave surfaces.

(c) Large surfaces should be broken up by randomly distributed irregularities such as convex spherical bumps and cylinders, and serrated surfaces. Absorbing material broken into small patches also aids diffusion. At the present time, radio broadcasting studios, theaters and auditoriums are being built according to these rules for best acoustic qualities.

The measurement methods which have been described in this chapter are being more and more widely used to give an objective indication of acoustic quality, and their application will result in continuing improvements in acoustic design and construction.

Audio Power Requirements

Most sound engineers have acquired enough experience to make good estimates of audio power requirements and equipment manufacturers try to supply packaged units to fit different types of situations. The purchaser can usually take his recommendations and obtain the desired results. However, most professional sound men will want to be able to make their own estimates. The present science of acoustics provides a basis for calculating the acoustic and electrical power required to produce a given sound intensity in different types of rooms. A few simple equations are all that is involved and these have been combined into a simple nomograph which will be described. With this graph the power required to produce any sound level in nearly all types of rooms can be estimated in a few seconds.

It should be stated that the accuracy of the graph is limited. This is not the fault of the equations involved but because of the difficulty of defining sound levels and due to the lack of accurate data on speaker efficiency. Even so, the graph will allow some useful estimates and, with a suitable safety factor, amplifier wattage ratings may be selected. The graph is also a very revealing source of information on the change in amplifier output at different sound levels and in different types and sizes of rooms.

Loudness is a sensation in the mind of the hearer and is a rather complicated matter but it is generally related to sound pressure or intensity. A great many measurements have been made of the sound intensities produced by various noises and musical instruments. A little study of this data will provide a measuring stick by which we may refer to various degrees of loudness. Fig. 23-13 shows several of these measurements on the standard decibel scale. These decibel ratings are obtained by measuring the intensity of the sound waves and by using the formula: *Sound intensity in decibels* $= 10 \log_{10} (I/I_o)$ where I is the intensity being measured and I_o is the standard sound intensity. This standard is usually selected as the weakest 1000 cycle tone that can be heard by a normal ear in a silent room. This is a sound intensity of 10^{-16} watts per square centimeter. Since the logarithm of one is zero the standard level is expressed as 0 on the decibel scale. The decibel scale is a

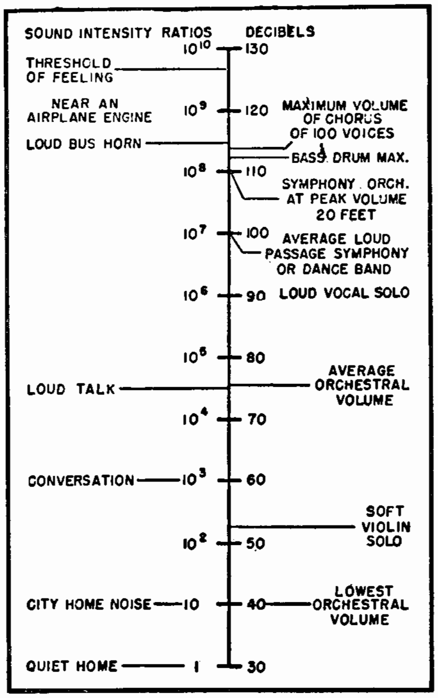


Fig. 23-13. Loudness levels of everyday sounds shown in db and in sound intensity ratios.

ratio scale and this is especially useful in loudness measurements because the ear responds to sound intensity in approximately a ratio function. See Chapter 8. That is, each time the sound level doubles in intensity it sounds approximately like an increase in equal steps of loudness. Thus an increase in sound intensity from 35 to 40 decibels sounds about the same as an increase from 55 to 60 decibels. In both cases the sound power increases by a ratio of about 3 but the actual power increase is 100 times greater in the second case.

Fig. 23-13 shows the relationship between the sound intensity in acoustic power ratio versus the decibel notation. The graph covers a range of 10 billion-fold in sound pressures, from 30 decibels, the sound level of an average quiet home, to 130 decibels, the threshold of feeling or pain. Note the decibel ratings of the different sound levels that an audio system will be called upon to reproduce and the tremendous sound intensity ratios which this represents. It should be remembered that most of these sound measurements have been made with sound instruments which have a certain time lag. With complicated sounds the instantaneous sound pressure may be considerably higher. Engineers generally recommend that a system be capable of producing sound levels 6 decibels above the maximum sound meter readings. In Fig. 23-13, 113 decibels is the loudest musical sound measurement made, but instantaneous sound peaks may extend to 120 decibels and this safety margin should be kept in mind.

How loud should an audio system sound? In its design it must be planned for the maximum level that will be required. Most home radios reproduce music at about 20 to 40 decibels lower than the original sound source. Of course it makes a difference if the music is to be a background for conversation or if it is an occasion for serious listening or dancing. High fidelity and music fans often listen at normal volume levels even though the program is not always played that way. Dance bands and symphony orchestras have a characteristic sound because people enjoy that kind of loudness and most listeners have similar preferences when low distortion audio is available and listening conditions permit. Normal volume level is not the same thing as putting a symphony orchestra in a small living room. That would be much too loud. We simply want to recreate the loudness heard in the symphony hall in the living room. Naturally much less acoustic power will be required. In reproducing music at lower-than-normal levels there are not the savings in power requirements that might be expected. This is because of the increased bass boost needed for compensating the loss in aural sensitivity to bass notes at low volumes. The *Fletcher-Munson* curves show that if a musical passage, normally heard at an 80 decibel level, is reproduced at a 40 decibel level about 30 decibels of bass boost at

50 cycles is required for balanced listening. Since the bass tones contain much of the peak power this 40 decibel reduction in listening level will allow a reduction in power requirements of only 10 decibels. Normal loudness levels are, therefore, a reasonable requirement for high quality audio equipment designed for listening to music. For other purposes different loudness requirements must be met and these will have to be decided by the designer. The graph will aid in calculating power requirements for any sound level.

In the open air sound intensity diminishes with the distance from the source of the sound but in a room or small hall there is a great deal of reflection so that a source of sound builds up the loudness to a fairly uniform level. The ability of the sound source to produce a given sound intensity in a room depends on the size of the room and the amount of sound reflection or absorption in the room. The properties of a room affecting the reflection or absorption of sound are measured by the reverberation time and were previously discussed. The reverberation time of most living rooms is within ⅓ of a second to 1 second. Small halls may go up to 2½ seconds. Values of about ½ to 1 second are considered quite good acoustically for living rooms or small halls.

There is a simple method for calculating the reverberation time of a room by summing the effective absorption areas of the entire room. Each area of the room of different absorption surfaces is multiplied by the coefficient of absorption for that material and the total for the entire room is obtained and is called A. The reverberation time in seconds is therefore: $T = .05V/A$ where V is the volume of the room. The table, Fig. 23-14, shows a typical calculation of this sort. This calculation was made on a room to estimate the amplifier requirements.

It is possible to make a shrewd guess of the reverberation time of a room with a little practice at clapping and listening. The method is to make sure the room is quiet, clap the hands loudly and sharply and listen for the time of die-away. Careful listening will reveal a noticeable time of die-away even in a heavily draped room. It is surprising how different rooms can be in this respect. A stop watch is helpful in learning to estimate fractions of a second. The clap should be loud at about 90 or 100 decibels. Since the noise level of the room will be about 30 or 40 decibels the sound will be masked by the room noise after it has dropped about 60 decibels or the standard one millionth of its original loudness. Check your first listening estimates against the set of calculations shown in Fig. 23-14.

ABSORPTION COEFFICIENTS

Glass	.025
Plaster, Brick, Linoleum	.03
Wood Panel	.08
Carpets, Rugs	.40
Draperies, Upholstered areas	.50
Wood Chairs, Small Tables (each)	.3
Persons (each)	4.0

$T = .05V/A$

T = Reverberation time in sec.
V = Volume of room in cu. ft.
A = Absorption power of room in effective sq. ft.

EXAMPLE

Small living room (13′ x 16′ x 9′)

Plaster (1061 sq. ft. × .03)	31.8
Wood areas (56 sq. ft. × .08)	4.0
Carpet area (120 sq. ft. × .40)	48.0
Draperies and upholstery (57 sq. ft. × .50)	28.5
Wood chairs and small tables (5 × .3)	1.5
Persons (3 × 4.0)	12.0
	125.8

$T = .05\ (13 \times 16 \times 9)/125.8 = .75$ sec.

Fig. 23-14. The calculations of reverberation time.

The power required to produce a loudness or sound intensity of 120 decibels is: $P = 0.00012\ V/T$ where P is the power in watts and V is the volume of the room in cubic feet and T is the reverberation time in seconds. Other sound intensities require proportionately different power levels. The power referred to in this equation is the acoustic power, the sound power actually put out by the speaker, not the power put into the speaker. In the room used in the calculations of Fig. 23-14 the acoustic power required to produce this sound level of 120 decibels is: $P = 0.00012 \times 1872/0.75 = 0.30$ watts.

Loudspeaker Efficiency

Speaker systems differ a great deal in efficiency, covering a range of about 2 to 40%. Few speaker manufacturers specify the efficiency of their speakers in a way which allows a calculation of power input to acoustic output but a little experience on this point can aid in making estimates. If one is fortunate to live in a city with a good comparative speaker listening studio it is possible to prepare a list of estimated efficiencies. Play some music and switch back and forth between two speakers, one with a known efficiency and, with an attenuator calibrated in decibels, cut the level of one until they are equal. The decibel attenuation allows a simple calculation of efficiencies. The speaker enclosure affects the efficiency and comparisons should be made while the speakers are housed in the type of cabinet which you expect to use.

As a general rule the more expensive the speaker the higher its efficiency, but this is not always true. Large magnets and light voice coils increase efficiency. Unfortunately good bass response does not go with light voice coils and diaphragms. Efficiency is also improved by good coupling of the speaker with the air as in horn systems and bass reflex cabinets. Standard speakers in the $10 class are generally about 2 to 3% efficient. Wider range speakers costing up to about $30 are generally not more efficient, the extra cost usually going into better bass and high frequency response. A number of speakers in the $40 to $70 class are about 4 to 6% efficient. These usually have magnets of 2 or 4 pounds as compared with ounces in the lower cost models. Higher efficiencies are obtained on some types of speaker systems using horn loading. The *Klipsch* corner horn system claims an efficiency of the order of 30%.

To return to the example shown in Fig. 23-14, the room is to be equipped with a coaxial speaker of good efficiency in a bass reflex cabinet. An efficiency of 5% is estimated for this model and so the electrical power calculated to produce the sound intensity of 120 decibels is: *Electrical power = Acoustic power/Speaker efficiency;* $Pe = .30/.05 = 6$ watts.

It is informative to calculate the power that will be used in this installation under different listening levels. At 113 decibels (top loudness of a symphony orchestra) the electrical power is 1.2 watts. Of course the instantaneous peaks may be above this level by a 6 decibel factor, or at the 6 watt output already calculated. The average loudness of a symphony orchestra will use only about a thousandth of a watt and lower volume levels will be down to a millionth of a watt and less.

Assume that a 10 watt amplifier would suffice for this installation. Output measurements of the amplifier at various loudness levels are made and readings obtained, for example, as follows:

	Calculated	*Measured*
Loud symphonic peak	1.2 w.	.6 w.
Average loud passage	.1 w.	.08 w.
Average music	1 mw.	2.4 mw.

The calculations outlined can be performed on the simple nomograph shown in Fig. 23-15. The method is as follows. Select the loudness level wanted for the room in question. Place a straightedge on this point on line 1 and also on the point on scale 2 corresponding to the volume of the room. Mark the crossing of the straightedge on line 3. Connect this point to the reverberation time of the room on scale 4. The intersection on scale 5 gives the acoustic power required. To change this to electrical output connect the point on scale 5 to the speaker efficiency on scale 6. The intersection on scale 7 shows the electrical power required. The right hand of the scale reads in watts and the left hand side gives the answer in decibels with a zero reference of .006 watts.

This nomograph has been used in estimating power requirements in a number of different applications. It has proven

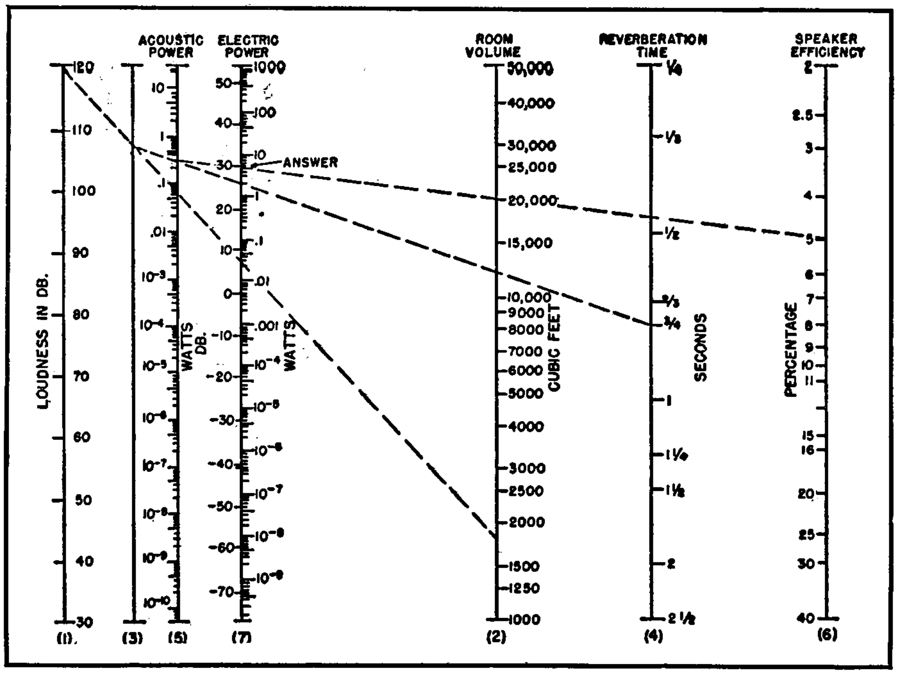


Fig. 23-15. Nomograph for calculating loudness and power requirements of sound systems.

to be very useful and even more educational. The limitations of the graph are the estimates involved in the reverberation time and the speaker efficiency. These factors must be used in any system of figuring power requirements plus an added safety factor which must be large enough to cover all contingencies. However, by the methods described it is usually possible to come within 50% of the actual power required.

It is amazing the tremendous range of power which an amplifier has to deliver. Peak power is needed only a very small fraction of the time. The average power required is only about a thousandth of the peak power. Amplifiers must have the needed reserve power for those times when musical peaks occur, power must be clean and low in distortion or else the thrill of musical volume is supplanted by a sound causing a wince and a shudder. A good audio system must be designed around the power requirements of the installation if it is to be free from overloading on the one hand or overpowered with resulting extra costs on the other.

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CHAPTER 28

Audio Measurements

Measurement of distortion, frequency, gain, pickup and cutter performance and the control of hum.

● The acoustical quality of an audio amplifier is related to the amount of distortion prevalent in the amplifier. If conforming to true Class A operation, the output plate current waveshape of the amplifier should duplicate the waveshape of the grid voltage input. Such not being the case, the amplifier has a certain percentage of harmonic distortion which, if excessive, deteriorates the audio quality and becomes annoying to the listener.

Types of Distortion

There are three types of distortion found in an amplifier; (1) *amplitude* distortion (2) *frequency* distortion, and (3) *phase shift*. In *amplitude* distortion, the fundamental (plus harmonics) are observed in the output. *Frequency* distortion is caused by the amplifier's inability to amplify all frequencies equally. *Phase shift* is present when the amplifier has different delays for all frequencies and when the amount of distortion increases as the tube is operated outside of the linear portion of the tube characteristic curve, as shown in Fig. 28-1.

In addition to harmonic distortion, there is *intermodulation* distortion in audio amplifiers. Both are caused by nonlinearity in the amplifier. Intermodulation results in the production of frequencies equal to the sums and differences of a low and high frequency (and harmonics).

Qualitative Distortion Test

The simplest test for distortion in a class A audio-frequency amplifier stage may be made, as shown in Fig. 28-2, by applying a signal voltage of proper level to the input and inspecting the circuit for one or all of the following abnormal conditions:

(a) Presence of dc grid current.

(b) Fluctuation of the dc plate current.

(c) Fluctuation of the dc cathode voltage, if the circuit employs cathode resistor bias.

Each of these indications generally occurs in a positive direction, and each will disappear upon removal of the signal. It must be borne in mind, however, that this method is purely rudimentary in nature and serves only to detect the presence of distortion. One or two of the indications may be absent, depending upon the main cause of the trouble.

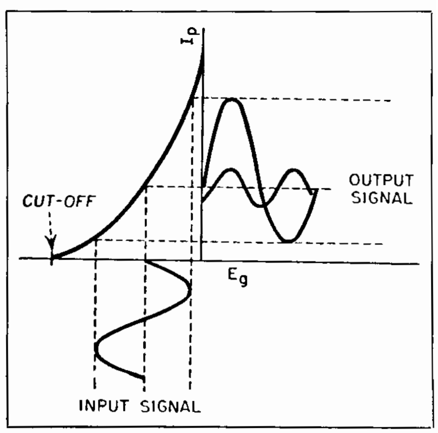


Fig. 28-1. Distortion in non-lineor amplifier.

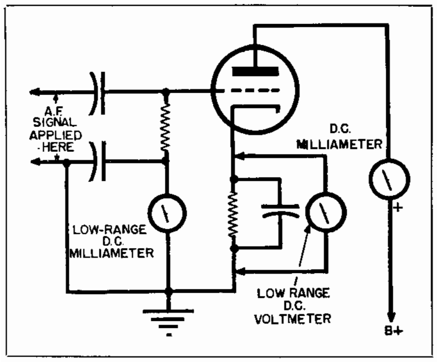


Fig. 28-2.

The three simple indications are well known and frequently used by servicemen and pa technicians having no equipment suitable for making quantitative distortion measurements, but must, in the course of routine testing, localize distortion without reference to the actual per-cent harmonic energy present.

The cathode circuit effects noted are due to fluctuations in the voltage drop across the cathode resistor, occasioned by variations in the dc component of plate current. The current indicated by the plate-circuit milliammeter is the average value of the fluctuating signal plate current, identical with the dc component, and is the current that produces the cathode resistor drop. These facts may be better comprehended when it is remembered that the fluctuating signal plate current (Fig. 28-3) is an alternating current, corresponding to the signal, superimposed upon a direct current. It will be evident from the fundamental relations of this combination that the average value of plate current, as indicated by the plate-circuit milliammeter, will be constant in the company of the alternating component under distortionless operating conditions.

Fig. 28-3 is a graphical representation of signal plate current. I_{max} is the maximum value reached by the fluctuating plate current; I_o, the zero-signal value; I_{min}, the minimum value. From these values, it may be shown that the per-cent second-harmonic content (often the most troublesome distortion factor) is equal to:

$$100\ \frac{I_{max}+I_{min}-2I_o}{2(I_{max}-I_{min})}$$

Quantitative methods of checking distortion are *harmonic analyses,* and are concerned with measurement of the actual amount of energy present in each separate harmonic of the signal frequency (or in the total harmonic content) and establishment of percentages with respect to the fundamental frequency. The most representative methods employed in wave analysis and the apparatus necessary thereto will be described presently.

