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THE APPLICATION OF ELECTRONICS TO THE PIANO

By
Benjamin F. Meissner

INTRODUCTION

The music of Nature, in her rippling brooks, her sighing breezes, her crashing surfs or howling gales, is as old as earth itself; the rhythm of electrons, of solar systems and galaxies proceeded from the beginning, if any there was, and goes on and on to the end -- if any there be. Of man-made music, the human voice is without doubt the oldest instrument of them all. Whether the twanged bow string, the wind-blown reed, the fall of a stick or stone, or the thump of a tom-tom came next in instrumental music, no one really knows. All have been handed down through the ages, with a new touch now and then, and all are to be seen in the modern orchestra. There we find plucked strings, bowed strings, struck strings; wind instruments of wood and brass; percussion instruments of stretched skin, tuned bars of wood and steel; untuned cymbals and gongs.

In the great music of large orchestras we find them all weaving together in a magnificent moving tapestry their variegated colors and harmonies and rhythms. Here, like actors in a play, whether sweet or blatant, dull or grotesque, soothing or harsh, agile or ponderous, they enact their parts on the musical stage.

Of all these instruments, there are some much more expressive than others, capable of a wider range of tone color, or character, or power, and thus better able to portray a wider range of mood or feeling. The violin family, for example, is extremely expressive; the artist has a degree of intimacy with it perhaps unequalled in any other instrument. He has control of pitch, power, duration, and timbre also, to a considerable extent; he can start or stop a tone, pass from one to another, and from one power to another, pluck or bow his strings in many different ways. He can produce tremolo, glissando or vibrato effects; he can even play two tones at once --- thus exceeding in this respect the abilities of the human voice.

Berlioz, discussing the great value of the bowed strings in an orchestra, says: "From them is evolved the greatest power of expression and an incontestable variety of qualities of tone; violins, particularly, are capable of a host of apparently inconsistent shades of expression. They possess, as a whole, force, lightness, grace; accents, both gloomy and gay; thought and passion. The only point is to know how to make them speak."

However, as Berlioz suggests, this very intimacy between the artist and his instrument makes it a very difficult one to play well. No instrument can sound so unpleasant as a violin or cello when inexpertly played. Further, their inability to sound more than two tones together is a severe limitation for unaccompanied performance. In melody they are superbly versatile; in harmony they are severely limited. The difficulties of bowing and fingering, for quality and pitch control of string instruments, was met by the introduction of instruments with many strings, like the harp and cymballum. Here the pitch intervals were exactly predetermined by tuning one string for each pitch, and the tone quality was also largely predetermined by design characteristics. Later, keyboards and actions were introduced which provided an easier playing technique. A great variety of these keyboard string instruments were introduced, a goodly collection of which may be found at the Metropolitan Museum in New York. Probably the best collection in the world is at the Deutches Museum in Munich.

The earliest string instruments with keyboards used the plucking or scraping type of action mechanism. Of these the Clavichord and Spinet came into extensive use. The more elaborate Harpsichord followed with multipedal control, for octave shifting - coupling arrangements, etc. These instruments used very small strings at relatively low tension, as in zithers, harps, etc. Their tones may be characterized as thin, delicate, and low in power. Dynamic control was not possible.

THE PIANO

When, in the year 1709, the Italian, Cristofori, invented the hammer action, a greatly superior control was provided, enabling the player to regulate tone power by the strength of key blow. The Italians and French called this instrument a "pianoforte," meaning soft-loud; the Germans called it a "Hammer Klavier."

The history of the instruments which preceded the piano shows plainly that the invention of the hammer action was merely the culmination of a long series of efforts on the part of many great craftsmen through three centuries, looking towards the production of a musical stringed instrument capable of doing for domestic use what the organ had always done in the church, namely, to furnish complete command over all existing resources of harmony as well as melody. I may add that the modern organ, and to a lesser extent, the piano, do a great deal more than this. While the piano has no true sostenuto like the organ, it does provide individual tone power control which organs do not have. The "touch-responsive" piano action greatly expands the player's control. The organist, however, has control of a great variety of tone colors, available at the touch of a stop, which he can use singly or in combination, by couplers, or by use of several keyboards. This is the real source of the beauty of organ music; without changes in registration, its performance becomes quite monotonous. The organist also has the swell pedal for mass power control over all tones together; this compensates to some degree for his lack of power control over individual tones provided by Cristofori for the pianist. The piano's touch-responsive keyboard does a great deal more than provide piano to forte control of volume. It provides the very important change of tone quality with volume possessed by all of the very expressive instruments, such as the bowed string types, or the human voice. These, when played piano, are soothing, sonorous, pleasing. At forte they become strident, even unpleasant. The physicist would describe this as a shift of energy from the lower to the higher partials. The radio engineer would compare it to overloading an amplifying tube.

The touch responsive piano, therefore, provides simultaneous, not separate, control of tone power and quality together, a decided improvement over the organ, which has control over power alone, and that only a mass control. The piano, particularly in the lower and middle registers, like the lion, will sigh or snarl or shriek, depending on whether you pet or punch or pound it.

COMPARISON OF PIANO WITH ITS PREDECESSORS

In Figure 1 are shown the various harmonic compositions of the piano and some of its predecessors. In all of these spectrograms the fundamental frequency was 128 cycles, an octave below middle C. The amplitudes have all been reduced to a common level so that

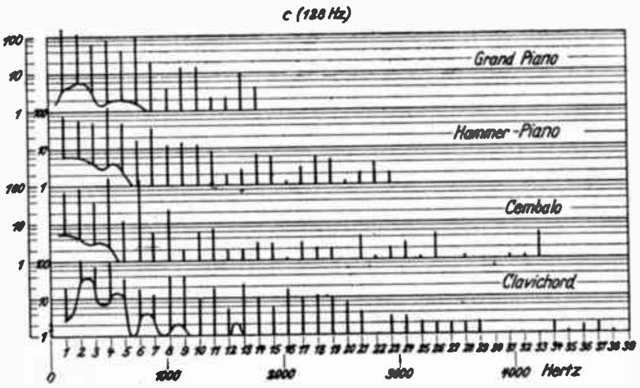

Fig. 1· Spectra of older pianoforte-like instruments.

FIG. 1
SPECTROGRAMS SHOWING LINE AND CONTINUOUS SPECTRA OF THE PIANO AND ITS PREDECESSORS.

this factor does not appear. The Clavichord, with a combined striking and scraping action mechanism, exhibits the thinnest, most delicate tone; today however, it might be termed "tinny." Next, the Cembalo, in the spinet - harpsichord family, with quill-plucked strings, has a tone less bright or thin than the Clavichord. The Hammer-Klavier, such as Chopin and Beethoven used a hundred years ago, gave a tone somewhat duller or darker still in quality. The modern grand tone is still darker, less strident or thin than the others.

In Figure 2 are shown seven spectrograms giving the harmonic composition of a grand piano on seven tones in octave intervals; notice the increase in harmonic richness from high to low tones; notice also the strong, continuous, or noise spectrums, shown by the continuous lines, produced by the hammer blow at the higher frequencies.

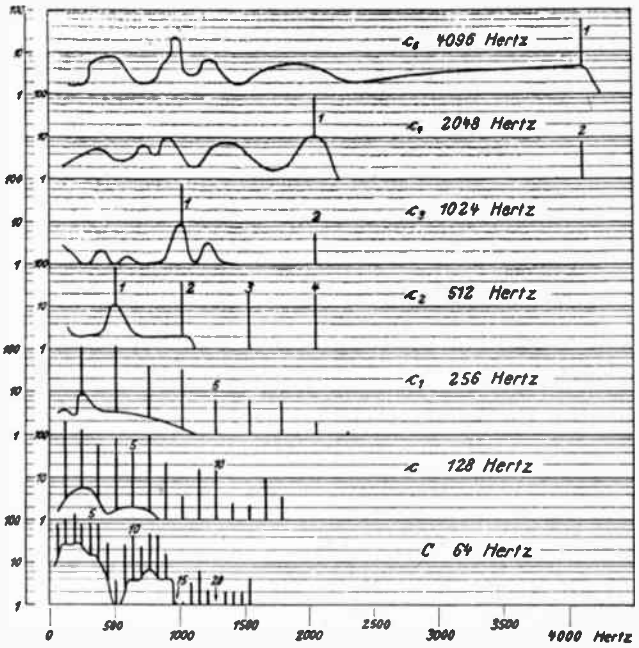

Fig. 2 . Spectra of a grand pianoforte.

FIG. 2
SPECTRA OF PIANO TONES IN OCTAVE INTERVALS.

EFFECT OF KEY BLOW

In Figure 3, I show the influence, on harmonic composition, of hammer velocity, as governed by strength of key blow, on the modern grand piano.

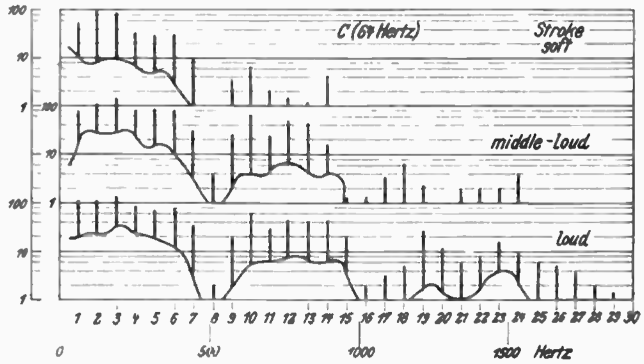

Fig. 3.Influence of the strength of stroke upon the spectrum.

FIG. 3
VARIATION OF HARMONIC COMPOSITION OF A PIANO TONE WITH FORCE OF KEY BLOW.

The top spectrum was produced playing softly, the middle, moderately loud; the bottom very loudly. Notice how the partials come in at higher and higher frequencies with increasing key blow; not shown again is the accompanying increase in general loudness, all three spectrograms having been reduced to a common amplitude level. The curves show the continuous frequency bands and relative amplitudes of the noises produced by the hammer blow, and the broad resonance effects of the soundboard. These sounds are induced in the soundboard by the shock excitation of the hammer blow and continue, with quite rapid damping, (fading) after the shock has passed. Those partials lying within these resonant ranges are, of course, amplified, as shown by their greater amplitudes at the peaks of the continuous curves.

MODERN PIANOS NOT MUCH IMPROVED

The piano of today is, with improvements in detail only, the pianoforte of a hundred years ago; no really significant change in principle has been made. Chief among the improvements is the perfection of steel wire with regard to tensile strength. Higher and higher tensile strengths have provided two significant improvements. First, the change from stretched string plus bar-like vibrational characteristics, to almost wholly string characteristics, and thus gradually wiping out the inharmonic partials of the former; and second, the increase of storage capacity for kinetic energy, which permits transfer of more and more of the energy imparted by the key blow, through the action and hammer, to the string. This improves both the beauty and the power of the tone. It permits, according to the choice of string to soundboard coupling, either a much louder tone, a much longer tone, or, as utilized in modern instruments, a moderate increase in both loudness and duration.

ADVANTAGES OF PIANO OVER OTHER INSTRUMENTS

To summarize then, the great popularity of the piano is due, first, to the ease with which perfect pitch intervals and perfect tone quality may be secured. The veriest tyro can play a series of pitch intervals just as precisely as the greatest artist; second, without the slightest difficulty, he can produce just as beautiful a quality in each tone; third, he can control the power and quality of each tone with the touch-responsive string vibrating mechanism; fourth, he can play, unaided by others, harmony as well as melody; this factor alone accounts for the great popularity of the piano.

While the piano is, undoubtedly, a great and beautiful musical instrument, there are many signs of long-desired improvements in various directions. Chief among these is control over the character or qualities of its tone.