The cathode-ray oscilloscope is notably useful in the observation of wave shapes. When the horizontal plates of the ray tube are energized by a saw-tooth wave sweep-oscillator-amplifier circuit to furnish the linear time base, and a signal voltage which it is desired to observe is applied to the vertical plates through a substantially flat-response amplifier, the cathode-ray trace will be an exact reproduction of the waveform of the applied signal voltage.

An audio-frequency amplifier may be checked for distortion with the oscilloscope in the manner illustrated in Fig. 28-4. At the left is an audio oscillator

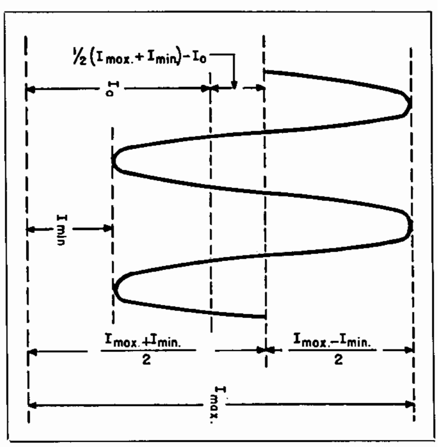


Fig. 28-3.

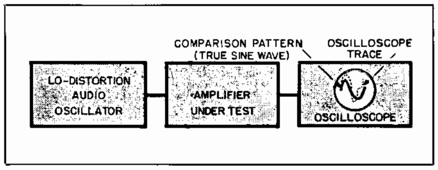


Fig. 28-4.

possessing an output voltage waveform of known purity, next is the amplifier under test, and at the right is an oscilloscope having horizontal and vertical amplifiers with substantially flat frequency responses.

It is the purpose of the oscillator to supply a signal of as pure waveform as practicable to the amplifier, and that of the oscilloscope to reproduce the wave shape of the signal after it has passed through the amplifier. In order that as little distortion as possible be introduced by the instruments themselves, the oscillator used for such a test must be of exceptionally high quality and the amplifiers in the oscilloscope must possess an excellent frequency characteristic. Likewise, the oscilloscope sweep circuit must be uncompromisingly linear in its characteristic.

If the amplifier had no distortion at all, the signal it delivered to the oscilloscope would be an exact reproduction of the input signal waveform. This is never encountered in practice, however, the most efficient amplifier arrangement being beset with the distortion characteristics of its tubes and other components.

For observations, a perfect sine wave (or, better still, a tracing of a single cycle from the test oscillator) might be inscribed on the transparent viewing screen of the oscilloscope, and signals from the amplifier matched to this pattern to discover variations from the original shape due to amplifier distortion. In making such a test, it would of course be necessary to adjust both oscilloscope amplifier gain controls in such a manner that the maximum amplitude and width of the signal trace coincided with those dimensions of the inscribed pattern.

With the low percentages encountered with most well-designed amplifying equipment, it will be difficult to estimate the percentage of harmonic content from the reproduced wave shape, in the oscilloscopic method, unless the operator makes use of the transparent screens furnished by some oscilloscope manufacturers for the purpose. These screens carry printed patterns of single cycles corresponding to the shapes obtained (variations from true sinusoidal) with various low percentages of distortion. Severe cases would result in images similar to Fig. 28-5 which is an exaggerated representation of pronounced third-harmonic content.

Distortion Measurements

This is probably the most basic measurement of performance in wide range audio equipment. It is only a short time ago that 5% total harmonic distortion was considered the criterion of high quality performance. In recent years this entire attitude has been recognized as naive, and an increasingly critical listening audience has demanded new standards of fidelity.[1]

Intermodulation distortion, another product of non-linearity, has received wide acceptance as a measurement that correlates particularly well with listening observations. Although a number of articles have appeared periodically in the literature, there is still considerable confusion in the minds of many audio engineers and technicians regarding the

[1]Goodell, J. D., "Audio Measurements." Radio-Electronic Engineering, Nov. 1948.

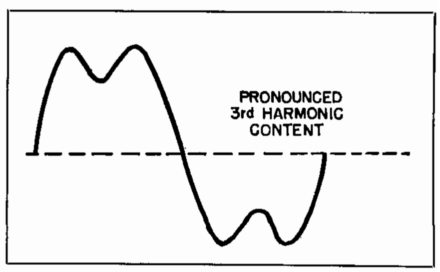


Fig. 28-5.

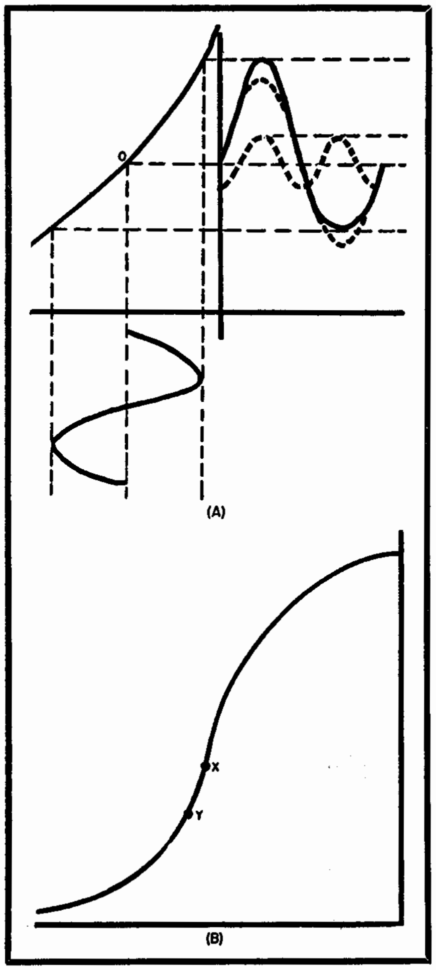


Fig. 28-6. (A) Impressing a frequency on a square law characteristic gives a distorted output consisting principally of the fundamental and second harmonic. (B) Curve is symmetrical about X, leading to odd harmonics. If operating point is moved to Y, both odd and even harmonics result.

real significance of intermodulation distortion and the application of intermodulation distortion analyzers in the industry.

Harmonic distortion and intermodulation distortion are intimately related in the sense that they stem from the same kind of non-linearity. Correlation between the two measurements is difficult for many reasons but may be accomplished on an approximate basis if the order of the harmonics is properly "weighted" in setting up the data. It is quite obvious that if harmonic distortion is eliminated, this, by definition, means elimination of non-linearity; there will be no intermodulation distortion. However, it is important to emphasize the fact that intermodulation is the more sensitive of the two measurements. When harmonic distortion has been reduced to a value that becomes difficult to observe with accuracy, appreciable percentages of intermodulation may still be measured with relative ease. Thus, the first and most obvious advantage of intermodulation distortion analysis becomes evident. It is more sensitive, and low percentages of non-linearity are reflected in larger percentages of intermodulation than harmonic distortion.

Harmonic distortion is relatively easy to represent and understand, but a few rarely mentioned phenomena are worthy of presentation. A simple case is where a frequency f_1 is applied to a device with a square law characteristic curve, as shown in Fig. 28-6A. The output resulting from lack of symmetry is obviously distorted and may be resolved into two essential components consisting of f_1 and $2f_1$. Where the characteristic curve departs from symmetry in a manner more complex than a simple square law, the terms in the output will contain additional multiples of f_1.

If a non-linear characteristic curve is symmetrical, the harmonic frequencies generated will be $3f$, $5f$, $7f$ . . . . ., consisting entirely of odd harmonics, a condition that is closely approached in push-pull amplifiers. Assymetry in a non-linear device leads to even harmonics. This, if the operating point X is chosen as an operating axis on a characteristic curve such as appears in Fig. 28-6B, so that the lower portion of the curve "mirrors" the upper portion, the output will contain only odd harmonics. In Fig. 28-6A the curve is assymetrical and even harmonics are produced. Now if the operating point in Fig. 28-6B is moved to Y, the system will produce both odd and even harmonics. This is an interesting aspect of self bias conditions where the static operating point may be selected symmetrically but may shift toward assymetry under dynamic drive. It also has appreciable bearing on the difference in output harmonic content between strictly Class A and Class AB_1 conditions.

Harmonic distortion is, by definition, the production of spurious frequencies that are mathematical multiples of the fundamental. The quality, or timbre, of a musical tone varies with the number and relative intensity of the harmonics. Under certain circumstances phase relationships are an added and important factor. At any rate, the addition of harmonic multiples of the fundamental do not introduce distortion that is objectionable on the basis of dissonance. The ear recognizes added harmonics or reenforced harmonics as a change in the quality of a tone.

When two fundamental tones are applied simultaneously to a non-linear element, the output waveform contains not only the harmonics of the fundamentals but also sum and difference frequencies of the fundamentals and harmonics, as well as combination tones made up from the harmonics only. If you apply f_1 and f_2, the output will contain f_1, f_2, $f_1 - f_2$, $f_1 + f_2$, $2f_1$, $2f_2$, $2f_1 - f_2$, $2f_2 - f_1 \ldots\ldots$, and the total structure becomes extremely complex. The resulting sum and difference tones rarely bear any harmonic relationship to the fundamental. These sounds have no relationship to the musical notes played by the musician, and they are distinctly dissonant. It is as though a violinist were to play *Mozart* with a faint accompaniment from a string quartette playing *Schonberg*. It is not strange then that listening tests have often indicated greater correlation with intermodulation measurements than with indications of total harmonic distortion. The intensity and number of intermodulation products produced increases not only with the magnitude of the harmonics but it is also a function of their order. It was early recognized that the presence of higher order harmonics was especially objectionable, and it has now become evident that the real reason for this is the increased contribution of violently dissonant combination tones.

These circumstances have been recognized for at least ten years. It is unfortunate that the wheels of progress and standardization move so slowly that total harmonic distortion is still the most commonly published criterion. This is partly true because standards are not firmly established, and partly because intermodulation measurements are often more revealing than some manufacturers find desirable. At the present time the publication of intermodulation distortion percentages has meaning only if the instruments used are specified together with the frequencies chosen and their relative amplitudes.

The application of audio test equipment falls into two broad categories. One is the design of new equipment and the other is production testing. Intermodulation analyzers are not only invaluable as designing tools but also have great value in quality control over production.

In establishing optimum values for a circuit design, it is simple to obtain empiric answers with the intermodulation analyzer. Decade boxes are inserted for the various components, and the effect of adjustments over a wide range of values may be rapidly observed in terms of intermodulation percentage.

Theoretically it may be said that if only two fixed frequencies are used for such a procedure, it is possible to design for a dip in the intermodulation reading that is peculiar to the frequencies used. Under these circumstances design parameters might be chosen that would not produce optimum results at other frequencies. However, in connection with practical design problems in audio amplifiers, this does not appear important if suitable frequencies are used.

It is obvious that considerations other than absolutely minimum distortion must govern the final choice of values. A simple example of this involves the cathode resistor for an input stage in a voltage amplifier. If the selection is based only on minimum distortion, then the value chosen will often not provide

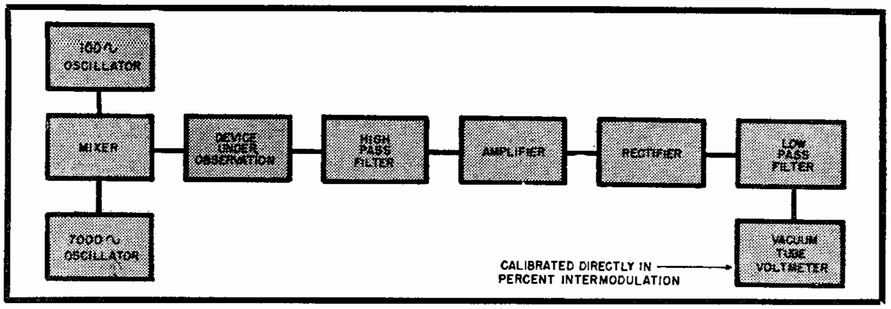


Fig. 28-7. Block diagram of a basic intermodulation meter.

sufficient bias to prevent the flow of grid current on peak signals.

For purposes of rigorous design, and particularly investigations of unconventional circuits, it is undoubtedly desirable to measure intermodulation distortion with a wide variety of applied frequencies. This may be accomplished with two oscillators and a wave analyzer. A great deal of interesting and valuable information may often be obtained by sliding two oscillators separated by only a few hundred cycles through the entire audio range and measuring the sum and difference frequencies generated in the equipment under test. This method is uniquely valuable in the high audio frequency ranges where the harmonics fall above the audio spectrum and above the range of the equipment under test so that harmonic distortion methods become impractical.

The basic form of an intermodulation distortion meter is shown in Fig. 28-7. Two frequencies are generated and mixed together. The most common frequencies used are 100 and 7000 cycles, and they are generally set in relative magnitude with a ratio of four to one, the low frequency being the largest. Obviously it is essential that the equipment under test be capable of passing the frequencies used with reasonably flat response. Where equipment is limited in range, a lower frequency than 7000 cps is used.

This complex waveform is fed to the input of the device under test, and the output of that device passes through a high pass filter. This eliminates the low frequency component together with approximately its first ten harmonics. The higher frequency may now be considered as a carrier that has been modulated. This waveform is applied to a rectifier and low pass filter, the remaining detected components being observed on a meter calibrated directly in intermodulation percentage.

A great deal of the early work on intermodulation distortion was accomplished in connection with sound motion pictures. The ASA "Standard Method of Making Intermodulation Tests on Variable-Density 16-Millimeter Sound Motion Picture Prints" approved March 19, 1946, is quoted in part as follows:

". . . Any distortion in the over-all process causes a change in high frequency amplitude in portions of the low frequency cycle. The ratio of the average variation in amplitude of the higher frequency in the reproduced wave to its original amplitude is called intermodulation. Intermodulation test results are not directly proportional to harmonic measurements, but in most cases an intermodulation figure of 10% corresponds to a harmonic reading of 2½%."

Norman Pickering reports that in his experience a reasonable rule of thumb method relates the intermodulation percentage to the sum of the percentages of each harmonic multiplied by the order of the harmonic. This estimate is based on using 100 cps and 7000 cps mixed 4 to 1. Thus, 3% second, 1.4% third, 0.5% fifth and 0.1% seventh would approximate 13.4% intermodulation. The total harmonic distortion indicated is

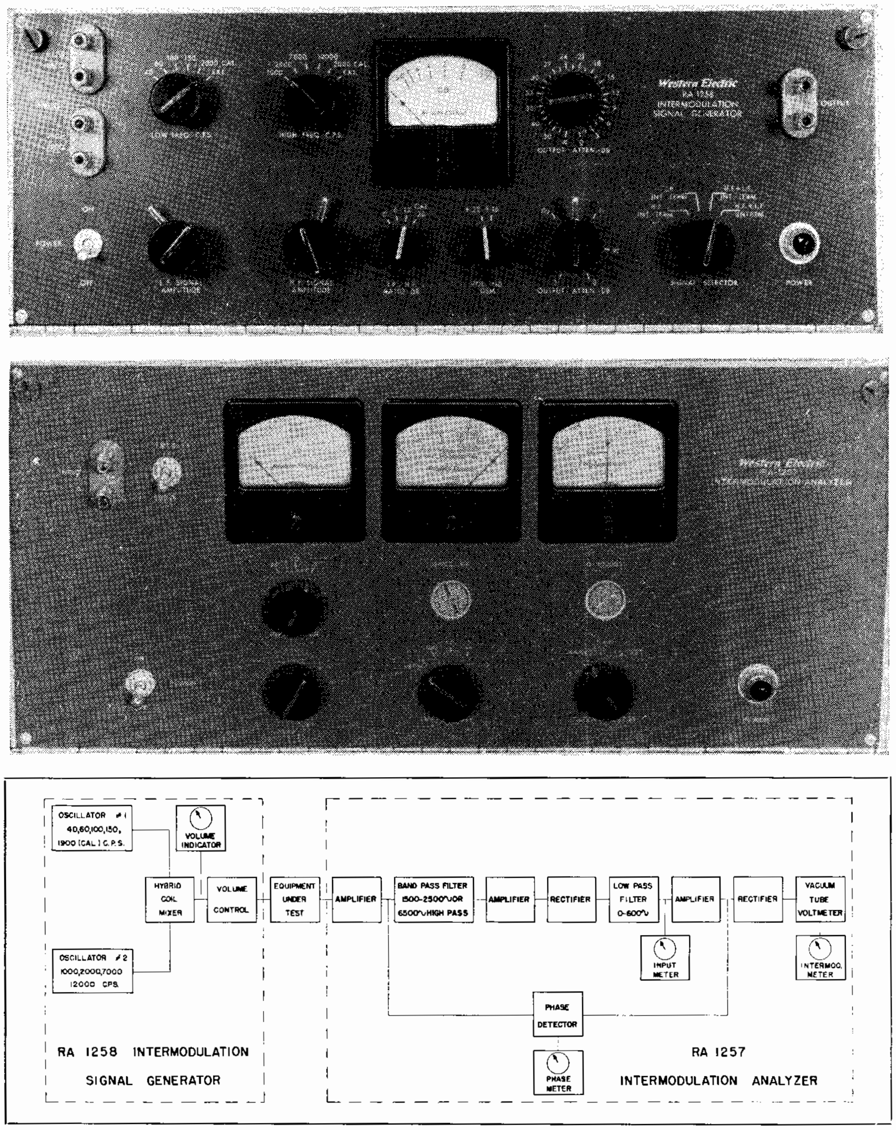


Fig. 28-8. Western Electric RA-1258 intermodulation signal generator (top), RA-1257 intermodulation analyzer (center), and block diagram of a complete setup for intermodulation measurements. (Courtesy Western Electric.)

5%, a figure that has been considered tolerable for high quality equipment. Yet experience in measurement and correlations with listening tests show that under the same conditions of measurement the intermodulation should be less than 5% for really high quality "clean" reproduction.

When used as a production test instrument for quality control, the intermodulation meter, Fig. 28-8, has many advantages. It is extremely simple to set up the equipment since it requires only an input and output connection. Power output may be standardized with a simple arrangement of resistive loads and a high quality vacuum tube voltmeter. Once a standard limit for intermodulation percentage has been established for a specific piece of equipment at a given power output, the measurement will rapidly show whether production units are satisfactory. The important point here is that excessive hum, a marked change in frequency response or any nonlinearity will change the intermodulation percentage reading. The meter will not tell the operator what is wrong with the equipment, but it will rapidly indicate that it is or is not satisfactory.

These methods of design and production testing are by no means limited in value to the manufacturer of extremely high quality equipment. It has been demonstrated that careful design with proper testing methods may make it possible to improve appreciably the performance of low priced table model radios. High quality results do not necessarily mean extremely high costs. Often careful analysis will actually show improvement through the elimination of components and simplification of circuits. The average listener will tolerate considerable modification of the original quality of tone but is very much disturbed over any extraneous added sounds. The reduction of intermodulation distortion is reflected in the cash registers of those manufacturers who have made this effort to improve their products.

It is not intended here to indicate that the design and production of high quality reproducers is strictly a matter of minimizing intermodulation distortion. Many forms of distortion and other factors are of importance. Transient response is greatly neglected, and noise reduction is certainly of paramount importance. But it is true that the customary standards involving only total harmonic distortion are entirely inadequate, and intermodulation measurements are of basic importance and value both in designing and production testing.

One approach to obtaining low values of distortion is particularly to be deplored. This concerns the all too common tendency to increase the power output unreasonably on the assumption that by so doing the distortion at power outputs actually used will be adequately low. In the first place this is fallacious, and secondly it is extremely uneconomical. Furthermore, it has led to a tendency on the part of many customers to have a greatly distorted concept of the power required for home installations. This is said with full recognition of the enormous ratio between average and peak power in music reproduction. However, 30 to 40 watts is hardly required in the average living room, and it would be very desirable if satisfactory standards of measurement were used for maximum allowable distortion at rated output power. This should either be expressed in intermodulation percentage with standardized methods of measurement, or if it must be in harmonic distortion, the percentage of each harmonic that can be reasonably measured should be included.

Intermodulation measurements have been extensively applied in studies of disc recording and considerable information obtained that would have been difficult to evaluate with other methods. Recently magnetic recording has been under intensive investigation and here, too, the intermodulation distortion analyzer has been of value.

Perhaps the most important single argument for intermodulation distortion measurements is that theoretically they should correlate well with listening tests, since they are a measure of distortion in a form that would be expected to be particularly objectionable to the listener. In practice, a number of tests run by independent investigators have revealed this as factual.

It is of considerable interest to note that the ear is capable of generating intermodulation products. A great deal of work has been done in studying this phenomenon and much of the information obtained is applicable, in general, to any non-linear system. For example, it is found that the different tones are usually of greater magnitude than the summation tones.

Newman, Stevens and Davis[2] using a wave analyzer were able to detect the presence of 66 different frequencies in the electrical response from the cochlea of a cat that had been stimulated with 700 and 1200 cps at 90 decibels above threshold. They point out that the large number of combination tones that may be present in the ear when the stimulus is limited to only two frequencies indicates that the complexity of the spectrum in the transmission system of the ear is incredible when the stimulus is orchestral.

Because of internal masking effects external distortion is less evident and less observable at very high levels of loudness than at low levels. This fact is often overlooked because of the common practice of turning up the volume to listen for distortion from a reproducing system. This is done because the tendency for any system to distort becomes greater as it is driven harder, but actually a given percentage of distortion may be detected most readily in listening observations at low levels. The reason for this is that the ear produces internal distortions that tend to mask distortion in the signal when the level is high. This applies to harmonic distortion and also to intermodulation, although there is somewhat less probability that the intermodulation distortion produced by the ear will correspond with external distortions in a manner that leads to masking. Again the tendency is for intermodulation to be more irritating.

It is often assumed that phase relationships are of negligible importance in producing objectionable distortion. That this is not true has been demonstrated. A harmonic in the signal may be reenforced or canceled by a harmonic or similar frequency generated in the structures of the ear. When two instruments play together, the harmonics from one may be in such phase relationship as to cancel harmonics in the other signal. Trimmer and Firestone[3] have shown that a change in the phase of a harmonic from a relatively low frequency fundamental may change the effective loudness of the harmonic and produce a distinct change in the timbre of the tone. The phase relationship that produces minimum stimulus also produces the greatest smoothness, while shifting the phase of the harmonic 180 degrees from this will provide maximum loudness and a definite roughness in quality.

Intermodulation distortion measurements are important and certainly such standards are more to be desired than existing total harmonic distortion representations.

Frequency Response Measurements

In the history of audio development work the characteristic used as a reference for the quality of audio equipment has changed continually. Thus, since one of the earliest problems was to obtain good low frequency response, there was a long period during which this was viewed as the most important individual factor to consider in evaluating a reproducing system. Later, when various improvements in loudspeakers and other components made reasonably good low frequency response practical, very high frequency response and complete reproduction of harmonic structures became of principle interest. Fairly recently there has been a strong swing toward emphasis on the desirability of reproducing the lowest octave down to 30 cycles per second. Realization of the problems inherent in wide range systems, where all forms of distortion become increasingly evident, has produced the concept that the last improvement to make in any system is to broaden the frequency range and that this should be done only after all other forms of distortion have

[2]Newman, E. B., Stevens, S. S. and Davis, H., "Factors in the Production of Aural Harmonics and Combination Tones." J. Acous. Soc. Amer., 1937, 9, 107-118.

[3]Trimmer J. D., and Firestone, F. A. "An Investigation of Subjective Tones by Means of the Steady Tone Phase Effect." J. Acous. Soc. Amer., 1937, 9, 23-29.

been largely eliminated. This, in a way, is the completion of a development cycle.

Any audio engineer who has been active in the industry for the last fifteen to twenty years has gone through the stage where everyone listened for bass reproduction, then the period during which a system that did not produce a good strong hiss or surface noise from even the best records was frowned upon, back to listening for the lowest pedal tones in organ records and, finally, to observing how successfully needle scratch could be eliminated. It is a long and interesting show with a continually changing criterion of performance. It has seemed as though the solution of one problem inevitably brings into focus another problem even more difficult to solve.

The most obvious method of making frequency response measurements is to apply a signal from an oscillator to the input of the amplifier and vary the input frequency while observing the output signal on a suitable voltmeter or on an oscilloscope. For observations calibrated directly in decibels a high quality vacuum tube voltmeter is probably the most desirable output indicator. For many reasons it is often more desirable and convenient to use an oscilloscope. One obvious reason is that a satisfactory oscilloscope is more likely to be available than a vacuum tube voltmeter that is sufficiently dependable with regard to maintenance of calibrations and with regard to frequency response inherent in the instrument.

When an oscilloscope is used the simplest method is to sweep the oscillator slowly through the frequency range under investigation and observe the relative amplitude at various frequencies. When the measurement involves tone controls and filters, the oscilloscope is particularly desirable because the waveform of the signal may be observed simultaneously.

Sweep Frequency Techniques

Various devices have been made available for observing the entire audio response range instantaneously on the oscilloscope. Among these are the sweep frequency records, Fig. 28-9A, and sweep frequency oscillators made by several manufacturers. In the case of at least one of the latter instruments, it is easily possible to establish calibrations in terms of decibel scales and the frequency range is presented logarithmically. All of these methods are valuable for various purposes. The sweep frequency records have their principal advantage in the fact that the output of the pickup cartridge (Fig. 28-10A and B) is included in the measurement, but on the other hand this is a disadvantage for observing the characteristics of an amplifier unless the pickup is almost perfect because it introduces a variable that must be allowed for and in any event may not be entirely dependable. However, it is the least expensive method of obtaining a sweep frequency source and presenting the effective response curve (as in Fig. 28-9B) on the oscilloscope. For purposes of convenience in making rapid comparisons and adjustments this method is to be highly recommended.

Fig. 28-9A. Clarkstan 102M microgroove audio sweep frequency record.

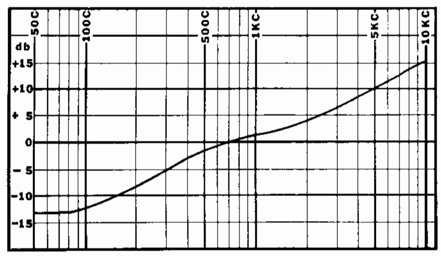


Fig. 28-9B. Frequency response of the 102M sweep frequency record.

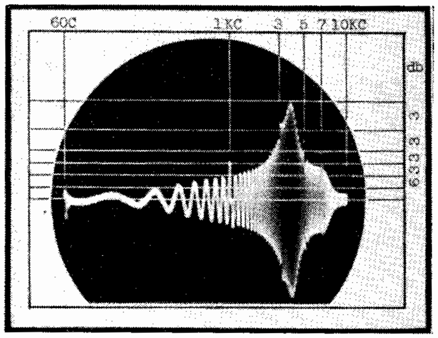


Fig. 28-10A. Unretouched oscillograph of the response of the No. 253 offset stylus when used with the RV 20 1 pickup. No equalization. (This oscillogram was taken using the Clarkstan No. 1000A Audio Sweep Frequency Record.)

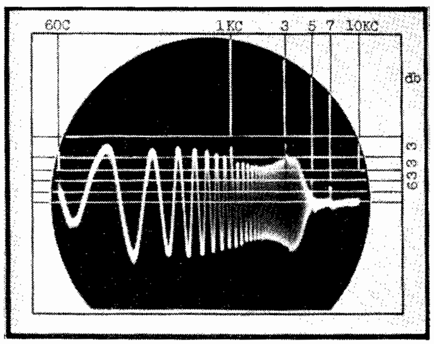


Fig. 28-10B. The oscilloscopic response of the No. 253 offset stylus and No. 201 pickup with complete equalization. (This oscillogram was taken using the Clarkstan No. 1000A Audio Sweep Frequency Record.)

Although the sweep frequency records and some of the sweep frequency oscillators provide marker pips for determining the specific location of various frequencies as observed on the oscilloscope, the accuracy with which this may be determined visually leaves much to be desired. In one of the instruments available sharp filters are provided with selective arrangements that make it possible to find a specific frequency on the curve quite accurately.

An effective and simple manner of observation of frequency response characteristics is very much to be desired because it facilitates design work, and speeds up production testing, contributing to lower costs and superior results.

There is one method of observing frequency characteristics that is believed not to have been previously described. This method has certain advantages over others in simplicity of observation and the accuracy with which the observation may be made, eliminating the possibility of error that may occur because the input signal is not ordinarily observed simultaneously with the output signal.

The method of frequency response observation to be described depends to some degree for its advantages on the fact that over the entire frequency response range there will almost inevitably be a certain amount of observable phase shift, even in a single stage of voltage amplification.

The effects involved and the patterns observed are by no means new, but the interpretation is applied to frequency response observations instead of the customary frequency determination or relative phase measurements.

Phase Shift Technique

The procedure is to apply, simultaneously, the output of an oscillator to the horizontal axis of an oscilloscope and the input of the amplifier under test.[8] The output of the amplifier is then applied to the vertical axis of the oscilloscope. If the input and output levels of the signal as well as the input controls to the oscilloscope are properly adjusted, one of the familiar *Lissajous* patterns shown in Fig. 28-11 will be observed. If at the frequency used there is no relative phase shift between the amplifier input and output, the pattern observed will be a straight line. At some frequency in the audio range this will be found to be the case, although the frequency for zero relative phase shift may be anywhere in the range. Having found a frequency at which this is the case, the amplitude of the two inputs to the scope should be adjusted to provide equal signal strengths to the plates so that the angle of this line

[8]Goodell, J. D., "Audio Measurements," Radio-Electronic Engineering, Jan., 1949.

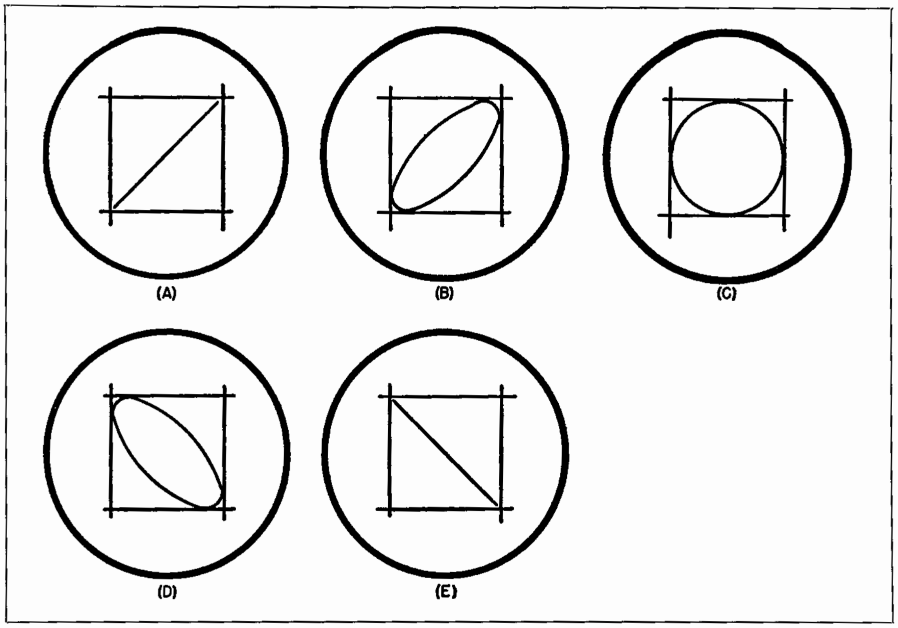


Fig. 28-11. Phase shift patterns used in frequency response observations described in text.

will represent the diagonal of a square, as shown in Fig. 28-11A or 28-11E.

Now, if the frequency is varied above and below this zero relative phase shift point, it will be observed that the shift in phase through the amplifier circuits will progress smoothly (in most amplifiers) so that the pattern gradually opens up. If the zero relative phase shift frequency is in the range above approximately 400 cps, this opening up will usually occur above and below the selected frequency. If the frequency response of the amplifier is perfectly flat, the pattern will progress through an ellipse to a perfect circle. If the perfect circle condition is reached before coming to the limits of the section of the spectrum under observation, it will then progress to an elliptical pattern again and back to a straight line representing the opposite diagonal from the one at which it started. This is shown in Fig. 28-11A through 28-11E.

In other words, if the frequency response of both the oscillator and the amplifier is perfectly flat, then there will always be a point on the pattern that is in contact with each side of the square (of which the initial pattern represents a diagonal). If the amplifier response falls off, then the vertical height of the pattern will decrease proportionately and the points that would normally be in contact with the top and bottom sides will recede toward the center. If the oscillator response falls off but the amplifier is flat, then both the vertical and horizontal dimensions of the pattern will decrease proportionately and all points that would otherwise be in contact with the sides of the square will recede toward the center.

Conversely, if the frequency response of the amplifier has a rising characteristic then the pattern will grow in a vertical direction, and if the oscillator has a rising characteristic the pattern will expand both vertically and horizontally. If the fault is with the oscillator, the axis of the elliptical patterns will not rotate. If the amplifier characteristic rises, the axis will rotate toward the vertical; if the characteristic tends to fall off, the axis will rotate toward the horizontal, as shown in Fig. 28-12.

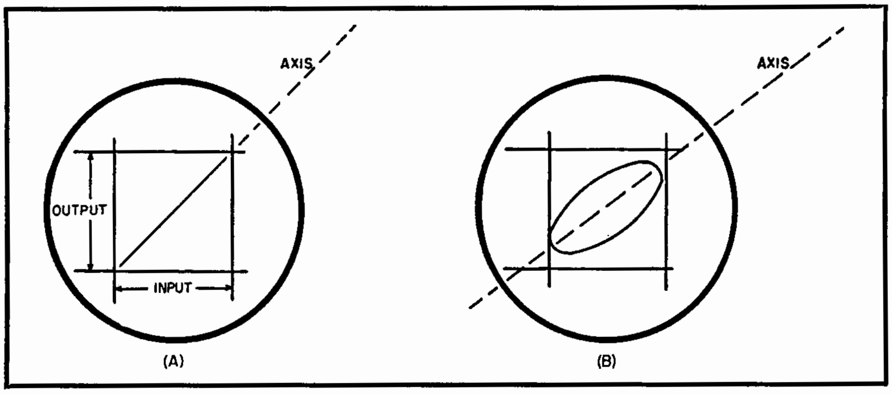


Fig. 28-12. Phase shift pattern—output vs. input. In (B) input is constant and output is attenuated. Note that axis rotates towards the horizontal.

The obvious advantage of this method is the ability to observe the oscillator response simultaneously with the response of the amplifier under test and immediately to ascribe a fault to the input signal or to the amplifier. When an effort is being made to obtain very accurate response characteristics, and particularly where this is being done in a production test set-up, it is well worthwhile to rule out (or in) the characteristics of the oscillator. Most oscillators are subject to an appreciable degree of change in their characteristics over a period of time. Rarely is all the test equipment used in production setups or on designing benches checked as often as it should be to insure protection against incorrect observations and conclusions. A system that eliminates this source of confusion has definite advantages.

With this method it is not possible to observe the waveform with any degree of accuracy, although serious distortion will show up in the lack of symmetry in the pattern. A flat-topped waveform from the amplifier will appear somewhat as shown in Fig. 28-13A. Various types of distortion will produce other patterns. (Fig. 28-13.)

In final production testing of amplifiers it is sometimes important to adjust the tone controls so that flat response will be obtained when the indicators point straight up rather than in the mechanical center of rotation where the controls are continuously variable. This may be accomplished quickly with the phase shift patterns. Having once established the frequency for zero effective phase shift in a particular amplifier design, the tone controls may be adjusted

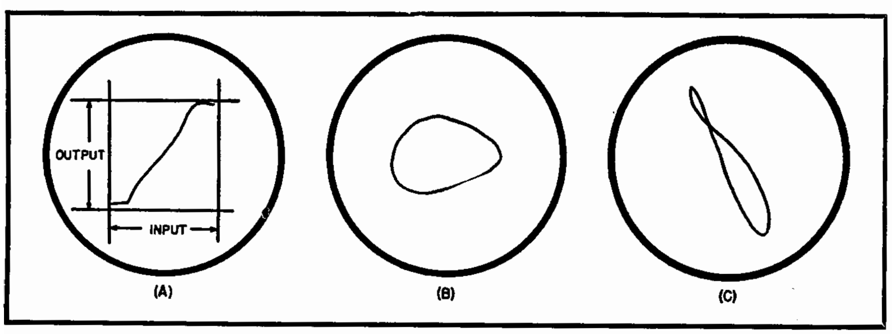


Fig. 28-13. Phase shift patterns—output vs. input. (A) Flat top distortion in output. (B) and (C) Various forms of non-linear distortion.

rapidly in production testing very close to "flat." This is done by setting the reference frequency on the oscillator and adjusting the controls until the pattern is a straight line. Any deviation from center setting will result in an opening up of the pattern. This may be accomplished by rotating the mechanical fastening of the potentiometers or by rotating the setting of the knobs if the shafts are not designed with a fixed flat contact point for the adjusting screw. A quick slide through the frequency range will then make a fine adjustment possible with speed and accuracy.

If it is found impossible to adjust the tone controls to achieve a straight line at the frequency normally correct for a specific amplifier design, it is a quick indication that the inherent response of the amplifier is faulty. It will be noted that any adjustment of the tone controls from the flat position, whether it be in a boosting or attenuating direction, will result in shifting the frequency at which the straight line pattern is obtained. High frequency attenuation, for example, will tend to move this frequency in a downward direction while increasing the high frequency response will move it in an upward direction in any specific amplifier.

In observing the phase shift across a single stage of voltage amplification and making the measurement directly between the grid and plate, it will usually be found that the frequency for zero effective phase shift is close to zero frequency. If the stage is reasonably flat in frequency response, the pattern will open very slowly and will not become a complete circle even at the limits of the audio range around 20,000 cps.