TONE CHARACTER

Tone character, that is, the characteristics which so markedly distinguish the tones of different musical instruments, one from the other, may be separated into several groups. The most important of these are:-

1) *Harmonic Composition:* This may be expressed as the amplitude ratios of the various harmonically related partial or component simple tones in the complex sound. It determines tone quality or timbre, as was illustrated in Figure 1. In many instruments this changes with the amplitude of the driving force. This was shown for the piano in Figure 3. It may also vary during the tone production at constant amplitude in some instruments, depending on the manner of excitation, as in the bowed strings. In the piano it changes constantly from beginning to end of the tone. There may also be inharmonically related partial tones present --- as in bells, or bars held at one end, or strings under insufficient tension.

2) *Dynamic Characteristics:* This refers to the power characteristics of the tone generally, but also more particularly to its relative power from beginning to end. The clarinet or oboe stops of a great organ may produce very close approximations to the true timbres of those instruments, but there is a very marked difference in power. But, more significantly, the shape of the tone wave train envelope may take on many different variations and result in many different tone characters. For example, the chief difference in character between a piano tone and that of a saxophone or clarinet is that the former starts loud and dies away rapidly, while the latter starts at zero, builds up rapidly, and continues, ordinarily at uniform loudness, except as modified by the player's blowing technique. In Figure 4 are shown two oscillograms representing these two types of dynamic inception character.

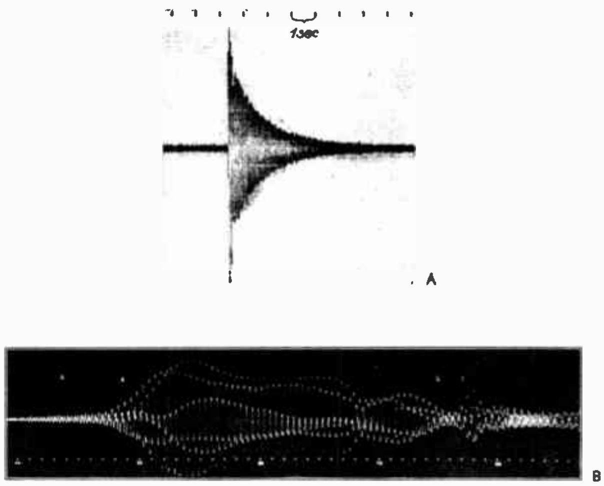

FIG. 4
ENVELOPE SHAPE IS A STRONG DETERMINANT OF TONE CHARACTER. (A) PIANO TONE; (B) WIND INSTRUMENT TONE.

A shows a piano tone with abrupt inception and immediate logarithmic decay; this character is typical of all percussion and plucked string tones. *B* shows the gradual inception of the letter *a* as spoken in the name Raleigh; this is typical of all wind instruments, of which the human voice is a perfect example. A banjo tone may have very nearly the same harmonic composition as a piano tone but differs very noticeably from it because of a much higher rate of damping, that is, rate of tone power diminution. The early phonograph records of pianos sounded a great deal like banjos and guitars principally because the high audibility threshold of the phonographic apparatus brought out only that part of the piano tone occuring in the first few seconds, and thus greatly increasing its apparent damping; and also, of course, because of the omission principally, of lower frequencies. The dynamic envelopes of the many partials may all have different forms. For example, in the piano, the higher partials die out much quicker than the lower.

3) *Incidental Noises:* In the percussion and plucked string instruments, in addition to the always falling dynamic characteristic of the tone, there are also significant incidental noises at the beginning of the tone. In bowed string tones, the bow-scrape noises are always present more or less and are as much a part of the tone character as the distribution of energy in its component partials, or as the shape of its dynamic envelope. In Figure 5, I show an oscillogram of a violin tone. Here is plainly a high frequency tone, superimposed upon one of low frequency. Clearly, the low frequency is the rate at which the string clings to and releases from the bow. The high frequency is that of the string itself. Both, of course, are complex in composition.

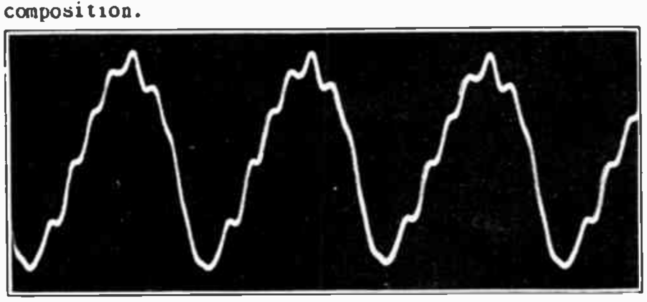

FIG. 5
OSCILLOGRAM OF A VIOLIN TONE

CHANGES IN FREQUENCY AND AMPLITUDE

4) There are also other distinguishing characteristics such as tremolo and vibrato, known to the radio engineer respectively as frequency and amplitude modulation. Another is frequency or pitch range; with regard to the latter, a flute and a tuba might have very similar characteristics in every respect yet no one could fail to distinguish them owing to their very different frequency ranges.

These are the principal elements found in the make up of tone character. I will make no attempt to break them down into finer subdivisions, combinations or subcombinations.

With such a set of ingredients it is possible to set together musical tones of practically any desired character; with suitable control over the utterances of these characters, it is further possible to construct and correlate them in the various melodic, harmonic, dynamic, and rhythmic patterns, which we call Music.

DESIRABLE PIANO IMPROVEMENTS

As I have previously stated, there are many directions in which the piano may be greatly improved. In the first place, its single type of tone quality, when compared with the many tone qualities of the organ or the orchestra, presents a very great limitation; secondly, the piano is always a percussion instrument; its tone always starts with a crash of percussion transients and then dies rather rapidly away, logarithmic fashion, with more or less continuous change in harmonic composition. The strong percussion noise and rapid fading out is especially noticeable in the upper registers. In all the wind instruments, and, within control, on the bowed string instruments, the tone starts with low amplitude and quickly builds up to normal. Here is a complete reversal of the dynamic inception characteristic of the piano tone; and this reversal, even without the percussion transients, is sufficient to produce, out of a given harmonic composition, two tones of decidedly different character. A very great improvement, then, would be provided with the ability to control the dynamic inception characteristic of the tone, thus producing tones characteristic of percussion or of wind instruments at will.

Just as important as the dynamic inception of the tone, is its dynamic nature after inception. The piano always has a falling tone power characteristic while the organ quickly reaches a steady state and holds it so long as the key is held. Both may be brought to an end with a rapid, though not sharp, falling characteristic by key release. In the piano the damper acts quickly though not instantaneously. In the organ and other wind instruments the vibrating air column continues apace until its stored energy also is dissipated. The ability to predetermine the tone damping, that is, the ability to make it sustain indefinitely, without damping, and to provide any desired degree of damping, constitutes also a very important control of tone character. While I have not incorporated in this instrument arrangements for producing indefinitely sustainable or undamped tones, that, with some increased complication, is possible. However, I do have arrangements for increasing the normally quite low damping rate of the tones.

The ability to inject certain incidental noises, to provide tremolo and vibrato effects, at will, further adds to our control over tone character.

Here then you have my conception of the advantages, the limitations, and the very desirable improvements for the most popular of all home musical instruments, the piano. To keep the most important advantages, to efface the worst limitations, and to add the best improvements then, was the problem which, some years ago, I set myself to solve. The instrument to be demonstrated later, represents in a general way the results of my efforts to date. It is not a commercial design, nor does it embody all of the many improvements developed in my laboratory. My efforts thus far have been devoted rather to an exploration of the possibilities and the development of a variety of solutions for the many individual problems than to a combination of them suitable for commerce. It is an operating model, in which some very significant improvements are included; in which some, perhaps equally important, are not yet included. While some of you might be interested in following the course of its technical development, that, obviously, time does not permit.

THE ELECTRONIC PIANO

Now I will proceed with a technical explanation of the operating principles and the instrument itself. What you see here in Figure 6 is essentially although not quite a piano. It has a conventional body or case,

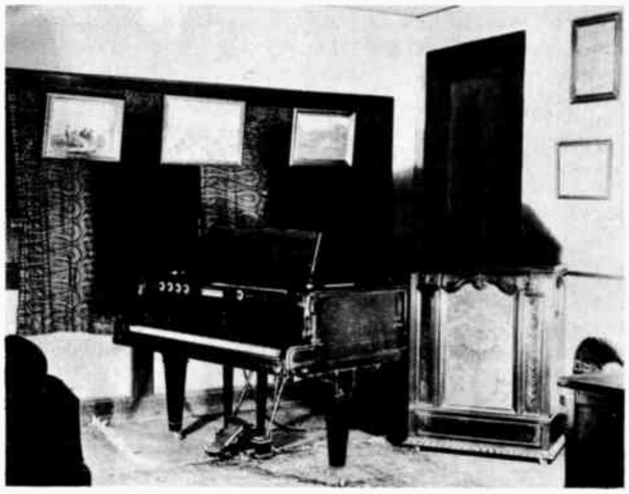

FIG. 6
THE MIESSNER ELECTRONIC PIANO

an inner iron plate strung with strings over a vibratile bridge; tuning pins for tuning the strings; a keyboard and hammer and damper action for selectively setting the strings into vibration, and for stopping that vibration. However, it has undergone some rather important changes. You note a number of control knobs and buttons above the keyboard; a swell pedal in place of the unicorde or soft pedal; an additional cabinet containing an amplifier and reproducers.

In Figure 7, showing the interior of the piano, you

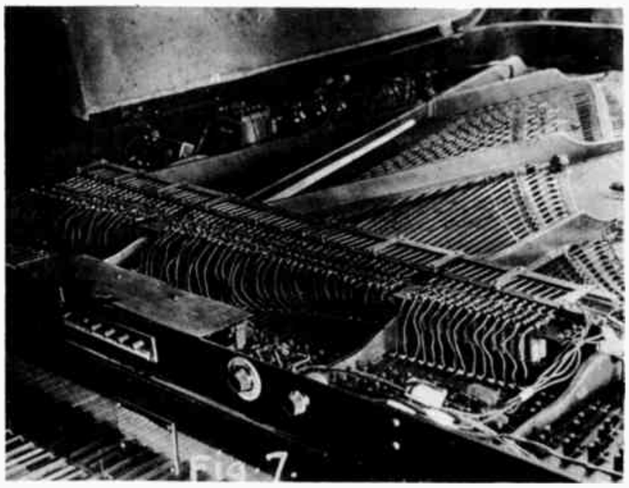

FIG. 7
INTERIOR OF ELECTRONIC PIANO

see many more changes. You see through the strings, in the right middle of the picture, that the soundboard has been cut away, leaving the reinforcing ribs; you see a small amplifier, and additional apparatus, which I will later describe more in detail. The soundboard has been cut away from its supporting ribs in order to retain the benefits of inter-string coupling, and to greatly reduce their coupling, through the soundboard, to the atmosphere. While often called a resonator the soundboard is, of course, really no such thing. It is simply a coupling device, with its action only very broadly resonant. It is certainly no more a resonator than the diaphragm of a loud speaker. The piano string has altogether too small a surface for efficient transfer of its vibratory energy into motions of the surrounding air. This purely mechanical soundboard device is therefore used between the two, and incidentally with a step-down action of about 1,000 to 1, so that the large motion of the small surface string is converted into a small motion of the large soundboard.

Since we wish to use a different, far more flexible and controllable system for translating string vibrations into sound waves, and since we wish practically complete control of all translation, we have cut away the soundboard. If this had not been done many of the effects we wish to secure would be blanketed by the normal piano sound sent out from this board. Further, the small residual sound emanating from the coupling between air and strings and ribs, and other vibratory parts of the structure as a whole, is absorbed by internal padding, or hindered in its escape by closed bottom and top. A careful, wholly new design would permit a still greater reduction in this purely mechanico-acoustic coupling. On the other hand, this weak direct sound is useful for certain tonal effects or for practise at low volume.

These changes are merely preparatory to the introduction of the new electro-acoustic translation system. They have not materially affected the rest of the instrument, except to reduce the rate at which the string declines in its vibration amplitude, because of the diminished hysterisis and radiation losses of the vibratile system.