Phase shift and frequency response are intimately related and phase shift is an indication of attenuation. This fact may be applied with validity only to a simple system such as a single stage of voltage amplification. With complete amplifiers the end result in terms of frequency is produced by the cumulative characteristics of all the elements in the system. One stage may have a drooping characteristic while a subsequent stage rises in an almost perfectly complementary manner. The result may be an almost perfect over-all frequency response; yet a phase shift of 0.014° per cycle of frequency would be typical between approximately 700 and 14,000 cps. It is not impossible to design a single or even a two stage amplifier with practically zero attenuation, and thus zero phase shift. It is rarely encountered in audio circuits. Phase shift in audio work with respect to listening observations is in much the same category as Vitamin E with respect to human nutrition. Its importance has not been firmly established. However, considerable work has been done to determine its value and there is considerable accumulated evidence of a practical as well as a theoretical nature to show that it should not be completely overlooked.

If phase shift is linear in any system with respect to frequency expressed in degrees per cycle of frequency, there will be no waveform distortion. The reason, of course, is that such a condition represents a constant time delay for all frequencies. As a simple example with arbitrary and not necessarily practical figures, a specific system might be considered that has a phase shift of 0.18 degrees per cycle of frequency. Thus, 500 cycles per second would be shifted 90 degrees; 1000 cycles, 180 degrees and 2000 cycles, 360 degrees. If all of these frequencies were considered to have started out from the point of origin in time in a positive direction from the reference line, then after the phase shift took place their relationship with respect to time would not have changed. This is indicated in Fig. 28-14. The resultant complex wave form would be unchanged. No phase distortion would be said to have taken place.

If, on the other hand, there is a definite time delay resulting from nonlinear relationships between phase shift and frequency, the shape of the waveform will be changed. With conditions where the signal passes through a large number of amplifiers, each with an appreciable phase distortion characteristic, the results are obviously undesirable.

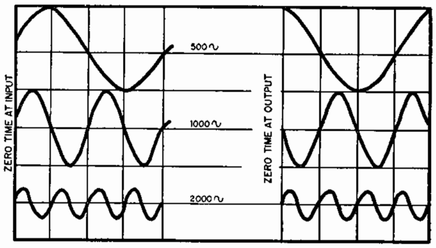


Fig. 28-14. Input and output phase relationships with theoretical linear phase shift of 0.18° per cycle. Note that time delay is constant at all frequencies—hence resultant waveform will not be changed.

With the amount of phase distortion ordinarily encountered in reasonably high quality audio amplifiers, the importance of the effect upon reproduced sound has not been definitely determined. The statement that phase shift is totally unimportant is based on the assumption that the ear rigorously follows Ohm's acoustical law, that it acts as a perfect wave analyzer and has no appreciable inherent distortion. Such, of course, is far from the truth.

It is obvious that if two signals of the same frequency are present, the resultant amplitude will be a function of their respective amplitudes and also of their phase relationship. The effect of feeding the same signal to two loudspeakers placed side by side but connected in opposite phase where the result of even a strong signal is almost complete silence is quite common. The same effect is observable in certain locations in rooms with public address systems where cancellation takes place, and also in many theatres. The principal reason for using an enclosure around a loudspeaker is to prevent the back wave from canceling the front wave in regions where the two may be exactly out of phase. Clearly this same principle applied for two frequencies in a complex signal and their phase relationship will strongly affect the observed resultant.

The same theory applies to many conditions where a difference in phase between frequencies that are not exactly the same will affect the resultant amplitude. Finally, the ear generates harmonics in its own non-linear structures. If a complex waveform consisting of a fundamental frequency and a number of harmonics is applied to the ear, the ear will generate harmonic structures from the fundamental and from each of the harmonics, provided that the signal level is adequate. If the harmonics in the exterior signal are in phase with the aural harmonics, the total harmonic structure observed by the central nervous system will be greater in amplitude than if the opposite condition exists. This has been demonstrated experimentally many times.

No one will debate the fact that in listening to live music or reproduced music the location of the observer in the room will affect the signal he hears to a very great extent, and that this is at least to some degree a result of phase shifts from direct and reflected waveforms.

If all this is true, it does not seem reasonable that phase shifts in a recording and reproducing system should be completely neglected. In degenerative feedback amplifier circuits it is necessary to maintain a minimum phase shift over extremely wide frequency response ranges where large amounts of feedback (such as 30 db.) are used. This may be a second order factor in the desirable listening results achieved with such designs.

In connection with the phase shift described, it is worthwhile to mention the simple method of approximating the

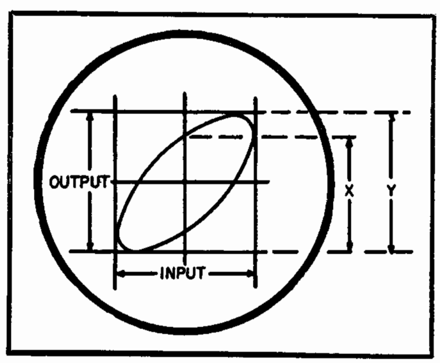


Fig. 28-15. Simple method of determining approximate phase shift by measurement of pattern—$\sin\phi = X/Y$.

amount of phase shift by physical measurement of the patterns. This is indicated in Fig. 28-15 where the sine of the angle is equal to the ratio of the dimension X to the dimension Y. In any form of these patterns X is the distance from the center to the point at which the pattern crosses the vertical line, and Y is the distance from the horizontal reference to the greatest vertical amplitude.

Frequency measurements, of course, are important in connection with all the components of a reproducing system. When response curves are constructed on the basis of sound pressure measurements for loudspeakers, they are usually so far from linear that the technician accustomed to observing amplifier response characteristics is often amazed that they represent high quality units. It is fortunate that the industry is progressing toward standardization of methods for making response curves on loudspeakers and that an understanding of their interpretation on a comparison basis is becoming more widespread.

Frequency records are often used to attempt to observe with listening tests the over-all response of a system from pickup to the ear. In practice there are many dangers inherent in this method, and conclusions drawn from the results will not always correspond to a judgment made in terms of straightforward music reproduction.

[4]LeBel, C. J. "Psycho-Acoustical Aspects of Listener Preference Tests," Audio Engineering, August, 1947.

[5]Roys, H. E., "Intermodulation Distortion Analysis as Applied to Disk Recording and Reproducing Equipment." Proc. I.R.E., October, 1947.

[6]Hilliard, J. K. "Distortion Tests by the Intermodulation Method," Proc. I.R.E., December, 1941.

[7]Frayne, J. G., and Scoville, R. R., "Analysis and Measurements of Distortion in Variable-density Recording." Jour. Soc. Mot. Pict. Eng., June, 1939.

Frequency Bridges

Certain bridge circuits, notably the Wien bridge (see Fig. 28-16) can be used for the identification of frequencies in the audio-frequency spectrum. If an alternating voltage is delivered to the bridge circuit, the latter may be adjusted for a null at that particular signal frequency. The null point would not hold for the same voltage of another frequency. Thus the adjustable element of the bridge might be calibrated to read directly in cycles-per-second.

The Wien bridge in its most useful form for this purpose would have its constants so chosen that the ratio arm R_2 is twice the ohmic value of R_1, the condensers C_1 and C_2 are equal in capacitance, and the two simultaneously adjustable resistance legs R_3 and R_4, are at all positions equal. Under these conditions, the frequency of the impressed voltage at null would be equal to:

$$f = 1/2\pi RC$$

Where:

f is the frequency in cycles-per-second,
R is the resistance of R_3 or R_4 in ohms,
C is the capacitance of C_1 or C_2 in farads.

Since the bridge may be balanced for only one frequency at a time, it would appear that any residual voltage indicated by the vacuum-tube voltmeter M, at null would be due to some other frequency or frequencies (such as harmonics of the fundamental). This harmonic voltage would be due to the total of harmonic voltages present. As such, the bridge might be connected, as shown in Fig. 28-16, to the output circuit of an audio-frequency amplifier which is passing a signal from a high-quality audio oscillator.

While the device might be used as shown as such a harmonic totalizer, the percentage total harmonic content with respect to the readings of the meter before and after null would not be reliable, nor would its error be uniform for all frequencies. These facts are due to the peculiar inability of the bridge to attenuate various harmonics equally.

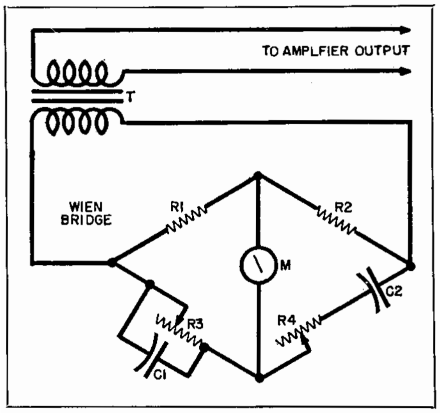


Fig. 28-16.

Another popular type of bridge harmonic totalizer is shown in Fig. 28-17. Here, three legs of the bridge, R_2, R_3, and R_4, contain pure resistance, while the fourth leg contains the shielded parallel resonant circuit, *LC*, which is resonant at the test frequency. The transformer *T*, like the one shown in the bridge previously described, must have an excellent frequency characteristic.

At resonant frequency of *LC*, the inductive reactance of the tuned circuit equals the capacitive reactance, the former is cancelled by the latter, and the bridge balances as if all four legs were pure resistance. Any voltage applied by the circuit to the vacuum-tube voltmeter is then due to harmonics of the test frequency (and it is assumed that these harmonics have been delivered to the bridge by the amplifier under measurement).

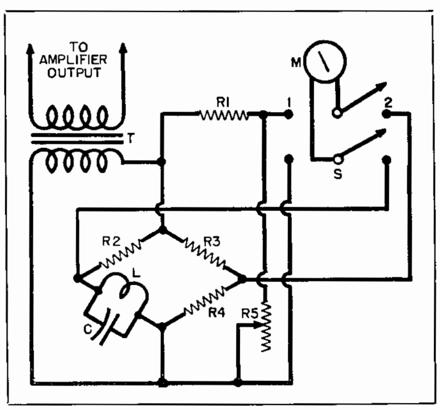


Fig. 28-17.

In operation, the double-pole, double-throw switch, *S*, is thrown to position 2 and the bridge balanced with the assistance of the vacuum-tube voltmeter, *M*, as a null indicator. The reading at null (due to harmonics) is recorded. The switch is then thrown to position 1 and R_5 is adjusted until the meter gives the same reading (as before at null). The following calculation may be performed to determine the percent of total harmonics from this operation:

$$\%H = 100R_5/(R_4 + R_5)$$

A dial indicator attached to the potentiometer R_5 may be calibrated directly in these percentages.

Square Wave Testing

Checking the response of an amplifier to a square-wave signal provides one of the fastest and easiest methods for testing frequency response, phase shift, transient response and similar characteristics.[9]

A close approach to a square wave may be obtained by "clipping" the peaks of a sine-wave signal as illustrated in Fig. 28-18. A number of clipper circuits

[9]Garner, L. E., "Wide Frequency Range Square-Wave Clipper," Radio & Television News, Mar. 1950.

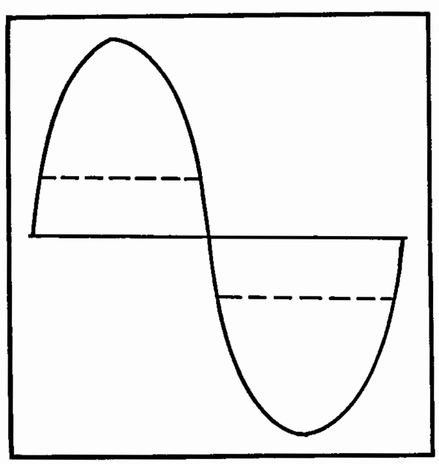


Fig. 28-18. Principle of clipper circuit. Peaks of sine wave are clipped to get square wave.

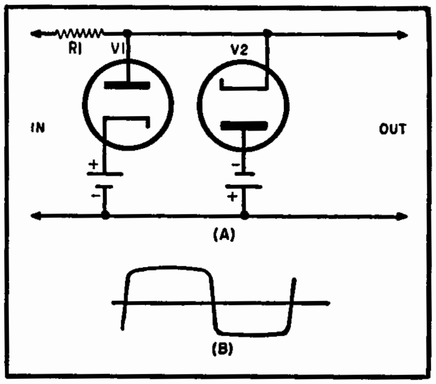


Fig. 28-19. Diagram of square-wave clipper. Its waveshape is not as sharp as that of the final circuit shown in Fig. 28-20.

may be used to do this, the most popular being illustrated in Fig. 28-19A.

In operation, whenever the voltage of the input sine-wave signal exceeds the dc voltage applied to the diodes by means of small cells, the diodes conduct. Diode V_1 conducts on positive peaks and diode V_2 conducts on negative peaks. When either diode conducts, it acts as a short circuit and the input signal is dropped across series resistor R_1.

The effectiveness of clipping in this manner depends on the combined value of the diode resistance (when conducting) and the battery resistance in comparison with the value of R_1. If R_1 is very large compared to the combined diode and battery resistance, reasonably good clipping is obtained.

This circuit, though widely used, does not give a really close approach to a "perfect" square wave, and hence is not suitable where more exacting tests are required. First, regardless of how large R_1 is made, the diode and battery still have some resistance and a small voltage will appear across them. This voltage will vary with the changing resistance of the diode. Thus, a "rounded" square wave is obtained, with both trailing and leading edges rounded somewhat and with a slight bow instead of a perfectly flat top, as illustrated in Fig. 28-19B.

In addition to this disadvantage, the circuit also has a limited frequency range, for as R_1 is made larger, distributed capacities become important and limit the frequency at which even a close approach to a square wave can be obtained.

By using a different arrangement, the clipper circuit shown in Fig. 28-20A can be obtained. This circuit, when properly driven, will provide almost perfect square waves, with sharp corners and a flat top, over an extremely wide frequency range.

In operation, diodes V_1 and V_2 are normally conducting and thus act as resistors, passing any signal applied to the input. However, when the peak of the input signal exceeds the battery voltage, then one of the diodes stops conducting and acts as an open circuit, preventing further passage of the signal and effectively clipping the peaks.

On negative peaks, the plate of V_1 is made negative with respect to its cathode and hence it stops conducting and acts to open the circuit. On positive peaks, diode V_1 continues to conduct, but the cathode of V_2 is made positive with respect to its plate and this tube acts to open the circuit.

Application

When testing either a single stage or a complete amplifier, the equipment is arranged as shown in block diagram form in Fig. 28-22. A good oscilloscope and a sine-wave signal source are re-

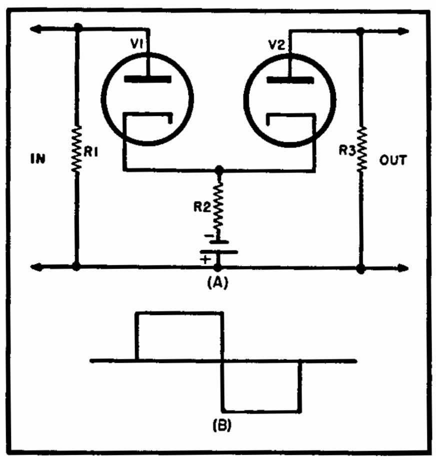


Fig. 28-20. Clipper circuit. See text for values of components.

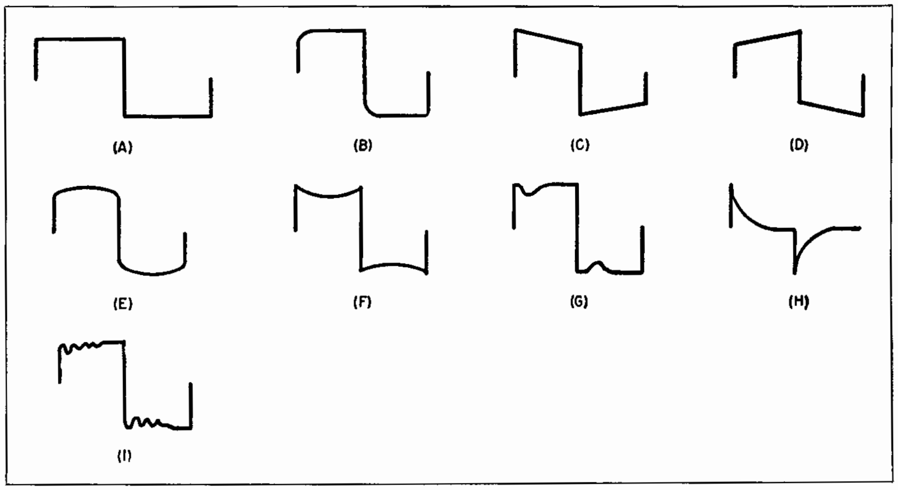


Fig. 28-21. In square-wave testing the output waveshape is noted and compared with the square-wave input. If any deviation occurs, it indicates some form of trouble in the unit under test. Each of the above waveshapes indicates a particular type of defect. See text for a complete analysis of the circuit faults represented.

quired in addition to the clipper. The square-wave signal at the output of the clipper is first observed on the oscilloscope. Next, the output signal from the amplifier is observed and any departures from a perfect square wave noted.

It is best to adjust the horizontal sweep of the oscilloscope so that at least two complete cycles can be observed on the screen.

An input square wave and distorted square waves showing the effect of different amplifier characteristics are shown in Fig. 28-21. The perfect input square wave is shown in Fig. 28-21A.

A drop-off in high frequency response in the amplifier causes "rounding" of the leading edges of the square-wave signal as shown in Fig. 28-21B. Usually, the rounding off will be easily noticeable if there is a decided drop in amplifier gain

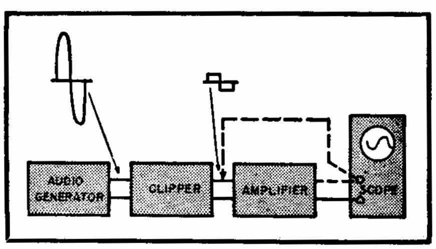


Fig. 28-22. Test setup used for checking audio amplifiers.

by the tenth harmonic (or less) of the square-wave fundamental frequency. Thus, if a 1000 cps square wave is passed without appreciable rounding, you can be reasonably sure that the amplifier is "flat" to 10 kc. But this gives no indication of the response below the fundamental frequency of the square wave. To do this, a lower frequency square wave is required.

Since the clipper, when properly driven, can easily supply a 20 kc. square wave, it can be used for checking the response of wide-band amplifiers (up to 200,000 cps) as well as audio amplifiers.

If there is phase shift in the amplifier so that phase leads at low frequencies, the top of the square wave is tilted as shown in Fig. 28-21C. If phase lags, the tilt is as shown in Fig. 28-21D. The amount of "tilt" depends on the degree of phase shift.

The effect of accentuated gain at low frequencies is shown in Fig. 28-21E, while the effect of a drop in gain is shown in Fig. 28-21F. The drop in gain is at the fundamental frequency of the square wave. It is assumed that there is no phase shift in both cases.

Low frequency phase shift and response tests are usually made with a 20 to 60 cps square wave, the exact frequency depending on the low frequency response of the amplifier being checked.

A dip at one point in the square wave, as illustrated in Fig. 28-21G may be caused by a drop in amplifier gain over a narrow range of frequencies (or at one frequency). If the drop in gain occurs at the square-wave frequency, then the dip spreads over the entire half cycle and we get the condition of Fig. 28-21F.

Too low a value of coupling condenser, too small a value grid resistor, or a partially open coupling condenser may cause differentiation of the square wave, resulting in a pulsed output signal as shown in Fig. 28-21H.

The transient response of the amplifier may be checked by noting if there is any overshoot or damped oscillations following the leading edge of a high frequency square wave as shown in Fig. 28-21I. A damped oscillation of this type may be caused by distributed capacities and lead inductances resonating at a low frequency, causing a sharp rise in amplifier gain at that point. This condition may also be caused by an undamped peaking coil in a video or scope amplifier.

The frequency at which the circuit resonates (and at which the peak in response occurs) can be determined by spreading the observed signal on the oscilloscope screen until the individual "cycles" in the damped oscillation can be counted. The number of individual cycles, multiplied by square-wave fundamental frequency, gives the approximate frequency at which the peak occurs. Although the value determined in this manner is not absolutely accurate, it is sufficient for all practical work.

In general, the low frequency characteristics of an amplifier are checked by applying a square-wave signal with a frequency near the lower limit of the amplifier. If the flat top of the square wave is tilted, phase shift occurs. If the leading edge is gradually rounded, there is a gradual falling off in amplifier gain at higher frequencies. If there is a peak or a dip in the signal there is either a peak or a drop (respectively) in amplifier gain at some particular frequency. The frequency at which the peak or drop occurs can be determined approximately by the ratio of the time of the peak or dip with respect to the time for the complete cycle of the square wave.

The high frequency response characteristics of an amplifier are checked in the same manner. For high frequencies, however, in addition to the above mentioned characteristics, transient response can also be checked.

For pa amplifiers, square waves at frequencies of 60 cps and 1000 cps are normally sufficient. For high fidelity audio amplifiers, frequencies of 20 cps, 200 cps and 1500 cps should be employed. Finally, for wide-band amplifiers, additional square waves with frequencies about a decade apart should be used, the highest frequency being about one-tenth the upper frequency limit of the amplifier.

In all cases, however, make sure the scope has a flat enough response to enable you to observe a square wave at the frequency used.

Routine Tests for Audio Amplifiers

Sound systems have found such wide application in recent years that the maintenance and repair of this equipment has become a distinct occupation.

Modern amplifiers approximate radio receivers in the number of their circuits and the increasing difficulty with which faulty operation is diagnosed and localized in them. Since in order to be profitable an amplifier test must be performed as quickly and completely as possible and the diagnosis must be precise, the test must be made according to a well-organized, time-saving plan.

A procedure for the routine inspection of audio amplifiers in maintenance and trouble shooting is as follows: Some of the methods included have been borrowed from laboratory practice, others are common to radio servicing. The steps given below have been selected to disclose most amplifier troubles. The arrangement of tests and their sequence are believed logical for complete diagnosis of amplifier performance, whether trouble is present or not.

By sufficient rehearsal of the operations, the Audio Technician should acquire considerable dexterity in the speedy analysis of amplifier operation. The data he collects from the series of tests should enable him to recondition an amplifier completely or appraise its performance.

Following are the recommended tests in the order in which they should be made:

1. Tube Checking
2. AF Signal Tracing
3. Static Voltage and Current Measurements
4. Gain Measurement
5. Frequency Response Check
6. Distortion Check
7. Check for Feedback
8. Impedance Measurement
9. Power Output Measurement
10. Hum and Noise Level Checks

The tube check (1) is preferably one which gives an indication of some dynamic characteristic (such as transconductance or *amplification* factor), rather than emission. AF signal tracing (2) consists of following the progress of an audio signal through the various amplifier stages, noting amplification, reduction or distortion of the signal in the stages. The static voltages and currents (3) are measured at appropriate circuit points, generally directly at the tube socket terminals. A rectifier-type ac voltmeter is satisfactory for indicating the heater voltages, but an *electronic dc voltmeter* (VTVM) or potentiometer-type indicator is mandatory for plate and screen voltages. Most of the high-resistance non-electronic meters are totally useless for measurements in resistance-coupled amplifier circuits. The operations numbered from 4 to 10 in the list will now be discussed in detail.

Gain

Gain measurements reveal the actual amplification obtained in the entire amplifier or in any of its stages. Gain measurements are quantitative. Both *voltage gain* and *power gain* are measurable and are of importance to sound engineers. However, voltage gain is of commonest concern and is generally implied by use of the single word, *gain.*

Both voltage and power gain are determined by the increase in output signal over the input signal. In procedure, the ratio of output voltage or power to input voltage or power is determined by measurement. There is no name unit for gain, this characteristic being expressed simply as the quotient indicated by the voltage or power ratio. For example, a gain of 10 (or 10 times) corresponds to the ratio 10 : 1, and indicates the amount of amplification in *any* case where the output voltage or power is ten times the input value. Since this is true regardless of the actual magnitudes of the input and output signals, gain is seen to be a relative characteristic.

To determine the overall voltage gain of an amplifier, an af voltage is applied to the input terminals of the amplifier and successive readings taken of the voltage impressed across the input circuit and the output voltage developed across a terminating resistance or impedance. Voltage measurements are made with a vacuum-tube voltmeter in a circuit similar to those of Figs. 28-23 and 28-24. The ratio of the two voltages is the overall gain. However, most amplifier manufacturers state overall gain as the number of decibels corresponding to this ratio.

It is advisable to make overall gain measurements with all volume controls set at maximum volume and to adjust

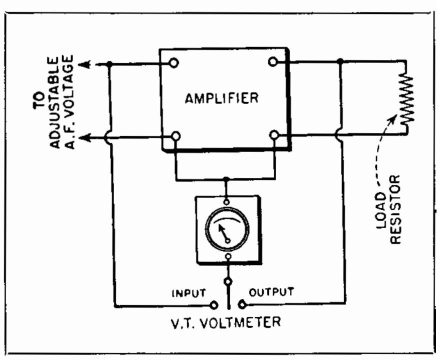


Fig. 28-23.

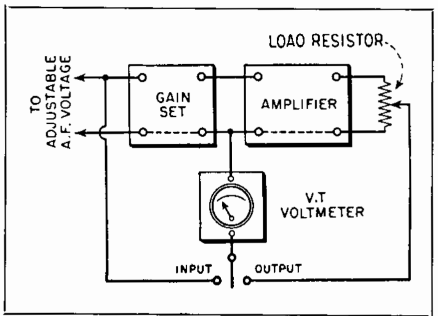


Fig. 28-24.

the voltage of the input signal to deliver an output voltage corresponding to the rated amplifier output power. The proper output voltage may be determined from the equation:

$$E = \sqrt{PR} \quad (1)$$

where E is the required e.m.f. (volts)
P the rated undistorted output power (watts)
R the load resistance or impedance (ohms)

An alternative method of measuring overall gain is shown in Fig. 28-24. The source of audio-frequency test voltage is properly terminated and then connected to the amplifier input through a standard *gain set*. The amplifier is suitably terminated by a non-inductive load resistor of appropriate wattage, and a vacuum-tube voltmeter is arranged to give successive indications of input and output voltages.

The tap on the load resistor enables the operator to select, for output indication, an af voltage within the full scale range of the vacuum-tube voltmeter, while full power output is supplied to the load. This is an advantage since the amplifier will be operated at full output during the measurements without injuring the meter.

With the voltmeter in the input position, the af voltage is adjusted to a value necessary to give full power output by means of the gain control in the oscillator. The meter is then switched to the output circuit and the gain set adjusted to restore the meter reading to this reference level. The gain may then be read directly from the dials of the gain set.

Tests for Feedback

Regenerative feedback is often quite perplexing to the amplifier technician since it may not cause the amplifier to oscillate at an audible frequency, but nevertheless is present in sufficient quantity to cause serious distortion.

When regeneration is suspected, each stage should be examined for its presence by means of grid current or cathode voltage measurements. The presence of grid current in any class of amplifier receiving no input signal is a certain indication of feedback.

Each stage should be examined separately, progressing from input to output circuits of the amplifier. The presence of feedback over several stages, rather than in a single one, may be verified by opening the connection between a stage in which abnormal no-excitation dc voltages or currents have been discovered in the stages immediately preceding and following. Currents or voltages should fall to normal values when the contributing stages are thus uncoupled. Likewise, abnormal currents and voltages will not be found in the isolated stages.

Audible feedback in a previously efficient amplifier generally arises in one stage. In developmental units, it is most often due to unwise layout. In the first case, the af voltage generated is passed along to each succeeding stage and into the loudspeaker. The offending stage is readily located if, in the absence of an input signal the output of each stage—from input to output ends of the amplifier—is inspected with a vacuum-tube voltmeter or sensitive headphones. The first stage delivering a howl to the headphones or deflecting the meter is the one in which audible feedback orginates.

Determination of Impedance

An investigation of the impedance of transformer windings logically will follow the discovery of distortion, although the impedance of an output transformer will be under surveillance when the loudspeaker signal is low. It may also be desirable to check an unknown transformer for secondary impedance when matching speakers to a strange amplifier.

Fig. 28-25 shows a simple method of measuring the impedance of a transformer winding. The unused winding is connected to its usual circuit points or to an equivalent resistor, R_L. The winding under test is connected in series with a variable resistor R_t and a source of 400 cycle controllable af voltage. R_t is a wire-wound rheostat or a laboratory decade box. A vacuum-tube voltmeter is arranged with a s.p.d.t. switch to measure successively the af voltage drop across the resistor and winding.

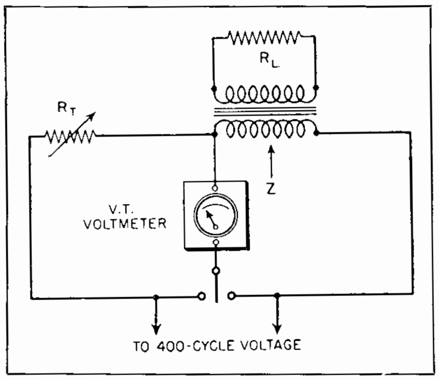


Fig. 28-25.

In making the test, R_t is adjusted until its voltage drop is identical with that of the transformer winding. Its resistance at this setting then equals the impedance of the winding. The input voltage level is adjusted for a good, readable deflection of the meter.

The value of R_t at "balance" may be determined by means of a good ohmmeter or, for more accurate purposes, by means of a resistance bridge.

The impedance of a transformer winding (with the other winding appropriately loaded) might be found also by measuring its inductance with a suitable bridge for the purpose, measuring its resistance, and calculating from the equation:

$$Z = \sqrt{R^2 + (2\pi fL)^2}$$

Where Z is the required impedance (ohms)

R the resistance (ohms)

f the frequency of the bridge voltage (cps)

L the measured inductance (henries)

If the turns ratio of a transformer has been determined previously, the impedance ratio or the separate impedances may be found with the aid of the following formulae:

$$\frac{N_s}{N_p} = \sqrt{\frac{Z_s}{Z_p}}, \frac{Z_s}{Z_p} = \frac{N_s^2}{N_p^2}, Z_s = \frac{Z_p N_s^2}{N_p^2}$$

$$\text{and } Z_p = \frac{Z_s N_p^2}{N_s^2}$$

Where N_p is the number of primary turns

N_s the number of secondary turns

Z_p the primary impedance, and

Z_s the secondary impedance or impedance of the load.

The turns ratio may be determined by applying a known alternating voltage across one winding and measuring the induced voltage across the other. The ratio of the two voltages will then correspond to the turns ratio.

Tube manufacturers state in their tables the values of load impedance (or load resistance) recommended for maximum undistorted power output. This load value is matched to the power-tube plate impedance by means of a transformer of proper turns ratio, as indicated above.

A rapid method for determining the impedance of an output transformer secondary of an amplifier in operation consists in adjusting a standard *output power meter* for maximum deflection and reading the impedance from the meter dials. The power meter is connected to the output transformer terminals, and a 400 cycle signal voltage introduced into the input circuit of the operating amplifier. The meter-range switch is set to accommodate the maximum amplifier power output, and the meter input impedance is adjusted for maximum deflection of the indicating instrument. The reading of the impedance dial at this setting is the impedance of the output transformer secondary.

Power Output Measurements

The audio output watts may be measured with the audio-frequency output power meter or determined from voltage or current readings taken in a properly terminated output circuit.

Since the output power meter presents to the amplifier under test a load impedance of widely adjustable value, this instrument may be connected directly to the amplifier output terminals without introducing any further load or matching device, provided the input impedance of the meter is set to the value of amplifier output impedance. If the latter is not known, the meter impedance may be adjusted for maximum deflection of the indicating instrument with the amplifier passing a 400-cycle signal, whereupon the output impedance may be read from the dials.

Advantages of the output power meter are that its wattage readings are direct, requiring no calculations or conversions, and its input circuit may be closely matched to the amplifier output circuit.

By the voltmeter or ammeter method, output power levels are calculated from voltage or current values measured in a terminating circuit of proper resistance or impedance. This method is illustrated in Fig. 28-26. Circuit A is that for the indication of af power output in terms of voltage; Circuit B for power indications in terms of current readings. In both examples, the load resistor, R, terminates the amplifier output. Its resistance must accordingly be equal to the rated output impedance of the amplifier and its wattage must be sufficient to withstand the maximum power output.

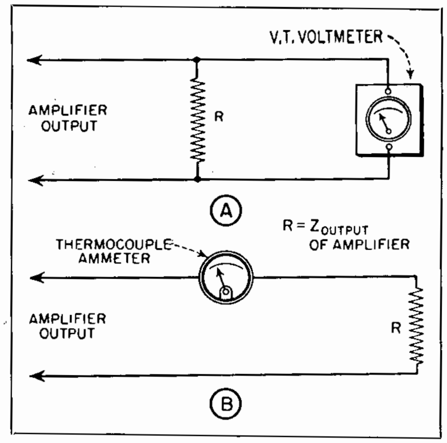


Fig. 28-26.

The ac voltmeter used in Fig. 28-26A must have good accuracy at the test frequency. It is strongly recommended that a vacuum-tube voltmeter be used in this position.

The amplifier power output is determined, in the voltmeter method, by means of the equation:

$$P = \frac{E^2}{R}$$

Where P is the audio output (watts)
E the meter reading (volts)
R the load resistance (ohms)

Alternatively, the value of the a.f. current flowing through the load resistor may be measured, as in Fig. 28-26B, and this value employed in a calculation of the a.f. power:

$$P = I^2 R$$

Where P is the audio output (watts)
I the meter reading (amperes)
R the load resistance (ohms)

For best results, the current instrument should be a thermocouple type ammeter or milliammeter. As a substitute for this type of current meter, a vacuum-tube voltmeter may be employed in connection with a suitable shunt for reading the current in terms of the voltage drop across the shunt.

Hum and Noise Level

Inherent hum and noise level of an amplifier or isolated stage may be quantitatively analyzed either by means of a vacuum-tube voltmeter of sufficient sensitivity or a standard output power meter.

Any output voltage present when the signal voltage is removed, and not due to feedback, may be attributed to hum and noise components. If the one is known to be present in the absence of the other, it may be identified by a simple headphone or speaker listening test.

The simple vacuum-tube voltmeter or output meter will indicate the total magnitude of hum and noise components and will not differentiate between the two nor separately evaluate them. To determine the amplitude of each component, it is necessary to select that component from the total. This operation is best performed by a wave analyzer, which is in principle a highly selective vacuum-tube voltmeter which may be tuned successively to the fundamental hum frequency and each of its harmonics. It will generally be necessary to measure only the fundamental, second, third, and fourth harmonic amplitudes with the wave analyzer.

Sound equipment manufacturers usually express the hum or noise level as a number of decibels below maximum output, and the experimenter may, if he desires, convert the measured values into these units.

After the hum level has been accounted for, the remainder of the no-signal output voltage may be attributed to noise. A certain amount of this arises from thermal agitation in the input stage. However, coupling capacitors in resistance-coupled stages should be inspected for low dielectric resistance, with a good "megger," when the noise level appears excessive. A good coupling capacitor should show a dielectric resistance equal to 500 times the value of the grid resistor used in the succeeding stage.

The Control of Hum

The reduction of hum is a problem that sooner or later confronts every designer or builder of ac powered audio equipment. When there is plenty of room on the chassis and the input signal level is fairly high, say 1 volt, the job may amount to no more than relocating a grid lead or remembering to ground the heater supply. High-gain equipment however, working from a microphone or other low-level input, can present some tricky problems.

Shielding, chassis layout, planning of ground leads, choice of tubes, and filtering all enter into the problem. No instruments are essential to do a good job, although an oscilloscope is very helpful and a high-sensitivity ac vacuum tube voltmeter quite useful.[10]

High-fidelity reproduction of sound demands a much lower relative hum level than does ordinary radio quality. In a good audio system the hum (as well as other noise) should be 50 or 60 db below average listening level. Out of a 5-inch speaker a 30 db ratio is tolerable. Public address amplifiers are often rated in hum level below full output. In a high-quality home installation this type of rating is misleading because a good 20-watt amplifier in a living room is usually operated at 1 or 2 watts output, while the pa system is run closer to full blast.

The low signal levels available from tape playback heads and variable-reluctance phonograph pickups, together with the bass equalization required, aggravate the situation. Two other factors tend to alleviate it somewhat; the comparatively small dynamic range of records and tape—30 to 50 db, and the low impedance of such pickups at 60 cycles. With these low-level input sources in wide use, however, good quality reproduction requires more careful attention to the reduction of hum than ever before.

Plate supply ripple due to insufficient filtering is always a pure 120-cycle note, and can be readily located by clipping extra condensers across the "B" supply. The actual design of power supply filters is a process of calculating the minimum size of chokes and condensers to insure a certain hum level. Modern electrolytic condensers are cheap and of excellent quality, being used in some of the finest

[10]Fleming, L., "Controlling Hum in Audio Amplifiers." Radio & Television News, Nov. 1950.

laboratory equipment. A single pi-section filter, using a good choke, is generally adequate for any power stage, and an additional *RC* section for each preceding voltage amplifier stage will insure freedom from ripple as well as providing plenty of decoupling. There should be some series impedance in the positive side of the filter system, not all in the negative side, to get rid of ripple coupled through the power transformer capacitance. It is important to remember that filtering action is better for a given section, the lower the dc current drain, so that an *RC* filter section supplying only a milliampere or two to a resistance-coupled stage will give you much more filtering than a similar section feeding a stage that draws a lot of current.

Many excellent commercial amplifiers use no filter chokes at all, and many fewer sections than suggested above. One can usually get satisfactory results with this type of design, but the proper procedure is to start out with plenty of filter, and then start cutting down and rearranging.

Push-pull stages require less filtering and decoupling than single-ended stages. Power pentodes are more tolerant of plate (not screen) supply ripple than are triodes, because of their high plate resistance, *i.e.*, a volt of ripple produces less plate current swing. Negative feedback over the output stage will attenuate hum originating in that stage, but will not reduce hum originating at the input end of the feedback loop.