MECHANICO-ELECTRO TRANSLATING SYSTEMS

There are numerous principles upon which a mechanico-electro vibration-conversion system may be based, for replacement of the soundboard. Among these may be mentioned modulations by the vibrating string of a magnetic or an electrostatic field, or a resistance; or operation of a piezo-electric, electro dynamic, magnetostriction, or other such device. In Figure 8, I show schematic representations of these string vibration conversion devices.

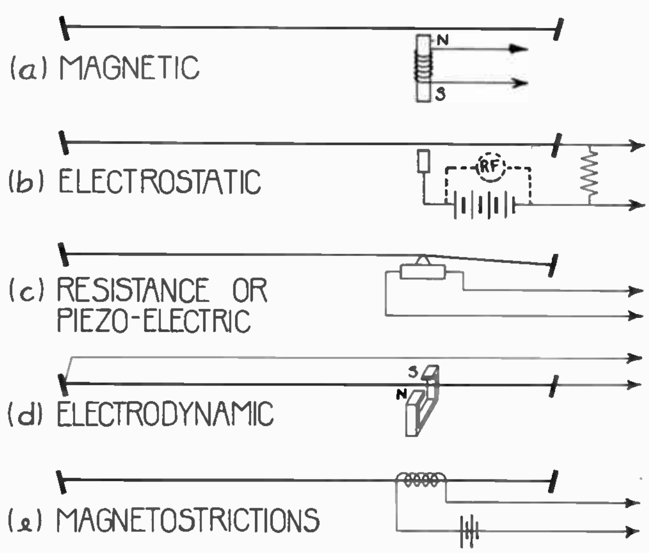

Fig. 8
TYPES OF TRANSLATING APPARATUS.

In (a) the magnetic field of the magnet threading the coil is modulated by vibrations of the steel string, thus producing a voltage vibration in the coil; if the string is magnetized the coil core need not be. In (b) the string vibration modulates its capacity to the pick-up conductor, and thus modulates either a steady voltage or a high frequency current. In (c) the string bears down with some pressure on a resistance or piezo-electric device and, as it vibrates, this pressure is modulated, thus varying the resistance, or generating a voltage in the piezo crystal. In (d) the string is a conductor vibrating in a magnetic field and thus has voltage generated in itself. In (e) the string is magnetized and its longitudinal extensions so disturb its magnetic field that a voltage is generated in the coil.

Of these the electrostatic principle appears simplest and most generally useful for the particular requirements at hand. As applied to my piano, a thin, narrow conductor is mounted near and underneath the strings, on a well-insulated support, as shown in Figure 9. You see three of these, a, b and c. Its insulation resistance should be of the order of 100 meg ohms. In general it is just far enough from the strings to prevent contact on the most violent vibrations which the strings undergo. When a voltage is applied thru a high resistance, between the strings and this conductor, vibratory variation of the distance modulates the capacity; since the total charge is fixed, by the high resistance, except for extremely low frequency modulations, the voltage across the capacity is modulated also; in its electrical operation here it resembles the familiar condenser microphone, or perhaps numerous condenser microphones connected in parallel, each with a tuned diaphragm. Note the cut-away soundboard again, the insulation between strings and plate, the voicing screws on the bridge for lowering or raising the strings with respect to the pick-ups.

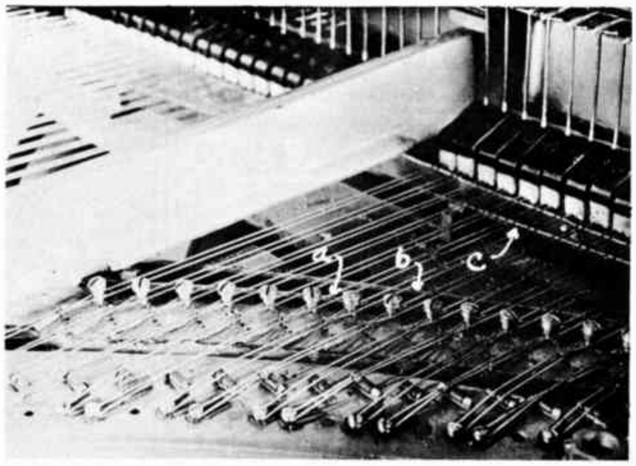

FIG. 9
DETAILS OF PICKUPS, VOICING SCREWS, AND STRING INSULATION

TYPES AND ARRANGEMENT OF ELECTROSTATIC PICKUPS

The shape, the position, and manner of mounting of this "pick-up" are of considerable importance. In Figure 10, I show some of these possible arrangements. In general, it should be very narrow along the string, unless very high partials are not to be translated. If, for example, it were as wide as the string is long, only the fundamental and a few low-numbered, odd partials would be translated.

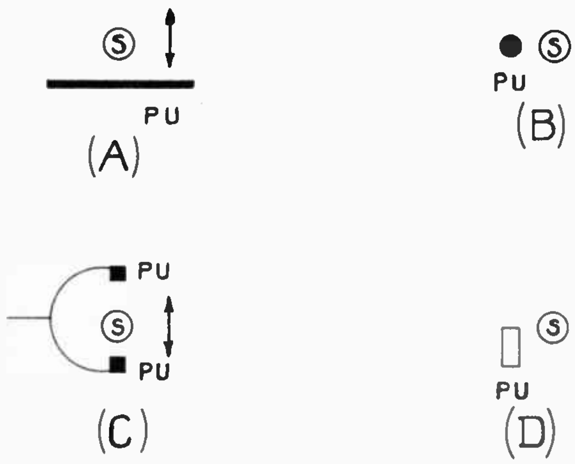

Fig. 10

ELECTROSTATIC PICKUP TYPES.

The depth of the conductor perpendicularly to the string axis, as well as the separation, affects the shape of the translated voltage wave form. The position of the pick-up along the string will affect the harmonic composition of the translated vibration. If at the string middle, only odd-numbered partials will translate, because there the even partials all have nodes of amplitude. At the string ends the higher partials have more nearly equal, or strong, amplitudes compared to the lower; this is reversed toward the mid position. A shows our normal pick-up. If two pick-ups be set in complementary positions about the middle of the string, either the odd or the even partials may be made to neutralize, depending on the relative phases of the pick-up outputs. If pick-ups be placed alongside a string, as shown at B, strong frequency-doubling action is provided; assymetric pick-up, providing a sometimes desirable distortion, is secured by two pick-ups on opposite sides of the string as shown at C; here, if the two spacings are equal, all frequencies will again be doubled; if assymetric, varying degrees of non-linearity in translation may be secured; if a single pick-up be placed obliquely to one side, as indicated at D, a combination of simple and doubling action occurs.

Another matter is concerned with the rigidity of the pick-up device. It may introduce spurious, unwanted frequencies or noises, if it, itself, can vibrate, or if it is attached to some part of the instrument which vibrates differently than the string-bridge system.

I have found that, for general purposes, a pick-up conductor, consisting of flat, one-sixteenth inch, copper braid, glued to the narrow side of a strip of bakelite one-eighth inch thick by one-half inch wide, and securely fastened, conductor-side up, to each of the wooden ribs, is a satisfactory type and mounting. This selection has, however, been rather largely influenced by our desire to secure, among other types of tone, one closely resembling that of the piano.

TONE QUALITY CONTROL BY SEVERAL PICKUPS

A plurality of pick-ups of different form, position along, or position around the string, may be used singly or in combination, still further to change the harmonic composition of the translated wave from that of the string itself. An over-all control, known in radio as a tone control, permitting tipping, about some central region, of the frequency-gain characteristic of the amplifier, provides still further control of harmonic composition. The possibilities for control of harmonic composition are thus seen to be almost limitless, at least for change of the harmonic structure provided by the struck string; and I may say that the string has a wealth of such material to work with. The string itself is certainly a simple, cheap, and rugged source of this harmonic material. Cristofori's mechanism for control of its vibration, while not simple, is certainly rugged and time-tested. In the present instrument, all of these possibilities for tone quality control are not included, and it is possible that a better selection and arrangement might have been made. However, this selected arrangement is simple and effective.

Three pick-ups, of the type previously described and already shown, are used. One of these is very close to the back end of the string, at about one-sixteenth string length, where the high partials are prominent; another is placed at about the one-sixth-from-back position; a third is placed in a position about one-third from the front end. To compensate to some degree for the irregular frequency-response characteristics of loud speakers, the strings are provided with individual pick-up spacing adjustment screws; these are mounted on the bridge and raise or lower the individual strings with respect to the pick-up strip. This adjustment operates much more effectively for the pick-up nearest the bridge than for the other two. A more perfect, though more complicated arrangement would provide fixed position strings and adjustable pick-ups for each string on each bar.

CIRCUITS OF THE ELECTRONIC PIANO

Before proceeding to detailed description of the various elements of the electronic system in the piano, I show, in Figure 11, the chief component parts and their arrangement. The piano key, K, thru the action, A, hammer, H, and damper, D, controls the mechanical vibrations of the string, S. The string charging circuits, coöperating with switching devices in the piano action --- control the polarizing charges on the strings.

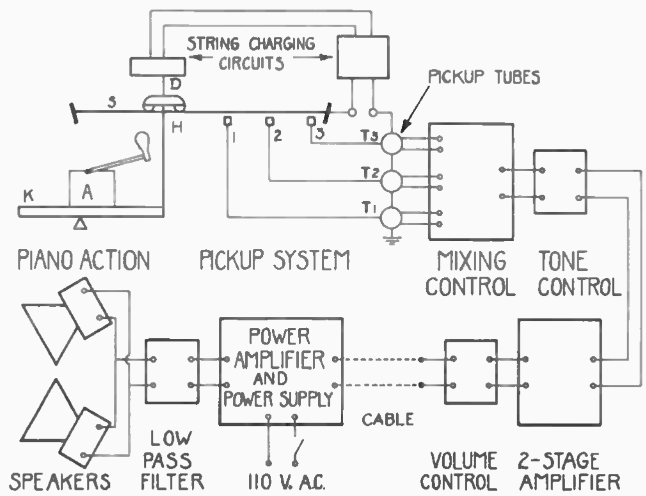

Fig. 11 GENERAL ARRANGEMENT OF ELECTRONIC PIANO CIRCUITS.

Pick-ups 1, 2 and 3, feed into individual pre-amplifier tubes T_1, T_2 and T_3 respectively. These tubes feed into a mixing circuit wherein the relative amplitudes of the amplified pick-up voltages may be varied in aiding or bucking phases. Following this is the tone control circuit for adjusting the frequency-amplification characteristic of the first stage of the following two stage amplifier. The second stage of this amplifier is a push-pull driver for the power amplifier. Next we have

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Annapolis, Maryland

18 January 1934

THE MASQUERADERS of the U.S.Naval Academy wish to announce the presentation of "THREE CORNERED MOON", on the following dates:

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Evening, 9 February
Matinee and evening- 10 February

1. The evening performance of Friday, 9 February, has been designated as "Officers' Night". Officers and civilians may attend any performance but should attend the one on the 9th if possible.

2. Please indicate below the number of tickets desired and for which performance.

3. Subscription is 85¢ per ticket. Make checks payable to "The Masquerader Fund."

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Enclosed please find money order / check / cash for ____________ dollars

as payment for ________ tickets for evening performance on

9 February 1934. (THE MASQUERADERS).

3 February 1934

10 February 1934 ; matinee; evening

(Indicate which performance by check mark)

a volume control operated by a swell pedal. All of the preceding apparatus is in the piano. The remaining power amplifier, power supply, low-pass filter and reproducers, are in a separate cabinet, connected to the piano by a multicable and plug.

OPERATION OF MIXING CIRCUIT

Referring now to Figure 12, the three pick-up strips connect to the grids respectively of three amplifier tubes. These tubes have output transformers 14, 14a and 14b, with parallel mid-tapped potentio-meters, all connected in a series mixing circuit so that any desired amplitude in either aiding or opposing phase may be used from either of the three pick-up tubes. This mixing circuit then feeds into a tone

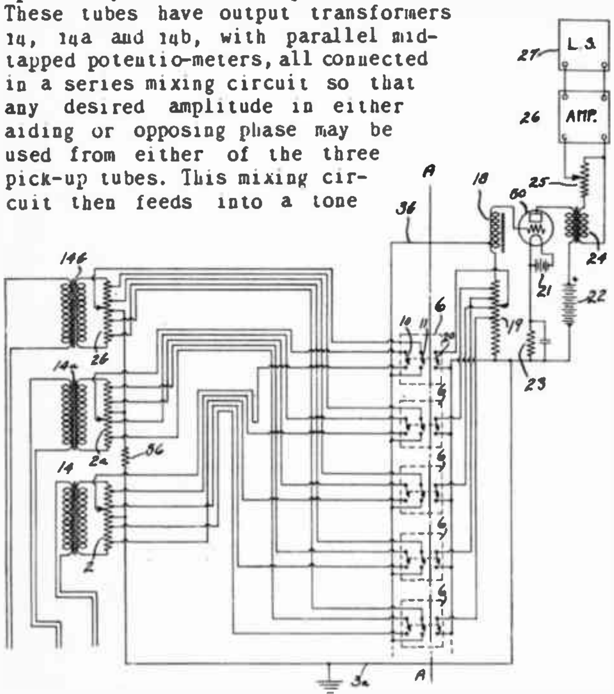


FIG. 12
TIMBRE-CONTROL CIRCUITS.

control circuit, so arranged that tipping of the voltage-frequency characteristic is provided. By means of this latter control the frequency characteristic may be made rising, flat, or falling, as desired. For these mixing and frequency-characteristic controls two interchangeable types are provided. One of these is a manual, or fine control, by continuously variable potentiometers; the other, a preselected type using multiple push-button-jack-switches and step-by-step potentiometers. Either manual or preselected control type may be switched into operation by another multipole, button-operated, jack switch. These switches are fitted with an interlocking mechanical arrangement so that only one may be set at a time; when a new button is pushed the old one releases by this single motion. A simple comparison will illustrate this arrangement perfectly. A painter may mix any desired visual hue by proportioning his three primary colors, red, yellow and blue, and then brightening or darkening it by additional use of white or black. Or he may use pre-mixed pigments in a great variety of hues. The primary acoustic color sources here are the three pick-ups; the brightening or darkening is obtained with the tone control. For delicate, in-between shades both painter and pianist would use the manual mixing method; for obtaining ordinary colors they would use the easier pre-mixed method.

To make the operation of this tone mixing system clearer, I show, in Figure 13, a schematic representation of a string vibrating with, for sake of simplicity, only three components, namely the first, or fundamental, the third, and the tenth, in amplitude ratios of, say, ten, five and two, respectively. If we place a pick-up at the string middle, A, and just beyond touching distance, it is clear that the translated amplitudes of the odd numbered first and third components will be the same respectively as in the string itself. Notice further, that the even numbered tenth does not translate at all owing to its vibrational node at that point. A

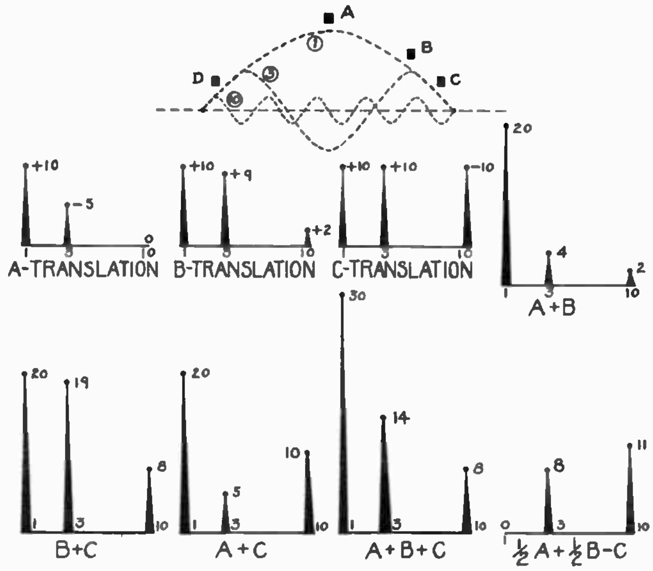


Fig. 13 STRING PARTIALS AND PICKUP SPECTROGRAMS.

pick-up at B, about one-sixth from string end, will give translated amplitude ratios of about ten, nine, and two for the three components. One at C, for the one-tenth position, gives ratios of about one, one, one. Another one-tenth position pick-up, placed on the same side of the string but at the opposite end, D, will have these same ratios but the phase of the even numbered partials will be reversed, while that of the odd will still be the same. If C and D are used in combination, with equal amplification, the odd numbered partials ten will neutralize. Such complementary-position pick-ups, no matter at what fractional string positions, may thus be used to cancel all even or all odd-numbered partials translated from a string, while a mid-point pick-up translates only odd numbered partials.

Let us see now what happens if we combine, say A with B. The fundamental component of A, with amplitude 10, and of B, with amplitude 10, add, making partial one of amplitude 20 on our scale; the translations of the third component, however, are in reverse phase but differing amplitudes. At A it has an amplitude of minus five, while at B it is plus nine; the algebraic sum, therefore, is plus four.

The tenth partial at A is zero, and, at B, two. The same numerical and tonal results are obtainable by combining A and D. Of no consequence is the fact that the relative phase of the tenth partial to the other two would thus be reversed. This has no influence on tone quality.

If we combine B and C together, again with equal amplification, we get another spectrum, as shown in the Figure; shown also is the combined translations of A and C as well as A, B and C together.

Let us now take advantage of phase reversal and of amplitude control of the translation of A, B and C, provided in our mixing system. Take, for example, A's and B's translations at one half amplitude and of the same phase, and combine these with C's translation at full amplitude with reversed phase. We have now eliminated the fundamental altogether, and turned the amplitude distribution completely about with respect to the remaining two partials. There is an endless variety of such mixtures possible with continuously variable control over the amplitudes of the translations of several such pick-ups in different positions and with reversible phase.

OPERATION OF TONE-CONTROL CIRCUIT

But we go yet another step further by applying the control over frequency-amplification characteristic in the amplifier circuit following the mixing system.

In Figure 14, I show two extremes and one mean adjustment of the continuously variable tone control. With it any of the tone spectrums obtainable from the mixing system may be reproduced about as delivered by the latter using A; may be depressed in amplitude in the lower and elevated in the upper frequencies as in C; or may be reduced somewhat in the lower without much change in the higher frequencies as in B.

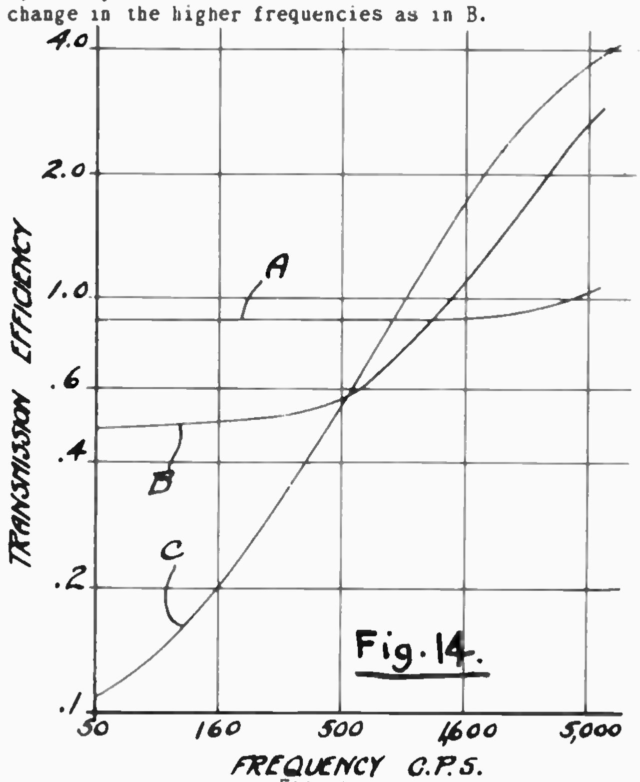


FIG. 14
CURVES FOR THREE ADJUSTMENTS OF TONE-CONTROL POTENTIOMETER

Obviously, by means of these various controls over harmonic composition, a great many varieties of tone quality may be obtained from the relatively fixed quality present in the piano string.

The output of this tone control system, as previously shown, is fed into the first common tube of the amplifier. This tube then feeds into a pair of 59 type tubes in push-pull and used as three-element tubes. Controlling this input is a master gain control potentiometer, used to set the maximum output sound power. These power-amplifier driver tubes feed into a 500 ohm matching transformer feeding into the power amplifier. In this 500 ohm line is placed a constant-impedance attenuator controlled by a foot pedal. With this the pianist has instantaneous control of tone power and with it numerous important musical effects may be secured.

I show the power amplifier, speakers, power supply, etc., in Figure 15. The power amplifier uses push-pull 845 tubes in a Class A circuit. Between these tubes and the speakers, in addition to a matching transformer, is a low pass filter, cutting off at about 6,000 cycles. This filter does not appreciably affect the desired tone quality, but it very greatly reduces those troublesome and unpleasant noises introduced by an amplifier at even subnormal output power. Two self-excited speakers are used, with crossed sound beams, for better distribution of the high frequency portion of the output sound spectrum. The entire amplifier is a.c. operated.

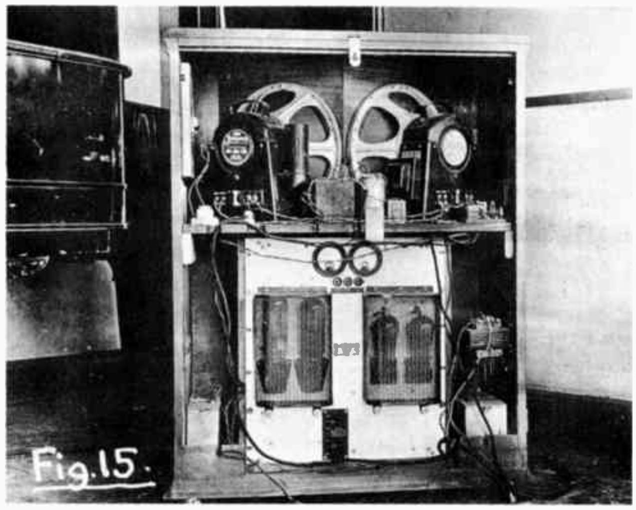


FIG. 15
POWER AMPLIFIER, POWER SUPPLY, SPEAKERS, ETC., IN SEPARATE CABINET.

That part of it up to the power amplifier is included in the piano as a pre-amplifier, the remainder is in a separate cabinet. Connection between them is made by a multiconductor cable and detachable multiple connector. The entire piano is electro-statically shielded; on the bottom by screening, and otherwise by conducting paint made of a solution of colloidal graphite.