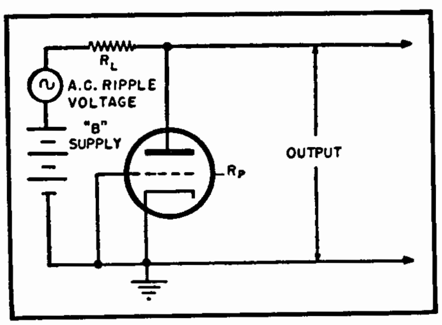


Fig. 28-27. Voltage divider effect attenuating plate supply ripple. The tube output is less than the "B" supply ripple by approximately the ratio of plate load resistance to plate impedance of the tube.

Plate Impedance Effect

The ripple voltage actually appearing at the plate of a tube is less than that at the "B+" terminal by the ratio: $R_p/(R_L + R_p)$ where R_L is the plate load resistance, and R_p the plate resistance of the tube. In Fig. 28-27, if the tube is a 6J5 operating with a 100,000 ohm load R_L and a plate impedance R_p of 10,000 ohms, only 1/11 of the "B" supply ripple will appear at the plate. An attenuation of at most 2 or 3 to one can be obtained with high-mu triodes, and practically none with pentodes.

Grounding and Balancing

In wide-range, low-level equipment with ac heater supply to the input stage it is advisable to use a hum-balancing potentiometer of 200 ohms or so across the heater supply, with the arm grounded. This adjustment will usually reduce the hum level of the input stage 10 db over the level obtained when using a fixed centertap. The total reduction over an arrangement with one side of the 6.3-volt supply grounded is from 20 to 30 db. The optimum adjustment varies slightly from tube to tube, and is less critical the lower the impedance in the grid circuit.

Heater supplies should be tied to ground or to a point having some definite dc potential (and bypassed to ground with 0.1 μfd. or so) but the reason why may not be obvious. A floating heater supply winding assumes a high ac potential, which it obtains through its capacitance to the high-voltage winding of the power transformer. Through capacitance and leakage from heater to grid, and other paths, this high ac potential couples large hum voltages into the signal circuits. It can even break down the heater-cathode insulation inside the tubes.

When one side of the heater supply is grounded, the ac potential available for such stray coupling into the signal circuits is only 6 volts instead of perhaps 300.

When the center-tap is grounded, two ac potentials of 3 volts each are on hand, to couple hum into the grid circuit, but are 180 degrees out of phase. Their effects approximately cancel. If the stray impedances from each side of the heater to the grid are exactly equal, an exact center-tap will neutralize the hum. Since they are not exactly the same an adjustment is necessary.

TYPE	HEATER	DESCRIPTION
12AY7	6.3/12.6 v.; .3/.15 amp.	Double triode, noval miniature, mu = 35
5879 (RCA)	6.3 v.; .15 amp.	Voltage amplifier pentode, noval miniature
347A (Western Electric)	6.3 v.; .6 amp.	Triode, glass octal with top cap, mu = 16

Table 1. Tubes recommended for low-hum, low-microphonic applications.

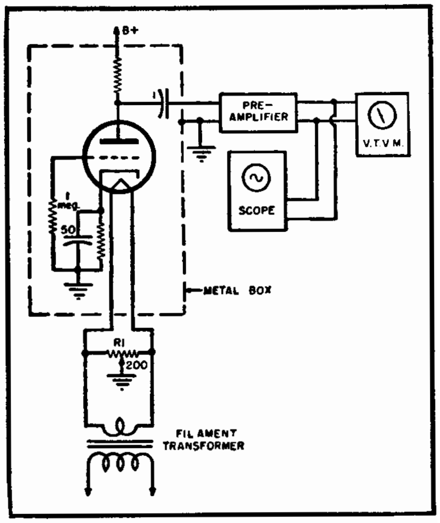


Fig. 28-28. Setup for measuring tube hum.

Choice of Tubes

Table 1 lists tubes recommended by the manufacturers for low hum, noise, and microphonic applications. Table 2 gives hum data on some standard receiving types.

For commercial service it is better to use one of the special types listed in Table 1, since the hum characteristics of ordinary types often depend on variables that are not controlled in production, and the special types are better and more consistent microphonically. Type 1620 was omitted from Table 2 because it is a selected 6J7.

A selected 6AU6, with a hum balance adjustment, can be so quiet that nothing but thermal noise can be seen on an oscilloscope, *i.e.*, an equivalent hum input at the grid of less than 10 microvolts, when using a 1 megohm grid resistor.

TYPE	HUM LEVEL (Referred to Grid)	REMARKS
6AU6	5-10 μv.	Fairly uniform
12AT7	10-30 μv.	Not very uniform, some tubes noisy and microphonic
6J7	30-50 μv.	Fairly uniform and quiet
6C4	10-100 μv.	Wide variation, some tubes very noisy

Table 2. The measured attainable hum levels of standard receiving tubes.

The data in Table 2 is based on measurements made on four tubes of at least two different makes of each type. The test circuit is shown in Fig. 28-28. Plate resistors are 470,000 ohms for pentode types and 100,000 ohms for triodes. Cathode and screen resistors follow tube manual recommendations. The 6AU6 is best, the 12AT7 next, and a close third is type 6J7. 6C4's are spotty and generally poor. All these types require selection for good results.

Figs. 28-29 and 28-30 show typical circuits recommended by the manufacturers for the 12AY7 and the 5879. An exception is that screen and cathode bypasses are shown larger for the 5879 than the data sheet recommends, because it is felt that cutting down the

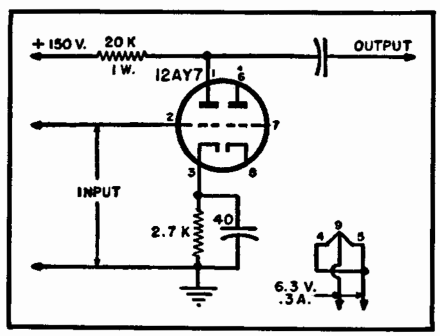


Fig. 28-29. Manufacturer's recommended circuit for 12AY7 as a low-level amplifier. Only one of the triode sections is shown.

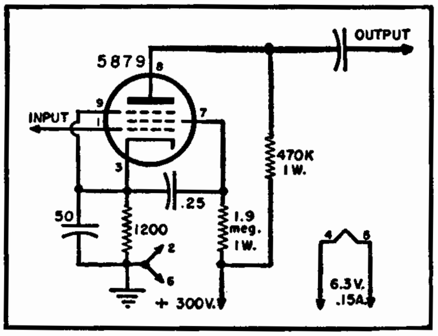


Fig. 28-30. Recommended circuit for RCA 5879 as a low-level audio amplifier. Pins 2 and 6 should be grounded. Gain is 150.

bass response is not the best way to reduce hum.

Fig. 28-31 shows a complete low-level input stage using a 6AU6, on which was measured a hum output, equivalent to less than 10 microvolts at the grid. The "B" supply was a conventional transformer and rectifier system with one pi-section of filter, but the only current drain was that of the stage shown.

Several things inside a tube contribute to ac ripple in its output. Consideration of these factors leads naturally to circuit planning for hum reduction. The first source of hum in the tube is leakage current between the heater and the cathode. This current is said to be around 0.04 microampere for voltage amplifier types and 1 microampere for power tubes. No definite data is available, but this leakage current probably increases greatly over these values as the tube ages. It is known to increase with heater voltage. Low-level audio stages should never have over 6.3 volts applied to the heater, and preferably about 5.8 volts. In the circuit of Fig. 28-31, increasing the supply from 5.8 volts to 6.6 volts doubled the hum output.

The effect of this leakage current is eliminated by keeping the dc potential low between heater and cathode (not over 10 or 20 volts) and by using a cathode bypass capacitance of not less than 40 μfd.

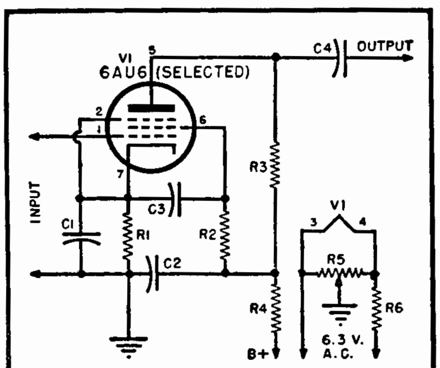
Electrolytic condensers are perfectly quiet. Grid bias cells however are apt to be noisy at millivolt levels.

Heater-Grid Leakage

Resistive leakage from heater to grid can be troublesome. If, for example there is 10,000 megohms of unbalanced leakage from a 6 volt source to the grid, and the input circuit impedance is 1 megohm, 0.6 millivolt of 60 cycle ac will appear at the grid. Leakage of this magnitude is not unusual across black phenolic tube bases and sockets, not to mention cotton insulation on pushback wire. With single-ended tubes, mica-filled or ceramic sockets are advisable.

Heater-Grid Capacitance

The capacitance between the heater leads and the grid is an important hum characteristic of a tube type. The low-noise types in Table 1 have internal shielding between the grid and heater. 1 $\mu\mu$fd. of stray unbalanced capacitance here will introduce 2 millovolts of hum into a 1 megohm grid circuit. The situation is saved in practice by hum balancing and by the fact that the actual capacitance inside miniature tubes is less than 1 $\mu\mu$fd., and that the impedance presented to the grid circuit by phonograph pickups, tape heads, and microphones is a good deal less than a megohm, pro-



R_1—2200 ohm, ½ w. res.
R_2—1.5 megohm, 1 w. res.
R_3—470,000 ohm, 1 w. res.
R_4—100,000 ohm, ½ w. res.
R_5—200 ohm wirewound pot.
R_6—2 ohm, ½ w. res.
C_1—50 μfd., 25 v. elec. cond.
C_2—10 μfd., 450 v. elec. cond.
C_3—.25 μfd., 400 v. cond.
C_4—.05 μfd., 600 v. cond.
V_1—6AU6 tube

Fig. 28-31. Low-noise preamp using 6AU6.

vided, that is, that the layout minimizes the stray capacitance which unavoidably remains.

Capacitive Hum Balancing

A fine adjustment on heater supply center-tapping can be obtained by trimming the capacitances between either side of the heater and the grid, as well as by using a potentiometer. Adequate trimming can usually be affected by lead dress alone. Bring an inch or less of heater lead close to the grid lead, using an insulating rod for a tool. One side of the ac supply will increase the hum level, the other side will reduce it. Observation of the tube output on an oscilloscope or meter is rather necessary during this operation.

Wiring Techniques

Most of the considerations outlined above as relating to hum in tubes apply as well to the wiring that goes to the tubes. Capacitance from any ac wiring to a low-level signal circuit introduces a 60 cycle voltage therein in direct proportion to the ac voltage present, the stray capacitance, and the impedance of the affected circuit.

Suppose we have a grid circuit, Fig. 28-32A, with a resistance R of 1 megohm, having its high side coupled to a source of ac E_H through a stray capacitance C_H. If the hum voltage E_H is 1 volt, there will appear at the grid 1/2300 of a volt, or about half a millivolt, since the reactance of 1$\mu\mu$fd. at 60 cycles is about 2300 megohms. This is only 26 db below 10 millivolts. If C_H should be coupled to a 300-volt power transformer lead, the grid circuit would receive 300/2300 volt, or 0.15 volt—an intolerable level. 1 $\mu\mu$fd. is about what you get between two pieces of pushback wire an inch long twisted loosely together.

Fig. 28-32B illustrates the same situation when the signal source C is capacitive, as crystal microphones and pickups. The stray "hum" capacitance and the signal source act simply as a capacitive voltage divider. If C is 2000 $\mu\mu$fd. and E_H 1 volt, the voltage at the grid will be 1/2000 volt.

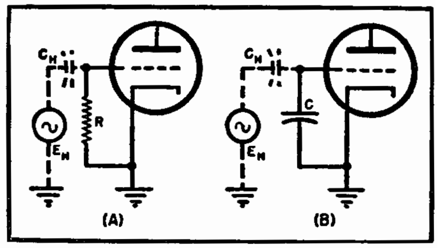


Fig. 28-32. (A) Capacitive hum pickup. (B) Capacitive hum pickup in crystal microphone or in phonograph input circuits.

All this does not mean that one should shield the daylights out of all low-level wiring, and locate input stages a yard away from everything else. Rather, it is necessary to be aware of the magnitudes of these effects, and to recognize a defective wiring layout. Lavish use of tinned copper braid is rarely necessary. It is really needed only to salvage a poorly-planned chassis layout, or in unusually compact assemblies, or when gain and tone controls are grouped together at some point removed from the associated circuits. The capacitance between two wires can be very close to zero if (1) they are a couple of inches apart, (2) they are kept close to the chassis, not up in the air; and (3) advantage is taken of natural cover to help in shielding, *i.e.*, bypass condensers, dc wiring, and other "cold" components.

Twisting of heater leads is not necessary if they are close together and grid leads are kept away. With top-cap tubes it is well to use braid on the grid lead and to put a "hat" shield over the grid cap.

Magnetization of Tubes

Glass tubes of a given type show more individual variation in hum level than metal tubes. The British journal *Wireless World* has reported that this phenomenon is associated with permanent magnetization of the tube elements, which seems to increase the hum level in a way not yet understood. The remedy is demagnetization with a gradually-removed 60 cycle field, the same as jewelers do with watches. The winding of an old power transformer, removed from the core, can be used for a demagnetizing coil. Connect the secondary across the 115 volt line (using series resistance if necessary, to keep it from getting too hot), insert the tube, then slowly withdraw it.

Chassis Currents

Gradients of the order of half a millivolt per foot often exist across a chassis on which a power transformer is mounted. The transformer winding induces a 60 cycle voltage, causing a circulating current in the chassis. Hence it is not good to include a chassis in series with a low-level input circuit. Fig. 28-33C illustrates what happens when the cathode and grid of a tube are returned to different points. E_H denotes a hum voltage distributed along the chassis. This voltage is included right in with the signal. The remedy is to return grid and cathode to the same point, as in Fig. 28-33F.

Ground Loops

When one starts wiring from a diagram without thinking too much about the purpose of each lead, the situation indicated in Fig. 28-33B may develop. Here, too many ground leads have been used, forming a shorted turn or ground loop. Stray magnetic fields eagerly induce ac voltage in such a loop, and the voltage across a portion of the loop is included in the grid circuit. The effect is usually very small, but an aggravated case can cause trouble.

If the grid lead and grid return be too widely separated, as indicated in Fig. 28-33E, another one-turn coil is created, ripe for the induction of 60 cycle voltage. Since the voltage induced in a loop is proportional to its area, the remedy is to keep the grid lead and return lead close together.

These magnetic effects are aggravated when they are on the low-impedance side of an input transformer. Phonograph motors have magnetic fields just like power transformers.

Magnetic Pickups

Fig. 28-33A shows a situation often encountered when using pickups of low output level, such as the *GE* and *Pickering*. Here the shielded pickup cable is grounded to the motor frame at one end, and to the amplifier chassis at the other. The motor frame has a fair-sized capacitance to the 110 volt line, as does the amplifier chassis. These capacitances are indicated as C_1 and C_2 in the diagram, with the line voltage in between. C_1 and C_2 can be as large as 0.1 μfd. if line isolating condensers are used. A hum voltage is thus applied across the ends of the cable shield, and the resulting voltage drop can be as much as 50 microvolts,

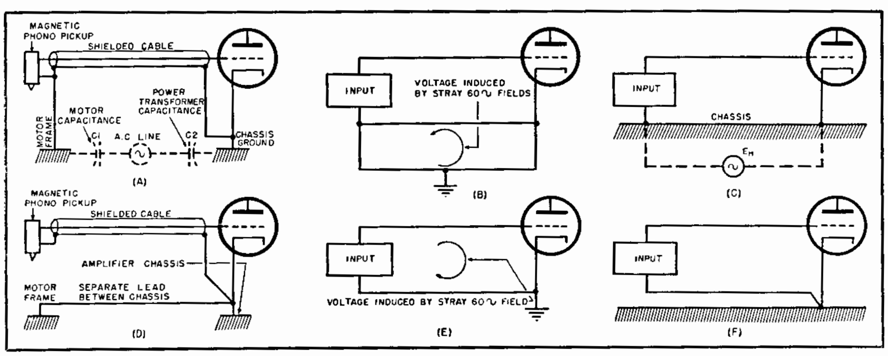


Fig. 28-33. (A) Hum induction by using cable shield to interconnect phonograph turntable and amplifier chassis. (B) A ground loop. One of the three ground leads should be opened. (C) Hum injected into input circuit from ac voltage gradient along chassis. (D) Correct method of connecting low-level phonograph pickup to amplifier. (E) An input wiring loop. This con give trouble in tope playbock input circuits if loop is close to motor or power transformer. Shielding grid leod alone will not help. Loop must be closed up by moking shield the only path between low side of signal source and cathode return of first tube. (F) Grounding grid return and cothode to some point to avoid hum from chassis gradients.

included directly in the grid circuit. The remedy is to use the shield only for the signal voltage, and to employ a separate wire to connect the two chassis, as shown in Fig. 28-33D. This wire should be close to the cable to avoid loop pickup effects which can be just as bad. Two-conductor shield cable can be used, too.

Tape Recorders

Tape playback is the most critical problem of all, because of the very low output level of tape heads—1 millivolt or so—and the heavy bass equalization required. All precautions must be observed. An important residual hum source is ac magnetic induction in the playback head itself. *Audio Engineering* has reported a simple method of neutralizing the residual induced hum. Attach a small piece of sheet iron, ½ inch square or so, to a stick, and try holding it in various positions near the playback head until a location is found that dips the hum level to a minimum. Then permanently fix the iron piece in place. *Permalloy* is better.

The iron acts as a distorter of the field, and, in the right position equalizes the effective fields applied to the two sections of the hum-bucking winding used in the head.

Audio Transformers

Unshielded low-level input transformers pick up hum from power transformer and motor fields, typically on the order of a few millivolts across the high winding. At microphone levels, an ordinary strap-mounted or similar input transformer simply cannot be used on the same chassis with a power transformer. This is why intercom sets always use an ac-dc type circuit.

Transformers of the two-coil "hum-bucking" construction are somewhat better. If adjustably mounted at least a foot from the power transformer, such a unit can be oriented to be quiet enough for pa purposes, but not for broadcast quality. The small "ouncer" size is about as good, since hum pickup decreases with size.

By far the best construction is that with multiple telescopic *Permalloy* shields. A cast iron case alone does little good. Multiple alloy shielded units are at least 30 db better than any other construction.

Hum Characteristics

Fig. 28-34 shows tracings taken from an oscilloscope screen, with typical waveforms of ac pickup from various sources. They have certain distinctive qualities. Straight capacitive pickup is a slightly rough 60-cycle wave. Plate supply ripple due to insufficient filtering is a very smooth 120 cycle note. Hum, due to inadequate cathode bypassing, or an ungrounded heater supply, is often an asymmetrical 60 cycle wave with sharp corner, Fig. 28-34B.

Inductive hum picked up by transformers is very rich in harmonics (Fig. 28-34C) but free from sharp peaks and spikes. It has a sort of musical sound, like the hum sometimes heard on telephone lines.