TONE CHARACTER CONTROL BY ENVELOPE SHAPE

For control over the tone character by envelope change a number of principles, as shown in Figure 16, are available. For example: to change the dynamic

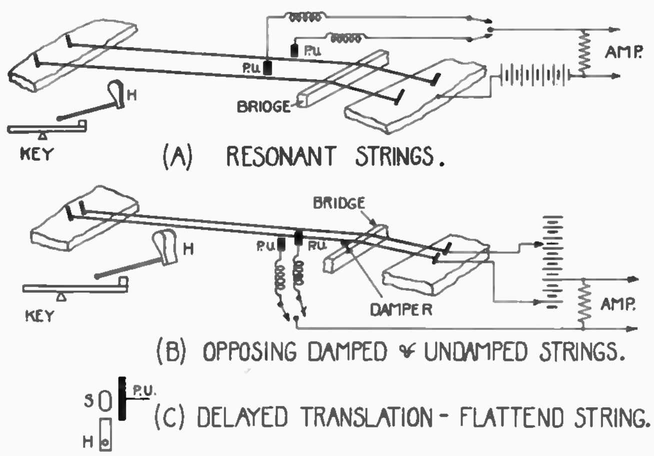


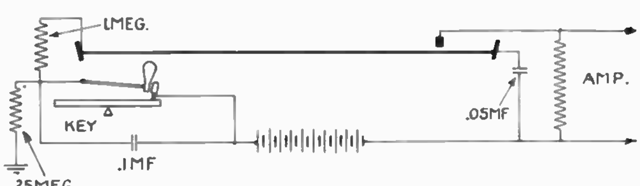
Fig. 16. TONE ENVELOPE CONTROL PRINCIPLES.

character from that of the piano, beginning with percussive transients and falling logarithmically in amplitude, to one without percussive transients and with an initial rising amplitude, like that of wind instruments, we may use the principle of resonance, as illustrated by A. If one string be struck, and if pick-up be confined to a coupled resonant string, this complete change of dynamic character will be provided. For percussion, piano-like, tone, translation may be from the struck string; intermediate effects may be had by combining the two strings' vibrations.

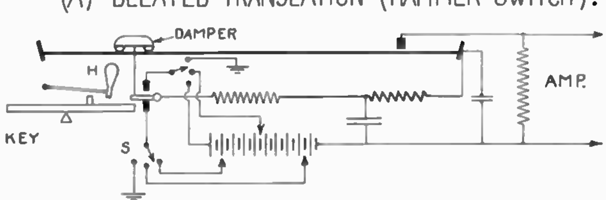
Or, we may strike two similar strings, damp one and pick-up from the two with either mechanically or electrically reversed phase, as shown at B. The initial portions of both neutralize, due to reversed polarities of charge; the undamped string, continuing, as the damped one falls rapidly in amplitude, provide together, the rising amplitude character, devoid of percussion transients peculiar to the wind instruments. Again, other tone-envelope shapes may be secured by translating from the damped string, alone, or from both, with aiding phase.

Another method, illustrated by C, is to use a somewhat flattened string, strike it along the long axis of its cross-section and apply pick-up normal to this striking direction. Such a string is in a condition of dynamic unbalance when first struck. It starts vibrating in the striking direction and then gradually swings about in a direction at right angles. A pick-up of length greater than the maximum over all vibration amplitude, positioned at right angle to the string's axis, and parallel to the string's initial vibration direction, will not translate during this early part of the tone, containing the percussion noises, but will translate efficiently as the direction of the string's vibration swings around normal to it. Thus the wind instrument starting characteristic is again obtained.

We may, in other methods, shown in Figure 17, so influence the pick-up apparatus by delaying its translation efficiency, that it is insensitive to the percussion transients, and so that its sensitivity builds up



(A) DELAYED TRANSLATION (HAMMER SWITCH).



(B) DELAYED TRANSLATION (DAMPER SWITCH).

Fig. 17. TONE ENVELOPE CONTROL PRINCIPLES.

and continues in any desired manner. With the electrostatic translation systems this delay may be secured by use of a switch operated by the key, hammer, damper, or other part of the action. If, as in A, the hammer be used, a contact between it and its back-check will close a circuit after the percussion transients have subsided. If that circuit include a condenser in parallel with the string, necessarily now insulated, and a resistance, in series to the switch and charging voltage source, then the delay is secured by delay in the application of the charging voltage on the string and the consequent delay in the building up of the translation efficiency for that string; this delay may be made as slow or fast as desired, by choice of the time constant of the resistor condenser circuit. In order that the operation may be repeated, the string and its associated condenser must be discharged when the key is released. It may be discharged, as in A, by a permanently connected resistance, across which the charging current maintains the charging voltage, so long as applied by the key-controlled switch; or it may be discharged, as in B, by use of a double-throw switch, which connects the string, through its delay circuit, to the charging voltage, when the key is down, and which, when the key is released, discharges the string and condenser, again through the delay circuit. The inclusion of the delay circuit also in the discharge path is not so much for the purpose of delaying the discharge as to prevent reproduced click noises resulting from too sudden a change in string potential. A condenser across the switch contacts prevents these noises when the switch opens. To further prevent too rapid potential changes, at least one of the switch contacts should be of a type providing decrease of resistance with increase of pressure. A small pad of soft leather impregnated with a dilute solution of colloidal graphite very simply provides this requirement.

Until very recently the present instrument had the hammer and back check type of switch in operation throughout. A detail view of this type of delay action is shown in Figure 18. Here the hammers and back checks,

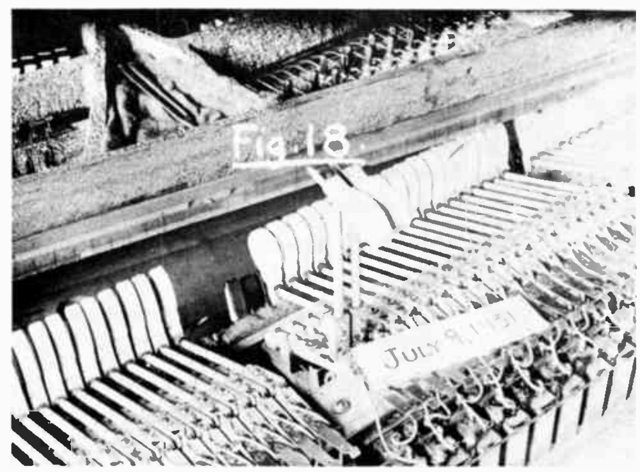


FIG. 18
HAMMER-BACK-CHECK TYPE OF DELAY-ACTION SWITCH.

with aquadagged-leather contact surfaces, connecting wires, and the condenser and resistors of the delay circuits, are shown. This scheme is still in use in the region above that provided with dampers, that is, in the top twenty-one notes. In the rest of the instrument we have recently installed the improved damper type switch. I show this, in detail, in Figure 19. Note the small rods on top of the dampers. These push up the flexible switch poles, which contact in up and down positions, with aquadagged felt pads cemented to brass

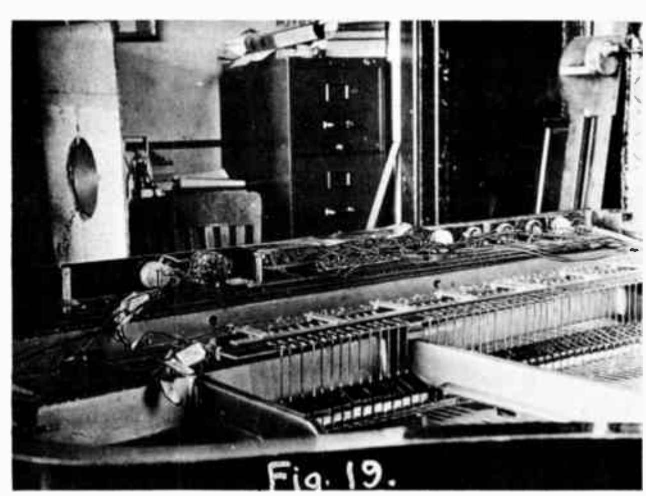


FIG. 19
DAMPER TYPE OF DELAY-ACTION SWITCH.

strips. The spring switch poles connect, through two section filters, to the strings. The brass strips are connected to the main, multipole switch, for control of their connections to the polarizing potential source or ground. Shown here also are the various control and switching connections.

ARTIFICIALLY-INCREASED TONE DAMPING

To reverse the delay action, that is, to permit normal initial translation of the string vibration with its percussion transients, but to artificially increase its rate of decay to any desired degree, the string may be normally fully charged but immediately start discharging as soon as the key is pressed. The rate of decay of translation efficiency may be varied by connecting this discharge circuit to some point on the charging voltage, source lower than normal, such as one third, one half, zero, or even reverse voltage.

There are important musical advantages with this artificial damping control. For example, with some types of music the normal damping of our strings, without a coupled soundboard, may be too low. The effect of this is a muddling or running together of tones in rapid passages. In previous demonstrations of the Electronic this muddling effect has been noticed by some musicians. Now it is possible to adjust this damping, thus providing any desired degree of distinction of individual tones played in rapid succession. The exact degree found in the best pianos can be obtained.

An important additional advantage is obtained with this scheme. The artificial damping control may be made to operate only on the early portion of the tones, and thus leave the remainder to go on with the low damping of the string, thus obtaining the very much desired sustaining power, without increase of muddiness in rapid playing. For special effects, such as harp, banjo, drum, etc., the higher damping rates are used.

The time, translation-efficiency characteristic obtainable with these delay circuits, may be adjusted to different forms as shown in Figure 20. Here (a) shows the zero voltage axis and, above it, the normal, steady-voltage impressed on the strings for full, unmodified, translation of the string vibrations. At (b)

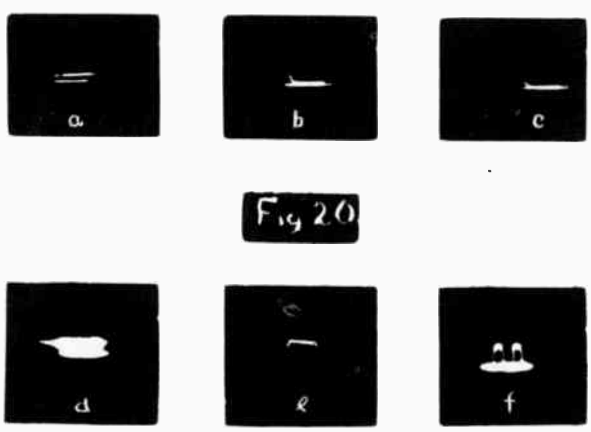


FIG. 20
OSCILLOGRAMS OF STRING-CHARGING VOLTAGE AS CONTROLLED BY DELAY CIRCUITS.

is shown first, normal charging voltage at the left, then the decrease to a moderate degree of artificial damping by decreased translation efficiency. At (c) a quite high artificial damping is provided by a quick drop to a low charging voltage. At (d) is shown the effect of two key blows in fairly rapid succession with a moderate artificial damping adjustment. At (e) is shown the rising characteristic adjustment for wind instrument tones. Here the string voltage is zero until its key is struck. Then the delay action switch connects the string, thru its delay circuit, to the charging voltage source. The smooth rise of the voltage to a steady value in about one fifth second is here shown. At (f) is shown the rise and fall of string voltage, for this wind instrument adjustment, when a note is rapidly repeated by key blow. As soon as the key is released its switch opens the charging circuit and discharges the string and its associated condenser, after which the next key blow repeats the charging action. So long as the key is held, the string voltage remains constant.

In Figure 21, I show some oscillograms of a vibrating piano string (middle C-261.6 cycles) without and with modifications in its envelope shape, introduced by these time-translation-efficiency controls.

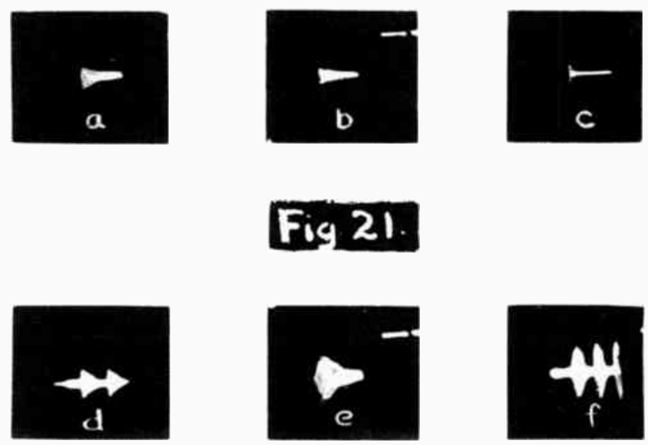


FIG. 21
OSCILLOGRAMS OF TRANSLATED STRING VIBRATIONS AS INFLUENCED BY DELAY CIRCUITS.