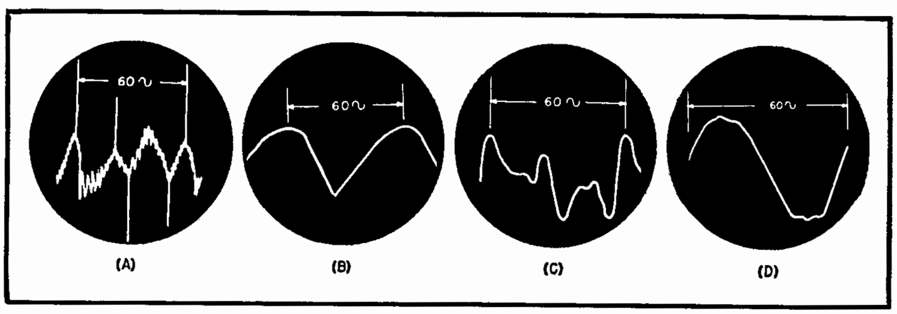


Fig. 28-34. (A) Noise pickup from fluorescent lamp. (B) Heater-cathode leakage current. (C) Inductive hum pickup by audio transformer. (D) Capacitive hum pickup from ac wiring.

SOURCE	REMEDY
1. "B" supply ripple	Filtering; use larger screen bypass in pentodes; use high load resistance with triodes.
2. Capacitive pickup from ac wiring to tube grids	Separation of leads; shielding; use potentiometer for grounding center of heater supply.
3. Heater-grid leakage	Mica-filled or ceramic sockets; keep wiring separated.
4. Heater-cathode leakage	Use cathode bypass of at least 40 μfd. Try positive dc bias of about 5 v. on heater supply.
5. Magnetic ac modulation of tube plate current	Locate input tube 6" or more from power transformer and phono motor. Use iron shield on glass tubes.
6. Chassis gradients	Ground grid return and cathode return to same point.
7. Gradients along cable shields	Ground mike or phono pickup shield at one point only; use separate lead to interconnect chassis; keep both leads close together.
8. Ground loops	Keep grid and return leads close together and away from power transformer. Interconnect ground tie points through only one path. Do not duplicate separate chassis grounds with wire interconnections.

Table 3. Some of the most common sources of hum and a listing of possible remedies.

Fluorescent lamp interference is characterized by sharp spikes and hash as in Fig. 28-34A. The frames of the fluorescent lamps that are placed near the bench should be well grounded and the lamps themselves kept at least two feet from the work.

Miscellaneous Sources

Broadcast receivers sometimes exhibit hum only in the presence of a received carrier. This is due to rectification of rf picked up by the house wiring, by copper oxide in loose connections therein. The rectified signal is modulated by ac. The remedy is to eliminate rf signals getting into the set from the ac line, by grounding both sides of the line to the chassis through condenser of about .005 μfd.

The vhf oscillators, such as oscillators in FM and TV receivers, acquire a 60 cycle frequency modulation in circuits where the cathode is "hot" to ground and the heater is not, *i.e.*, where cathode and heater are not at the same rf potential. The mechanism of this effect is not understood. The cure is to use a circuit where cathode and heater have the same rf potential. It is satisfactory to use either a grounded-cathode oscillator circuit, or a "hot" cathode circuit where cathode and heater are connected together, feeding the free side of the heater through an rf choke.

Center-tapped heater supplies are not too feasible in systems involving high-gain, multi-stage rf or if amplifiers, because one side of the heater has to be grounded at each tube, preferably right to the chassis, to prevent regeneration by coupling through the common heater wiring.

Cutter Head Measurement

There are several means used for determining the response of cutters and associated equipment. One is the use of an audio oscillator, vacuum-tube voltmeter and the regular output level meter on the recording amplifier. With this combination we can obtain definite patterns on the discs (Buchmann-Meyer effect) that will show exactly what is taking place. A disc that has been cut *properly* with a sharp stylus, preferably sapphire, will have grooves which are *shiny* in appearance and capable of reflecting light rays when placed in proper position with respect to the eye of the observer. Unfortunately, much detail is

[11]Dickerson, A. F., "Hum Reduction," Electronics, Dec. 1948, pp. 112-116.

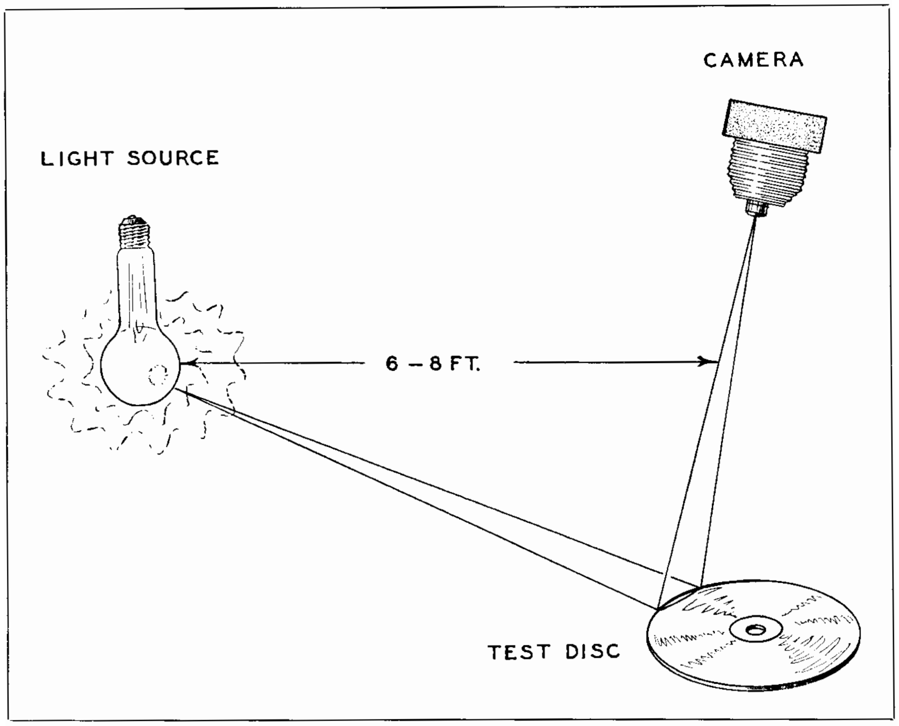


Fig. 28-35.

lost in reproducing the original photos, but they illustrate the effects.

An improperly cut disc (one cut with a dull stylus) cannot assume this shiny character and is not suited to the study and analysis of frequency response. The setup used to obtain the following photos makes use of a single light source. The author used a No. 2 photoflood lamp without reflector and a standard make of 5″ x 7″ camera.

The setup is shown in Fig. 28-35.

The electronic end of the setup consists of Audio Oscillator, a VTVM to keep the audio output of the oscillator at a fixed level, a recording amplifier, the db meter to observe the cutting level at the cutting head, and the recording table and fresh disc.

The records are all at a standstill in the illustrations and were made from the same general position of the camera except for distance.

The most important tests for the layman deal with the determination of cutter response as it is called upon to modulate the various frequencies on the disc. If these are lacking or improperly cut, or if too high or too low a cutting level is used, he may, with this setup, determine where the system is lacking and take proper steps for correction where needed. The procedure may be used on any type of lateral disc recorder.

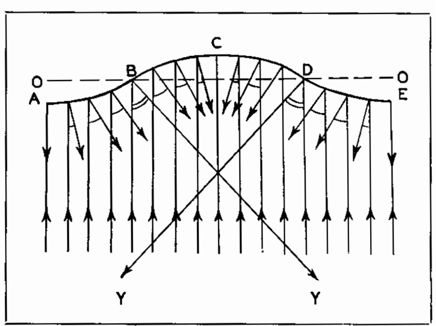


Fig. 28-36.

Light Patterns

Fig. 28-36 illustrates the principle of the optical method of measuring veloc-

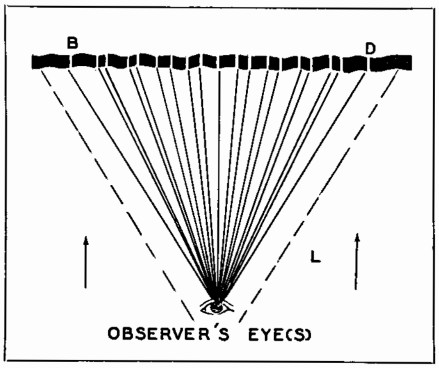


Fig. 28-37.

ity/amplitude on records, due to *Buchmann* and *Meyer* and so named after them. The curved line ABCDE at the top of the figure represents the side of a groove, in a sinusoidal section of the disc material. Parallel light (a point source) falls upon it and is reflected. The largest angle, Y, between an incident and reflected ray, occurs at the inflection (bending) point of the curve. If OO is the time axis (or the groove-width) the curve ABCDE depicts the displacement, so that Y is given by dx/dt, which is the velocity/amplitude, or lateral velocity, etc. No larger angles than Y are possible. In Fig. 28-37 a number of reflected rays is shown. The outermost ones radiate from the points of inflection. A luminous band can be seen if the line DB is moved along its continuation so that the reflecting spots dissolve into each other. The width of the light band is determined only by the velocity/amplitude of the curve, and is independent of the position of the observer's eye(s). When the line OO is a circle, *i.e.*, a groove, on a disc, the same result is obtained.

Now let us analyze the first test record, Fig. 28-38. This disc was cut at 33⅓ rpm on a 13¼ inch *disc*, inside-to-outside. The minimum diameter was 5″. A (plus) 18 db cutting level was maintained over the entire frequency range and the controls on the amplifier were set so that the audio response is essentially flat from approximately 20 cps to 15,000 cps. Note that the absence of equalization has a definite effect on the cutting capabilities.

The frequencies appearing on the disc, from the right (hub) to left on the illustration are as follows: 1000, 2000, 3000, 4000, 5000, 6000, 7000, 8000, 9000, 10,000, 11,000, 12,000 cps and then in the same steps in reverse back to 1000 cps and then on through the lower frequencies of 900, 800, 700, 600, 500, 400, 200, 100, 50 and 25 cps. The small light portions between each series of grooves that were cut, indicate the normal surface noise, or unmodulated grooves. A complete story has been told on the finished disc—and the interpretation follows. The middle register, from 1000 to 7000 cps shows a rather even groove modulation, as indicated by the width of the cut lines. The response begins to taper off at 7000 cps and we still observe that cutting is taking place up to over 10,000 cps. Then the surface noise level almost equals the modulated level.

From the above observation it is apparent that we must *boost* the frequencies *above* 7000 cps by several decibels in order that they be modulated on the disc, if the finished recording is to be brilliant in character, and if we expect to be able to hear these frequencies above the surface-noise level.

The effects of boosting may be observed in Fig. 28-39 and 28-40. Returning once more to the analysis of Fig. 28-38, we find that we have cut the disc properly from the 7000 cps range down to 50 cps. At 25 cps we can see a slight dropping off in response, but this is still above the surface noise by the amount of 2 or 3 db. The pattern on the outside of the disc is 400 cycles cut at a constant reference level (in this case) of (plus) 14 db.

If it were possible to compensate for the loss of the higher frequencies, we could increase the width of the pattern at those frequencies and this would be done by increasing the cutting level

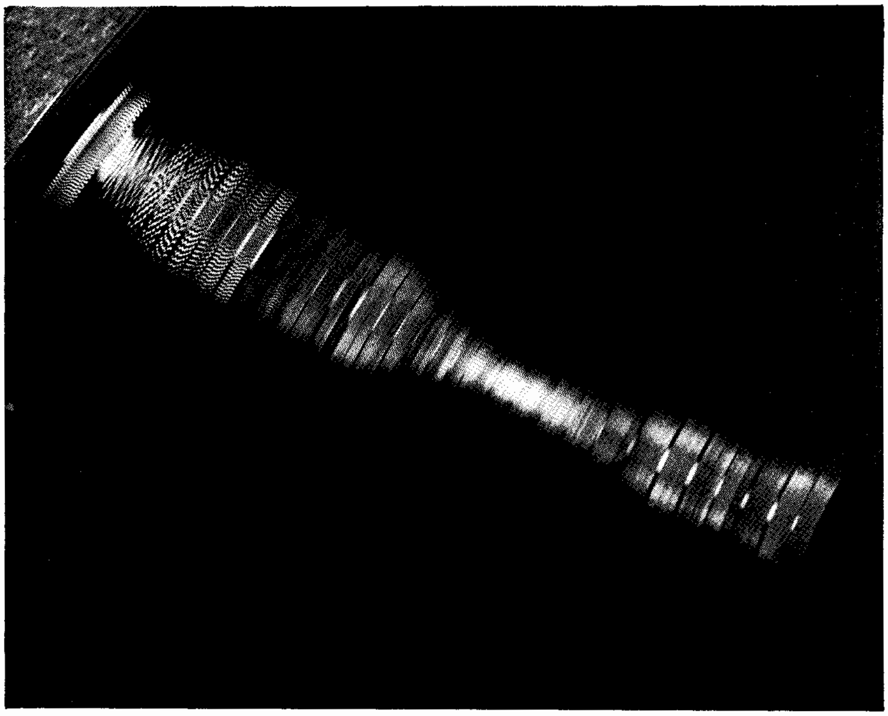


Fig. 28-38.

Fig. 28-39.

(volume) over the range that requires this treatment. This may be done by proper use of the bass and treble equalizers (tone controls).

Now we shall make another test recording to determine how far we can boost the high frequencies without getting into trouble from distortion, etc. Fig. 28-39 illustrates the effect. Remember that *the widest patterns are cut at the highest level,* and *the narrowest grooves are cut at the lowest level.* Starting at the right side of the illustration we find our first frequency, 10,000 cycles in this case. This disc was started at a minimum diameter of 6″ in order to attain more velocity (record speed) in order that the high frequencies could be more fully modulated.

The treble control on the amplifier was set for an increase at 6000 cycles of (plus) 8 db above normal. Thus, volume was increased automatically at this frequency so that a note of 6000 to 7000 cycles appearing at the amplifier would be boosted by that amount. The bass control was left at normal response. The audio oscillator output was kept at a *constant level* while the cutting took place.

Analysis of Fig. 28-39 shows that there is a definite peak both at 7000 cycles and 6000 cycles. In fact the two are cut at about the same amplitude. Counting the groove cuts from the inside-out, right to left, we find the following frequencies: 10,000, 9000, 8,000, 7000, 6000, 5000, 4000, 3000, 2000, 1000, 900, 800, 700, 500, 300, 100 and 50. Note that the frequencies below 300 cycles fall off rapidly in amplitude. This is necessary in cutting "Constant Velocity" records and is known as the "turnover" or point where the characteristics are altered in the equipment so that the bass frequencies will not overcut the groove walls, due to the greater stylus displacement. However—these low notes will still be reproduced satisfactorily and the bass control may be

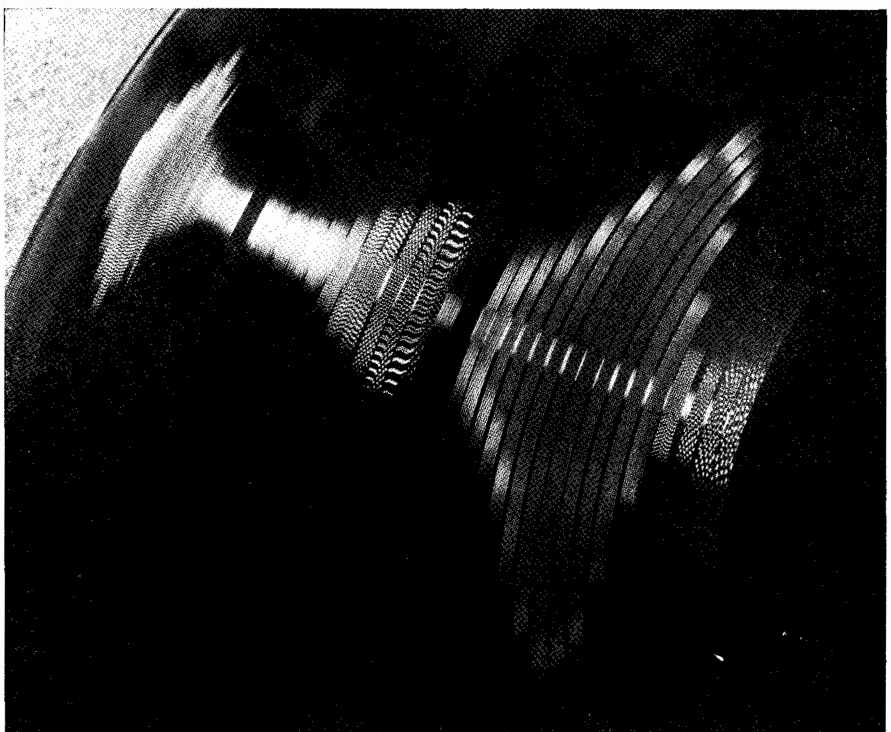


Fig. 28-40.

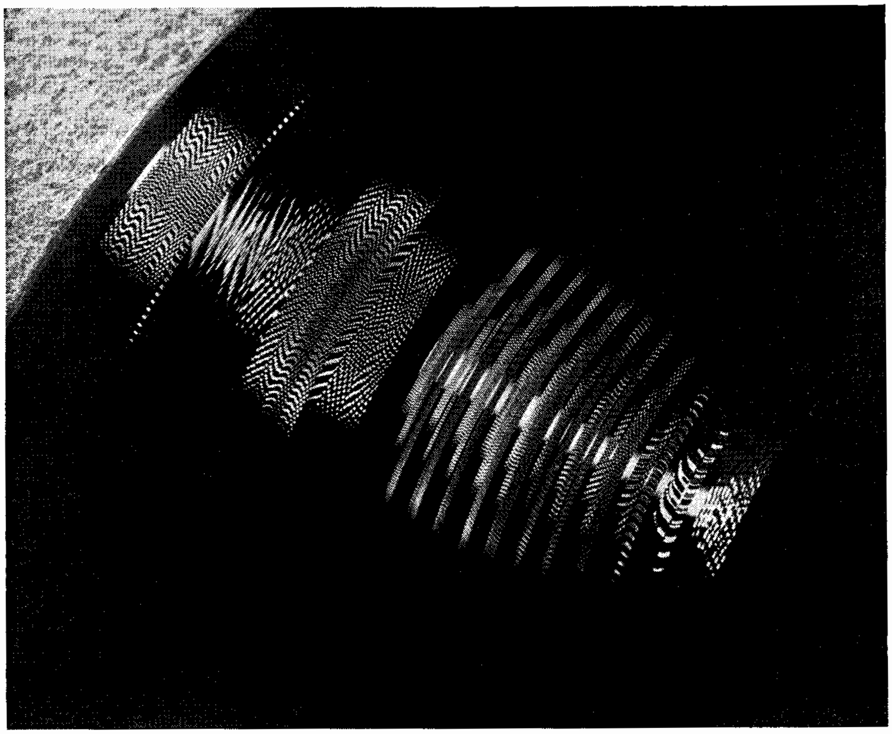


Fig. 28-41.

set to effect a boost when the record is played back.

The second portion of the disc, from the narrowest part out, is cut with a rising frequency in linear steps from 50 cps to 15,000 cps. Note that the response is very uniform from about 100 cps to well over 7000 cps. This is considered very satisfactory from the listeners point of view. However we can still improve matters to increase this response to include a greater audio range as we shall see later.