The normal, unmodified envelope of this string vibration, as it appears in voltage across the loudspeakers of our electrical translation system, is shown at (a). This does not show the whole tone, because it continued too long for the one-second sweep period of the oscillograph. Notice the abrupt inception and immediate decaying of the vibration at a slow rate. In (b) is shown the same tone with its envelope modified by a slowly-falling time-translation-efficiency characteristic. Observe here the moderate increase in damping of the tone. In (c) a quite rapidly falling characteristic was used, producing a tone of markedly greater damping. In (d) is shown the effect of an artificial damping of moderate degree plus the mechanical damping of the felt string-damper with a rapidly repeated tone.

WIND INSTRUMENT TONES FROM STRUCK STRINGS

In (a) is shown the delayed inception of the tone caused by the rising voltage characteristic adjustment. Comparing this with the unmodified string vibration shown in (a) we see that the abrupt inception of (a) has been changed to a gradual inception in (e), requiring about 1/5th second, thus changing the piano tone into that of a wind instrument.

In (f) is shown a thrice-repeated tone with this same delayed translation efficiency adjustment. The decrease in amplitude at the moment of key release is caused by both decreased translation efficiency and mechanical damper action. In all of these oscillograms showing separate tones the sweep circuit was timed to one second; in those showing repetition action it was adjusted to two seconds.

OSCILLOGRAPHIC APPARATUS

In Figure 22, I show a view of the oscillographic apparatus used in making these oscillograms. A Von Ardenne Cathode Ray oscillograph, o, o, was used with a Dumont Cathode Ray Tube, T, of the time delay type.

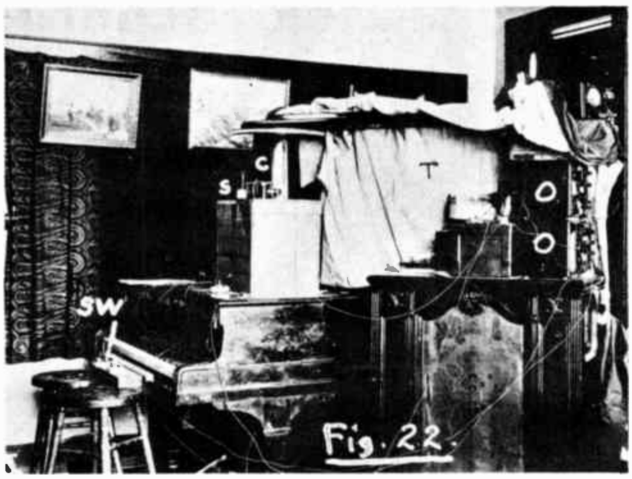


FIG. 22
OSCILLOGRAPHIC APPARATUS SETUP.

A Leica Camera, C, with an electromagnetic shutter-operating arrangement, S, took the pictures. The switch, SW, at left, served to close the shutter-operating magnet circuit, and to start the sweep circuit, when the switch handle struck the piano key. The oscillograph deflecting voltage was taken from a transformer across the speaker voice coils --- for the tone oscillograms last shown in Figure 21; for the voltage graphs of Figure 20 the oscillograph deflecting voltage was obtained from a high-resistance potentiometer between string and ground.

LOCATION OF DELAY CIRCUIT SWITCHES

The most advantageous point in the action, for operation of the delay circuit switch, is in the damper mechanism. With the back-check switch, the organ or wind instrument tones could not be held, after key release, by use of the damper-control sostenuto and sustaining pedals. The reason for this is, that key release opened the string-charging switch, thus discharging the string. Even though the string was allowed to continue in vibration, after key release, by holding the damper up with pedals, this vibration would not translate, due to the lack of translating efficiency caused by string discharge. The organ tone, therefore, could only be held by holding the key down, a technique confusing and limiting to a pianist, who relies very strongly on his sustaining pedal, and, to some degree, on his sostenuto pedal, for sustaining all or given tones after his fingers have quit the keys. These same objections apply equally to a switch located in any other part of the key and hammer action. If however, the switch is operated by some part of the damper mechanism these difficulties are eliminated, for the reason that the dampers are controllable by both keyboard and pedals. When so operated, desired tones, only, may be held by use of the sostenuto pedal, by depressing it after the corresponding keys have been struck, and while they are still held; or any and all tones, once started, may be held by means of the damper, or so called, loud pedal. By such a plan, which we have incorporated in this instrument, the wind instrument or organ tones may be controlled with a normal piano-playing technique respecting both keys and pedals.

ORGAN-PLAYING TECHNIQUE WITH PEDAL

For varying the inception of single tones or chords played slowly, the foot volume control may be used advantageously for introducing, merely by swell pedal control, groups or single tones of wind instruments in an otherwise piano performance. It also aids the hands in building up crescendos or in diminuendo passages.

ENVELOPE-CONTROL MASTER SWITCH

By means of a main, multipolar switch, the individual string-charging switches may be so connected that either percussion or organ type tones are secured throughout the keyboard range; or one half may be made organ, the other percussion, or the reverse of this may be secured. To prevent loud noises in the reproducers, caused by the sudden and large voltage shifts in the pick-up system, resulting from the operation of this master switch, inter-position and overlapping segments on this switch provide, with a contactor, for short circuiting the 500 ohm line to the power amplifier as the switch changes position.

CONTROLS

In Figure 23, I show the various controls in a close-up, front view of the Electronic. At the extreme left is a master volume control for setting the extreme volume obtainable by the foot swell. Next comes the three amplitude and phase reversing controls for the

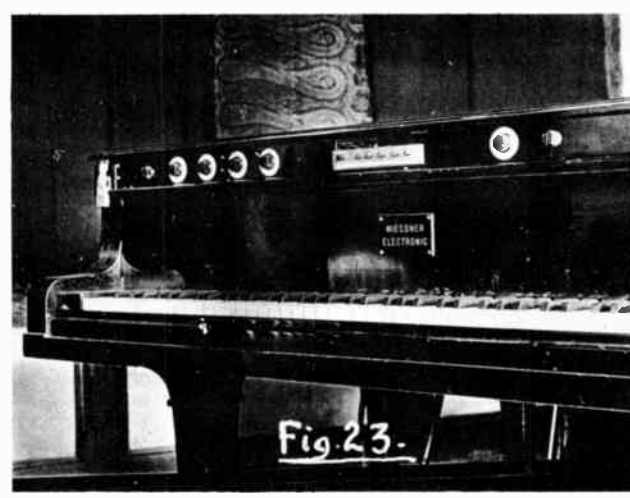


FIG. 23
CLOSE-UP OF ELECTRONIC PIANO CONTROLS.

manual mixing of the three pick-up tube amplifier outputs. Next is the amplifier-frequency characteristic control. In the middle is the preselected tone quality switch group of six buttons. The extreme left button is a red one and, when pushed in, serves to switch in the manual mixing controls to its left, or, when out, the other five buttons at its right, controlling preselected, tone mixtures. If one of these be pushed, with the red button already in operation, the red one springs out, disconnecting the manual controls, and the new one takes effect; or if one of the other preselected qualities be now desired, a push on its button will release the last and give the new quality. Next to the right is the master polarizing voltage control switch. This has four positions. One provides percussion tones throughout the piano keyboard; the next, wind instruments throughout; the next percussion below middle C and wind above; the last wind below middle C and percussion above. With any of the adjustments of this switch all of the tone quality controls are effective. The last knob, at the right, controls the artificial damping of percussion tones. Not shown in this picture are the swell, damper, and sostenuto pedals below on the lyre.

This completes, except for details, the description of what I have come to call my ELECTRONIC PIANO, wherein are brought together a very new and a very old art.

PATENT LITERATURE

For details of its construction, and for much more complete explanation of its operation; for other alternate methods and constructions, and improvements, and for descriptions of some wholly new electronic musical instruments, I refer you to our nineteen already issued patents, appearing as an appendix to my paper; numerous other patents will issue in the future.

COMMERCIAL DEVELOPMENTS

With regard to Commercial developments, these may be worked out in many ways.

No great skill is required in making an electronic piano. With the removal of the acoustic principle of its operation has gone practically all the artisanship of its construction. Furthermore, since the strings are no longer called upon to vibrate a large soundboard, they need not be nearly so long, or so large, or so high in tension. The iron plate on which they are strung may then be made smaller and lighter; the rim, and beams attached to it, no longer called upon to restrain the outward push of the soundboard, caused by the downward pressure of the strings, may now be much lighter; there need be only one or two strings where now there are three.

These changes in the piano itself make it much smaller, much lighter and much cheaper to manufacture. In fact the smallest baby grand provides more than sufficient size, and much more strength and weight than is required for our electronic design.

The electrical apparatus required to make it function is hardly more complicated than a good radio set. As a matter of good judgment, the audio frequency amplifier, reproducers, and power supply can serve, not only for the piano, but very logically, for a radio frequency amplifier and detector, and for a phonograph pick-up also. Such a complete instrument, providing radio, phonograph, and electronic piano performance, could surely be manufactured as cheaply as a good piano of moderate size. For very cheap instruments we may use the entire amplifying and reproducing equipment of an existing radio set, and provide in the piano, only a small tube oscillator, whose input to the radio set would be modulated by the vibration of the piano strings to and from the electrostatic pick-ups. Or, we may use only the audio amplifier and reproducers of the radio set, by providing a small preamplifier in the piano. Another branch of commercial activities might concern itself with conversion of existing acoustic pianos into Electronic pianos.

COMMERCIAL DESIGN

In Figure 24. I show a commercial design of an ELECTRONIC PIANO of this type, made under patent agreements with my own company, Miessner Inventions, Inc., and with Oskar Vierling, of Berlin, by the well known piano manufacturing company, August Förster, at Löbau Saxony, and in Czechoslovakia. These are now on the market in numerous foreign countries, and are known by the name "Electrochord."

SUMMARY OF IMPROVEMENTS

I have previously mentioned that all conceivable or useful controls over tone character such, for example, as feed back, to keep strings vibrating as long as keys are held, are not included in the present instrument. Among those included, in addition to those of a normal piano, are:-

1) A rather full control of harmonic composition, within the limits of the spectrum of the struck string.
2) Control of the tone envelope shape, except for indefinitely sustained tones.
3) Mass dynamic control by foot swell.

FIG. 24
COMMERCIAL DESIGN OF ELECTRONIC PIANO MADE IN GERMANY.

In addition, the divided keyboard permits, when desired, simultaneous play, in different registers, of tones differing in character by change of envelope shape.

CONCLUSION

In conclusion, and before beginning our demonstrations, I wish to express my thanks to the Radio Club of America, and to Columbia University and Doctor Dykema, for their kindness in arranging this meeting, and to my audience for its attentiveness to a long, and arduous technical paper.

I wish, further, to thank my brother, Dr. W. Otto Miessner, of Chicago, for his original suggestion some ten years ago of the piano as a research problem, and for his musical guidance from time to time; to Mr. Anton Rovinsky, for his many helpful suggestions; to Mr. Harold Bauer, for his interest and criticisms; to Dr. Erwin Meyer, of the Heinrich Hertz Institute, of Berlin, for use of his Spectral analyses of Musical instrument tones, shown in Figures 1, 2, 3 and 4; to Dr. Dayton C. Miller for oscillograms of violin and voice tones in Figures 4 and 5, from his "Science of Musical Sounds"; to my collaborator, Dr. Oskar Vierling, of the Heinrich Hertz Institute, of Berlin, with whom I have a cross-licensing patent agreement; and finally to Mr. Charles T. Jacobs, my technical collaborator and patent attorney, who has made numerous important technical contributions to this work during the past three years.

APPENDIX

LIST OF ISSUED PATENTS IN U. S. A.

SUBJECT	Number	Issue Date
Photo Electric Organ	1,886,687	Nov. 8, 1933
Multiple Pick-up Tone Control	1,906,607	May 2, 1933
Bucking String Organ	1,912,293	May 30, 1933
Sidewise Translation	1,915,858	June 27, 1933
Time Delay Organ	1,915,859	June 27, 1933
Translation-efficiency Variation	1,915,860	June 27, 1933
Multiple Translation	1,915,861	June 27, 1933
Tone Control By Sidewise Pick-up	1,929,027	Oct. 3, 1933
Amplifier Frequency Control	1,929,028	Oct. 3, 1933
Inter-vibrator Coupling	1,929,029	Oct. 3, 1933
Uncouplable Soundboard	1,929,030	Oct. 3, 1933
Sustained Tones by Feedback	1,929,031	Oct. 3, 1933
Tone Control Stop System	1,929,032	Oct. 3, 1933
Different Vibrators for Tone Types	1,933,294	Oct. 31, 1933
Resonant String Organ	1,933,295	Oct. 31, 1933
Non-linear Translation	1,933,296	Oct. 31, 1933
Longitudinally-vibrated Bars	1,933,297	Oct. 31, 1933
Flattened String Organ	1,933,298	Oct. 31, 1933
Timbre Control, Formants, etc.	1,933,299	Oct. 31, 1933

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THE BEHAVIOR OF GASEOUS DISCHARGE TELEVISION LAMPS AT HIGH FREQUENCIES

By
H. J. Brown*

Probably the most common source of light used for the reception of television images is a gaseous discharge known to the physicist as a glow. The discharge is usually thru one of the noble gases as argon or neon in a sealed glass envelope at a low pressure. The light output of a lamp so constituted is capable of rapid changes and while it is far from ideal in color the simplicity and ease of construction of such lamps and the convenience of their application has given them wide use in mechanical scanning systems.

In discussing the elements of a mechanically scanned television system it is relavent at the start to mention one important feature of the system as a whole.

While it is probably true that a mechanically scanned television system can only with great cost and complication realize the same large number of picture elements that are readily possible with the cathode ray type of scanning system, the ease with which precise spot placement and spot size may be realized in the mechanical type of system off-sets this advantage of the alternate type of system to a degree that has been estimated to be as high as fifty per cent.

In view of these facts and in view also of the fact that the mechanical difficulties in the mechanical systems are probably not without their counterparts in the electrical problems involved in the cathode ray type of system, it is felt that there is still a considerable justification for the investigation of the properties of the glow lamp with their application to the problems of television in mind as is herein reported.

The glow lamps discussed in this paper were of two distinct types. One, suitable for use with aperture disc, included flat plate cathodes 1½" square while the other was of the "crater" type with a circular cathode 0.060" in diameter. The gaseous atmosphere of the lamps was neon in both cases, at pressures giving a breakdown voltage of about 175 volts. The current rating of both lamps was about 30 milliamperes so that the current density at the cathode of the crater lamp was many times that of the other lamp. The emitted light in both cases came from the negative glow region of the discharge and was identical in appearance.

In a glow discharge as employed in the lamps here discussed the negative glow is a region characterized by high ion density, small net space charge and low electric field intensity. The dark region between the glow and the cathode has a high space charge, and a much greater field intensity. In fact substantially the entire voltage impressed upon the lamp appears across this cathode dark space, the electrons gaining velocity from this field to plunge them into the negative region with sufficient energy to create positive ions and to excite atoms with a consequent production of light. In Fig. 1 is shown the static characteristics of the two types of glow

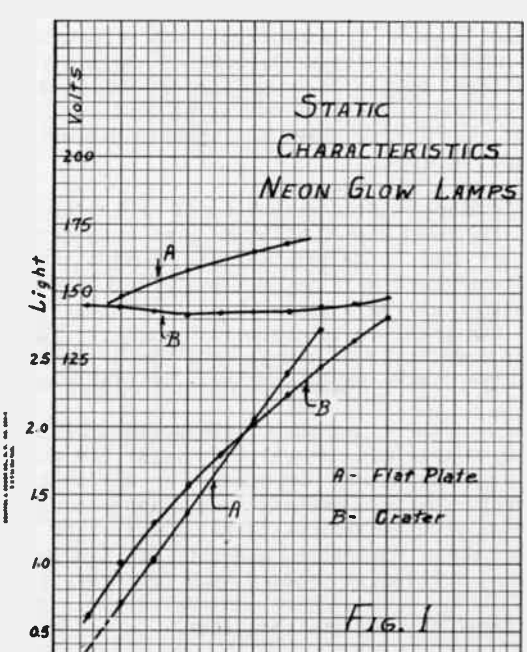

FIG. I

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lamps under test where it may be seen that the production of light is a simple, direct function of the current thru the lamp and is largely independent of the voltage across the lamp, from which it is obvious that the practical and important characteristic of these lamps is their current-light relation particularly in so far as this characteristic is related to the frequency of the current exciting the lamp.

When operating in a conventional television system, the current thru the lamp is a modulated direct current. This is equivalent, of course, to an unvarying current with an alternating current superimposed. The alternating component of the current should provide an alternating component of the emitted light if the lamp is to be suitable for television purposes, and a lamp of such characteristic may then be considered as a transducer between current and light and as such requires that consideration be given to the following important characteristics.

Frequency Discrimination.

If the degree to which the light is modulated for any degree of current modulation is not substantially the same at all useful frequencies of current modulation the light output of the lamp will not truly represent the wave form of the current and will, of course, emphasize certain frequencies or ranges of frequency and discriminate against other ranges of frequencies.

Phase Shift.

If the light output of the lamp is not in precise time agreement with the current thru the lamp thruout the cyclic variation of both, there is, a phase difference between the light and the current. Such a phase difference between current and light output is obviously present in the filament type of lamp as a result of its thermal inertia and in which the light lags behind the current in time relationship. In the gaseous conduction lamp, however, the light will often lead the current as is shown in what follows.

Harmonic Distortion.

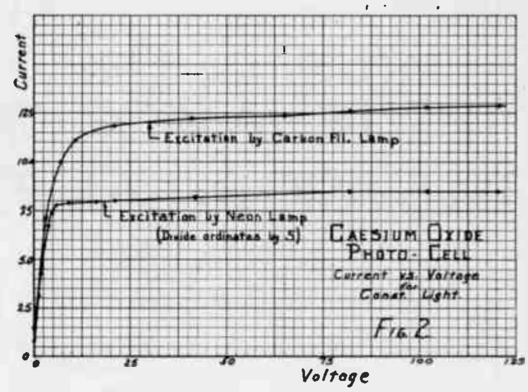
The dynamic light-current characteristic may depart from strict linearity in which event harmonics in the light output will be present over and above such harmonic content as may be present in the lamp current thus giving rise to harmonic distortion of the light output.

If these characteristics of a current-light transducer such as the lamps here considered are known and found to be satisfactory, only a knowledge of the characteristic impedance of the lamp is required to serve as a basis for the rational choice of the equipment to be associated with the lamp and for the prediction of the general operating properties of the system employing these several elements.

And if there is a reasonable degree of linearity between voltage and current, and light and current, a vector diagram may be drawn for any particular frequency showing the relative magnitudes and phases of these three variables. If any correlation exists between these variables a series of diagrams for different frequencies should tend to bring it out.

Measurement of Lamp as Transducer:

The obvious way to measure the light output of a lamp whose color is nearly invariable is to use a photo-cell. The electron emission from a photo-cell's cathode is strictly proportional to the quantity of light of fixed frequency impinging on it. Thus if saturation voltage is applied to the cell anode a current will flow, the magnitude of which is proportional to the light falling on the cell. In high vacuum cells there will be few collisions between electrons and gas molecules so that we need expect no internal effects, other than those due to the geometrical capacity between elements, tending to prevent the space current of the cell from following the comparatively slow flucuations of light encountered in these experiments. The external characteristic of the cell (Fig. 2) for constant impressed



light shows a definite Schottky effect but it hardly suggests the presence of ionization by collision, the curve being concave downward up to 135 volts. In our experiments we used a cell voltage of 40, large enough to insure saturation, and small enough to prevent appreciable ionization current.

It is especially interesting to note the difference between excitation by a line spectrum (neon lamp) and a continuous spectrum (filament lamp) in the sharpness of cutoff, as shown in Fig. 2.

The A.C. component of the photocell current must of course be built up to a value sufficient for measurement of magnitude, of phase, and of wave form. There must be no question of fact that this transfer takes place with a known degree of accuracy, and for this reason, the comparison method of measurement was used as indicated in Fig. 3 and 4.

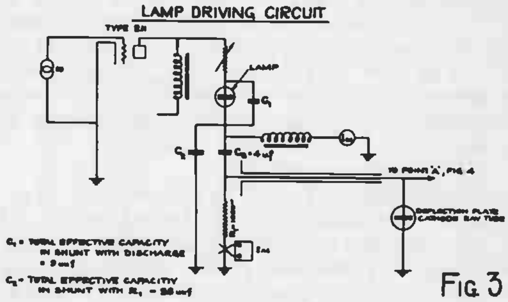


Fig. 3

The lamp driving circuit (Fig. 3) is arranged to superimpose an alternating current relatively free from harmonics on the steady direct current thru the lamp. This alternating current, flowing thru R_1 actuates one pair of deflection plates of a cathode ray tube and provides a calibrating voltage for the amplifier shown in Fig. 4, the magnitude of this current being measured by a thermocouple. The attenuator (Fig. 4) is so adjusted that the voltage introduced at the input of the amplifier is approximately the same as the A.C. voltage generated across R by the photocell. The output voltage from the amplifier actuates the voltmeter V_o and the second pair of deflection plates of the cathode ray tube. With the photo-cell hooded, the voltage V_{ac} (proportional to the A.C. component of the lamp current) is introduced into the amplifier which is then adjusted for unit output and zero phase shift. Point A is then disconnected and grounded and the cell is unhooded. The resultant current-light excursion on

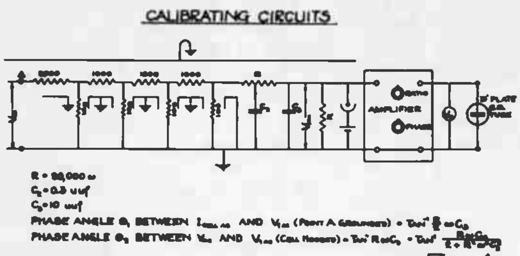


Fig. 4

the cathode ray tube will show the phase shift and any harmonics present. The voltmeter reading will tell the relative response.

In making these measurements a multistage wide range amplifier was used. Altho the measured response of the fundamental frequency in the light output is independent of the amplifier characteristics, to obtain a true current-light excursion the amplifier and input circuit must be distortionless thru the range of harmonics encountered in the light output. It is a simple matter to extend the distortionless range of the amplifier by increasing the number of stages while cutting down the gain per stage. But to increase the range of the input circuit it is necessary to reduce the value of resistor R. Random noises in the input circuit will tend to encroach on the signal as R is reduced making the measurements less satisfactory and for this reason a minimum value of 20,000 ohms was chosen for R. This gives a conservative upper limit of 160 k.c. for the calibration for the fundamental frequency and a lower value, depending on the harmonic content in the light output, for an accurate excursion. As it happens, however, the harmonics in the lower region are not noticeable and in the upper region are not qualitatively indicated on the excursion, so their effect will be neglected in the measurements.

The resistors used in the calibrating circuits and in the measurement of the lamp current are of the metallic film type. These elements have only easily calculable and measurable geometrical capacitances to affect their impedance. It is, therefore, a simple matter to choose values that will be substantially non-reactive. Only in the case of R (Fig. 4) was it necessary to supply especial mountings to reduce and establish this capacitance as represented by C_2. In the attenuator the effective capacitance in shunt with the resistors is largely negligible because of their low values and is eliminated from R by shielding.