The third portion has been cut with a 1000 cycle note in steps of 2 db from (plus) 12 to (plus) 26 db and returning in 2 db steps to (plus) 12 db. This illustrates the effect of *cutting level.* Note the wide pattern when high levels are reached. In fact, we actually overcut the record in order to illustrate the point. Now if we play back the test record and observe where distortion is heard, we have an accurate guide for the proper cutting level to use on our own particular equipment.

Other frequencies may be observed under the same conditions, and the patterns will give the same indications.

Fig. 28-40 illustrates a series of three tests. First is a frequency run from 50 to 12,000 cycles, cut at a level of (plus) 18 db. The treble control at the amplifier was set to full boost at 4000 cycles and the bass control to full boost. The steps are as follows: 50, 100, 250, 500, 1000, 2000, 3000, 4000, 5000, 6000, 7000, 8000, 9000, 10,000, 11,000, 12,000 cycles.

Observation discloses that the frequencies of 50, 100, 250 and 500 cycles are cut at a uniform amplitude. There is an increase of approximately 2 db at 1000 cycles and from there up to 5000 we observe steps of better than 2 db. for each successive frequency. The middle register has been boosted considerably from the normal flat response of the amplifier and the greatest peak volume is at 4000 cycles.

The second part seen on the disc was cut with the bass frequencies cut off below 200 cycles and the highs attenuated above 800 cycles. Comparison with the data kept on this disc shows that the frequencies above 1000 cycles fall in level to that of the disc and are completely missing when the record is played back. This cut was made to illustrate the effect of peaking at 300 cycles without the extreme high and low notes.

The outside cutting is at 1000 cycles, similar to that on Fig. 28-39. Here we used steps of 1 db instead of 2 db. This gives a sort of crescendo effect to the note and the action may be observed readily. The *maximum* cutting level was (plus) 18 db at the head.

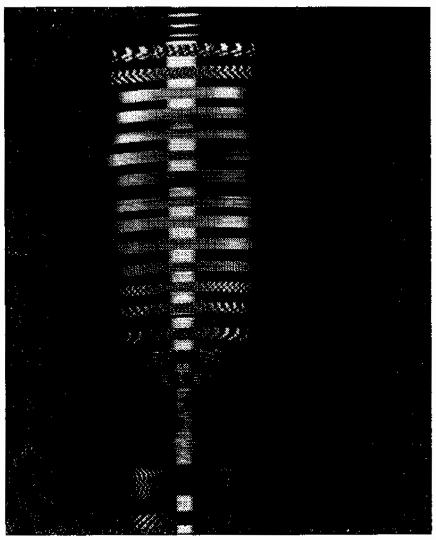


Fig. 28-42A. The Clarkstan 2000S steady state frequency record. (Courtesy Clarkstan).

Next we come to the patterns illustrated in Fig. 28-41. These are rather unusual in some respects and were made to illustrate the effect when notes are cut at various amplitudes. This test record is divided into two parts: The first cutting (inside) was made as follows: Frequencies of 50, 100, 200, 300, 400, 500, 600, 700, 800 900 and 1000 cycles were cut, each at three volume levels, (plus) 18 db, 14 db, and 10 db. Observe the heavy pattern left in the 50 cycle grooves and note how the taper widens as the volume is increased. This is more clearly illustrated on the rest of the cuts from 400 cycles on up. This test shows excellent uniform response from 500 cycles to 1000 cycles, in fact it is safe to state that this is within ½ db. The test could have been continued for the remainder of the available space and the pattern would indicate the response over whatever range we decided to use.

The remainder of the record was cut at 200, 400, 300, 100, 50, and 25 cps at a level of (plus) 18 db while the outside straight-sided cuts are 400 cycles at (plus) 18 db as a reference. Note that the response falls off at 50 cycles which is normal without any boost. These cuts are to be used for low-frequency reference standards and therefore no treatment was wanted.

COMMERCIAL FREQUENCY RECORDS

Several companies, including *Clarkstan*, produce special Vinylite records for

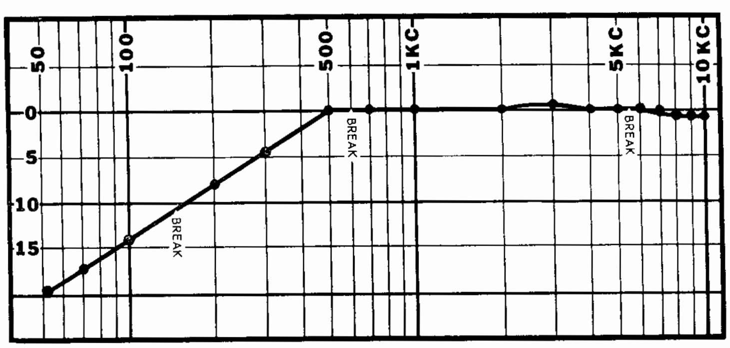


Fig. 28-42B. Frequency response curve of No. 2000S.

test purposes. The *Clarkstan* steady state frequency record, Fig. 28-42A, is recorded constant velocity above 500 cps and constant amplitude below 500 cps.

Frequencies:

500 cps, 1000 cps, KC: 10, 9, 8, 7, 6, Break, 5, 4, 3, 2, 1 and cps: 700, Break, 500, 300, 200, Break, 100, 70, 50, 1 KC and 500 cps.

Groove Width: .0058″
Playback stylus: .0027″ Radius 47° included angle.
Double Displacement at 1000 cps .00098″
Velocity: 7.9 cm/sec peak
Lines per in.: 105
Cutting Stylus: 87½° included angle .001″ rad.
Crossover: 500 cps
Cutting Head: Olson

The frequency response of the *Clarkstan* 2000S is shown in Fig. 28-42B.

A MICROGROOVE STEADY STATE FREQUENCY RECORD

The *Clarkstan* Number 2001S, Fig. 28-43A, recorded on 12-inch Vinylite at 33⅓ rpm, employs the modified NARTB curve.

Frequencies:

KC: 1, 10, 9, 8, 7, 6, Break, 5, 4, 3, 2, 1.5, 1 and cps: 700, Break, 500, 400, 300, 200, 150, Break, 100, 70, 50, and 1 KC.

Groove Width: .0023″
Playback stylus: .001″ Radius, 47° included angle
Double Displacement at 1000c:- .006″
Velocity: 4.3 cm/sec peak Lines per in.: 180
Cutting Stylus: 87° included angle, .0002″ radius
Type of Cutting Head: RCA—Type MI—11850-C (Temperature Compensated).

The frequency response of the *Clarkstan* 2001S is shown in Fig. 28-43B.

Microscope Method

The microscope method is suitable for initial calibration of a cutter, especially if adjustments can be made without removing the head. But the method is slow and tedious and is inaccurate at the higher frequencies where, due to constant stylus velocity, the amplitude of motion is small and the spot of light is no longer small in comparison with the amplitude of movement. Most recorders maintain constant amplitude stylus motion below a frequency, known as the cross-over frequency, and constant velocity above, so that, at the higher frequencies the amplitude decreases, since the product of frequency and amplitude

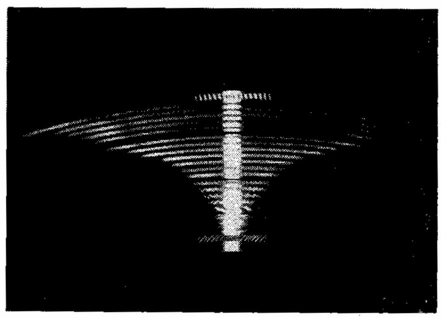


Fig. 28-43A. Clarkstan microgroove steady state frequency record. (Courtesy Clarkstan).

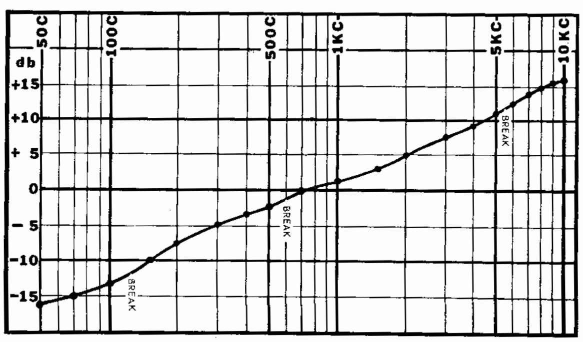


Fig. 28-43B. Frequency response curve of No. 2001S, Microgroove record.

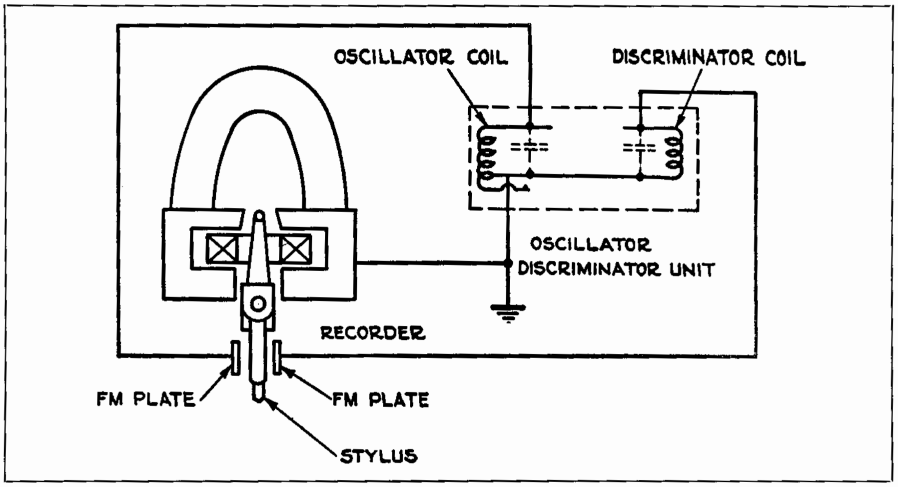


Fig. 28-44. Basic arrangement of the FM Calibrator. Small plates on either side of the stylus bar are connected to an oscillator discriminator unit mounted on the corriage. (Courtesy RCA).

must remain constant for constant velocity motion. Constant amplitude at the lower frequencies is of course necessary to prevent overcutting, unless excessive spacing of grooves is resorted to with the accompanying loss of playing time.

Photoelectric Cell Method

The microscope method was improved upon by substituting a photoelectric cell for the eye and having the stylus modulate a light beam being transmitted to the cell. Calibrators of this type have been in use for some years and in general have proven to be accurate and reliable. They do not, however, permit calibration while cutting a disc.

FM Method

The problem of being able to calibrate the recorder under actual cutting conditions was solved by an FM system.[12] Here is a device which is attached to the recorder without requiring much space or adding mass to the moving system, one which does not couple electrically to the driving coils of the recording head, and which can be so arranged as not to interfere with the cutting action of the stylus.

Fig. 28-44 shows the arrangement.* Two tiny plates, one on each side of the stylus shank or stylus bar, insulated from each other and from the recorder are spaced a few thousandths of an inch from the stylus. Neither mass nor stiffness is added to the moving system so there can be no change in its mechanical action. Flexible leads from these plates connect to the oscillator-discriminator unit mounted on the carriage located close to the recorder. Variation of capacitance between the plates and the stylus due to its motion changes the oscillator frequency and tuning of the discriminator.

Monitoring

The FM calibrator was designed primarily for calibrating purposes, but may also be used for monitoring. As such it is ideal when cutting frequency records for reproducer tests. The recorder can be carefully calibrated beforehand and the correct input level for each band determined. Then when cutting the final disc the calibrator may be used as a check on the recording level, making slight corrections if necessary, or if it is

[12] Roys, H. E., "An FM Calibrator for Disc Recording Heads," Broadcast News.

**The FM Calibrator described here was developed by Mr. Alexis Badmaieff, RCA Engineer.*

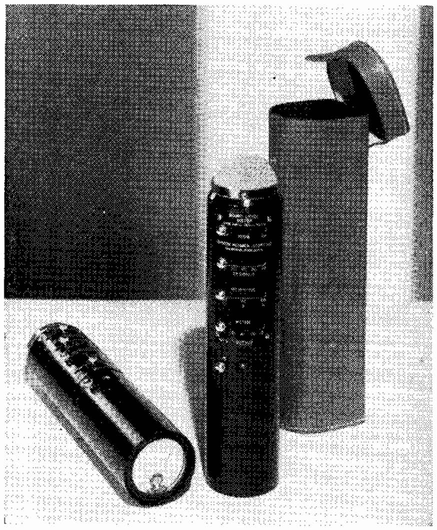


Fig. 28-45. The Scott 410-A Sound Level Meter. (Courtesy H. H. Scott, Inc.).

undesirable to change the level during recording, the correction can be noted and applied afterwards when using the disc.

Measurement of Sound

A sound-level meter which complies with the *American Standard* Sound-Level Meters for Measurement of Noise and Other Sounds, Z24.3-1944, should be used for measuring sound level. Fig. 28-45.

The sound-level meter measures weighted sound pressure level and the results are expressed as "sound level" on a decibel scale, the zero of which corresponds to a sound pressure of 0.0002 dyne per square centimeter at 1000 cycles. This reference point is approximately the minimum sound pressure that would be audible in a very quiet room to an observer having acute hearing. A sound level of 120 db is approaching the level where a sensation of feeling is produced and as the level is still further increased the sound becomes painful. Fig. 28-46 shows the approximate sound levels for a number of commonly encountered conditions. For example, a very quiet residence has a sound level of about 33 db, an average office about 57 db, and very noisy factories 90 db or more. It may also be useful to know that for levels above 40 db, a sound level change of from 6 to 9 db (depending on whether the noise is complex or essentially single frequency), corresponds to approximately doubling or halving the loudness sensation. At levels from 0 to 40 db, the loudness of the sound is doubled for an increase in sound level of from 4 to 7 db.

The sound level of a sound is purely an objective measurement. The results of subjective measurements are expressed as the loudness level of a sound in phons where the loudness level is equal in magnitude to the sound pressure level of the 1000-cycle tone which is judged by an adequate sound jury to be as loud as the sound. For the case of single-frequency tones, the relations between sound pressure level and loudness level have been determined empirically and are shown as a series of contour curves in the *American Standard* for Noise Measurement, Z24.2-1942. Since these contours vary materially as a function of level, three of these contours were selected to cover the range for measurement purposes and these have formed the basis for the response frequency characteristics built into the sound-level meter.

Three frequency response characteristics are provided in the *American Standard*, Z24.3-1944. Curve A approximates the 40-db loudness contour, Curve C is a flat response for higher levels. Whenever sound-level measurements are made, the weighting used, *i.e.*, 40, 70, or flat, should always be reported. Due to limitations in microphone design, uncertainties in their calibration, and variations in individual microphone and amplifier characteristics, deviations from the design objective responses given in the Z24.3-1944 Standard nearly always occur in practice. Allowable deviations in response are given in detail in the Z24.3-1944 Standard from which the six

(Db Referred to 0.0002 Dyne Per Square Centimeter at 1000 Cycles)

Indoor Sounds	Sound Levels	Outdoor Sounds
	120	Airplane: Motor 1600 rpm, 18 ft from Propeller
	110	
Boiler Factory Subway-Local Station, Express Passing		Riveter at 35 ft
	100	
Very Noisy Factories		
	90	
Inside Car of Moving Elevated Train		Heavy Street Traffic with Elevated Line, N. Y.
Very Noisy Restaurant Stenographic Room Average Factory	80	Average Motor Truck Passing at About 30 ft, N. Y.
	70	Average Automobile Passing at About 30 ft, N. Y., or Ordinary Conversation
Information Booth in Large Railway Station		
	60	
Moderate Restaurant Clatter or Average Office		Quiet Automobile
	50	
Quiet Office Average Residence		Minimum Instantaneous Noise in Mid-N. Y. at Night
	40	
Very Quiet Residence		
	30	
		Average Whisper, 5 ft
	20	
		Quiet Whisper, 5 ft, a Rustle of Leaves in a Gentle Breeze
	10	
	0	

Fig. 28-46. Sound levels of common sounds. (Courtesy American Standards).

illustrative values given below were taken:

Frequency—cps	*Allowable Deviations*
25	+ 6, — 9.5 db
60	± 3 db
600	± 2 db
1000	± 2 db
3000	+ 4, — 5.5 db
8000	+ 6, — 9.5 db

The method of calibrating, set up in the *American Standard* Sound-Level Meters for Measurement of Noise and Other Sounds, Z24.3-1944, does not permit a sound-level meter response to be consistently high or low over the whole frequency range by the amounts given in the above table. However, deviations as large as these may occur over narrow frequency ranges. Therefore, it must be borne in mind that allowable differences of up to 6 db may occur in the readings of two meters, on a noise, the sound level of which is controlled by a single-frequency component, or a narrow band of components lying in the region of 60 to 1500 cps. Above or below this frequency region the difference might be greater. For very complex noises much smaller differences would occur, of the order of ± 2 db.

Sound-level meter readings should be representative of the levels existing under operating conditions. If the sound level produced by a piece of apparatus is high near the apparatus, but is considerably attenuated under normal operating conditions at the place where relative quietness is desired, it may be desirable to make measurements at the latter place.

Because of the fact that sound-level meters are designed with three frequency response weightings for different loudness levels, it is frequently difficult on many low-frequency machinery noises to get consistent readings at changeover points. Therefore, machinery manufacturers have found it desirable, in many cases, to use only one weighting for their apparatus noise measurements. This weighting is specified in the specific test code applying to their apparatus and should be followed. When such specific codes are not available it is recommended that the 40-db weighting be used for sound levels up to about 55 db, 70-db weighting for sound levels from about 55 db to 85 db, and the flat weighting be used for higher levels. Where low-frequency noise predominates and large differences are observed between the results obtained using the 40-db and 70-db weightings, it is recommended that the 40-db weighting be used for sound levels up to about 45 db, 70-db weighting for sound levels from 65 db to 75 db, and flat weighting for sound levels above 90 db. For sound levels between 45 db and 65 db, obtained with 40-db and 70-db weightings, a more representative determination of the sound level can be obtained by averaging results obtained with these weightings. Similarly, for sound levels between 75 db and 90 db, obtained with the 70-db and flat weightings, more representative values can be obtained by averaging results obtained with these weightings.

The relation between loudness level and sound level as measured with a sound-level meter is quite complicated. For single-frequency tones, or noise in narrow frequency bands, the two types of levels would be numerically equal if the theoretically correct response curve were available in the sound-level meter, and are approximately equal if the averaging method described previously is used. For complex sounds the loudness levels in phons as judged by a sound jury will in general be numerically greater than the sound levels in db as measured by the sound-level meter. For moderately complex sounds in which the controlling energy at any particular level lies within two or three octaves, the difference will be in the order of 5 to 10 db, while for very complex noises in which components of approximately equal magnitude are rather uniformly spaced over the whole audible frequency range, the difference will be in the order of 5 to 15 db. In general, a moderate change in the level of sound, produced by amplification or attenuation, will produce a change in sound level which will approach very closely a corresponding change in loudness level in phons. This assumes that proper weightings are used in sound-level measurements.