Experimental Results;
Light vs. Current, Amplitude Response and Phase Shift:

The frequency response curves as a current-light transducer are shown in Fig. 5 for the two lamps whose static characteristics were given in Fig. 1. At low frequencies the excursion for both lamps has the same general form as the static characteristic and a very uniform response. In the case of the crater lamp, the response after falling to a minimum around 10,000 cycles slowly rises with frequency until at the upper limit of our measurements it is about 50 per cent greater than the low frequency value. The flat plate lamp has a 20 per cent increase in response to 40 k.c. where there is a sharp discontinuity. It then falls to a value at 160 k.c. not markedly different from its low frequency response.

While the above results are strictly true only for a degree of modulation of about 50 per cent, they are characteristic of the behavior of these lamps over wide ranges of modulation and polarizing current, with the exception of the break in the curve

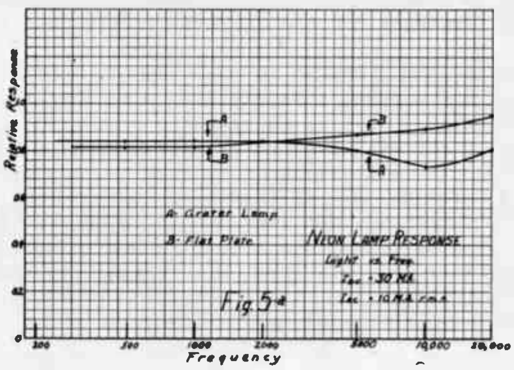


Fig. 5-a

of the flat plate lamp. This break is accompanied by so high a circuit impedance that it is impossible to force alternating current thru the lamp, and it is a property of this particular lamp for a narrow band of frequencies between 39 and 41 k.c.

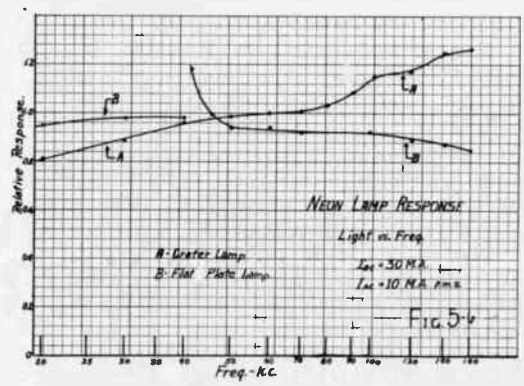


Fig. 5-b

There is a marked difference in the performance of the two types of lamps with respect to phase shift. All crater lamps tested having a small enclosed cathode showed a leading angle for the light with respect to the current while there was no observable phase shift with flat plate lamps. The angle of lead for the crater lamp whose response is plotted above is about 15 degrees for the above conditions at 50 k.c. It is proportional to frequency from zero to 90 k.c. after which tends to be constant. It is independent of the degree of modulation and decreases with increasing polarizing current.

Lamp Impedance:

The measured values of lamp impedance affords an interesting correlation with their observed light response. As with all instances of gaseous conduction, there is a time lag between ionization and the ionizing agent. In general it would be expected that devices such as these lamps, which depend on the creation of ions by collision to provide for the passage of current, would present inductive reactance to alternating currents. In Fig. 6 and 7 is shown the equivalent self inductance for the crater lamp as a function of frequency and current. This self inductance provides substantially the entire impedance of the lamp, the resistance component being small or negative. For high frequencies the inductance tends to approach a constant value determined only by the polarizing current. Thus, the alternating current lags the voltage by a large angle.

The flat plate lamp also shows self inductance. It tends to have a constant reactance

(Fig. 8) rather than a constant inductance; but as the resistance component is large and increasing

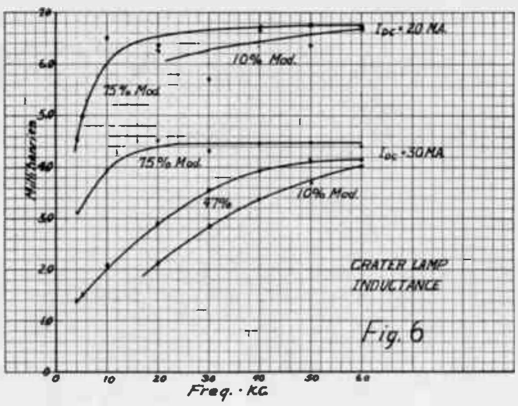


Fig. 6

with frequency, this lamp has a very much smaller phase angle between the applied current and voltage.

In Fig. 9 are given vector diagrams showing the relationship between the factors of impedance and light characteristics. It is definitely suggested

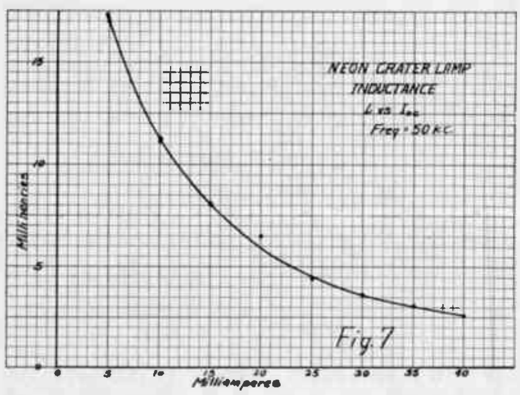


Fig. 7

by these diagrams and the response curves that the increasing voltage variation across the lamps is responsible for the rise in the light response with frequency. It will also be noticed that a small phase angle between current and voltage is associated with a negligible phase angle between current and light, while the large voltage-current angle of the crater lamp has with it a light, current phase shift.

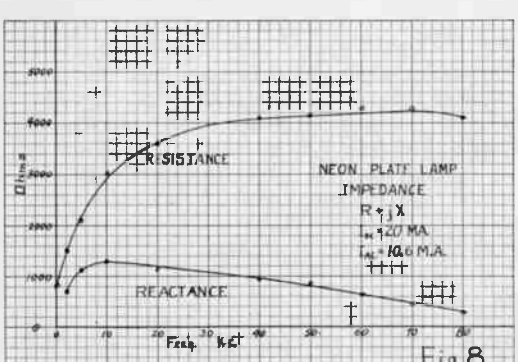


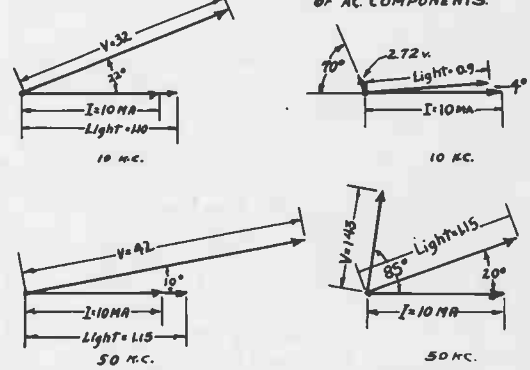
Fig. 8

The impedance variation also parallels the variation of the light response for the two lamps. The continuously rising impedance of the crater lamp has with it a continuous rise in the light response. The impedance of the flat plate lamp tends to approach a maximum in the same range as does the light response.

Circuit Application:

The circuit application of neon glow lamps is very simple. They should be driven thru a sufficiently high impedance circuit to minimize the effects of non-linearity, frequency discrimination, and phase shift inherent in the impedance characteristics. The pentode type of output tube with a plate resistance of the order of 100,000 ohms has shown itself to be satisfactory. A simple series connection of

NEON LAMP - VECTOR DIAGRAMS
OF AC COMPONENTS.



FLAT PLATE CRATER
FIG. 9

the lamp, tube plate and power supply is preferable to a shunt feed in that it provides a lower capacity shunting the lamp. This is very desirable as the crater lamp at high frequencies has a self inductance inversely proportional to the current. For a sinusoidal current thru the lamp there will be by necessity harmonics in the voltage. Shunt capacity around the lamp will decrease the voltage harmonics with a corresponding increase in the current harmonics flowing thru the lamp, as well lower the actual response. The upper limit of response of the crater lamp will obviously be limited by its inductance and its shunting capacity. With a series feed, this capacity need not be greater than 25 uuf. including the plate to ground capacity of the tube. As the lamp inductance is around 4 millihenrys, it may be safely said that, to the upper limit of the current-light response measurements, this lamp will cause little loss in picture quality.

Conclusions:

It is believed that the data here given shows quite positively that a crater lamp of conventional design will be found completely satisfactory for television systems employing 120 lines per frame and 25 pictures per second. And it is quite likely that they will be found useful for systems requiring even greater detail than this. Since no theoretical basis has been evolved to rationalize the results here reported, it is not possible to make any estimate of the limit of response of the behavior of lamps of other physical characteristics. However, in view of the fact that the interesting characteristics herein reported were found in lamps which had been built without any attempt at design for any special characteristics and without any understanding of the factors involved in the high frequency characteristics, it is believed that glow lamps can be developed that will provide satisfactory performance in any otherwise practical television system.

In closing I wish to acknowledge with sincere thanks the advice and assistance of Mr. J.B. Russell of Columbia University.

THE DESIGN OF RESISTANCE ATTENUATORS FOR RADIO FREQUENCY USE

By
Malcolm Ferris*

Resistance attenuators have recently been very widely used for radio frequencies up to about 20,000 kilocycles, especially in signal generators and similar apparatus.

The resistances used as elements in such attenuators must be so designed that skin effect, series inductance, and shunt capacity do not introduce appreciable error. I have found that wire wound resistors of the bifilar type are most satisfactory, as this type of winding has the lowest inductance. It is true that it also has high shunt capacity, but this can be held to values that are negligable even at 20,000 kilocycles, for resistances up to 25 ohms. For higher values of resistance, two or more 25 ohm resistors can be used in series. An attempt to wind a bifilar resistance of several hundred ohms in one unit would of course result in failure, due to high shunt capacity. This does not show, however, that the bifilar type of winding is unsuitable for use at high frequencies, but merely that suitable care must be exercised in using it.

Resistances of 5 to 100 ohms are the most practical to construct. Below 5 ohms, very careful design is necessary if the inductance of the leads is not to cause error, and above 100 ohms, shunt capacity becomes more troublesome.

Far more troublesome than the effect of inductance or capacity in the individual resistances, is the effect of mutual inductance, or capacity between sections of the attenuator which are operating at widely different voltage levels. Before considering these effects in detail, one factor that is of greatest importance should be especially noted.

Resistance and inductive reactance in series do not add algebraically, but combine vectorially in quadrature. For instance, a resistance of ten ohms and an inductive reactance of one ohm connected in series have a total impedance of 10.05 ohms, and NOT 11 ohms. This means that an inductive reactance of one ohm in series with a ten ohm resistor, will cause only one half of one per cent error, not ten per cent error, as might be expected. This is one of the important advantages of a resistance attenuator, and undoubtedly is one of the most important reasons for its extensive use.

Every engineer who has considered resistance attenuators has of course realized that undesired inductance and capacity may cause serious errors. A suggestion that has often been made is that two condensers in series would form an ideal voltage divider, free from frequency error. This would be true if it were possible to have such an arrangement entirely free from inductance. Unfortunately, this is impossible, and, as mentioned above, the undesired inductance causes a much more serious error than it would in a resistance attenuator.

The arrangement of apparatus shown schematically in the diagram of figure 1 is for the purpose of illustrating this effect. If the slider is moved

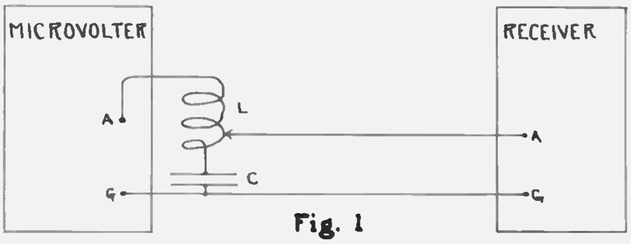


Fig. 1

along the inductance L, a nodal point will be found where the input to the receiver falls almost to zero. The photograph of figure 2 shows the actual arrangement of this apparatus, and the point of special

*Ferris Instrument Co., Boonton, N. J.

interest is the actual physical appearance of the inductance. It will be noted that this inductance is merely a short length of wire, two or three inches long. At 12,500 kilocycles, the nodal point occurs

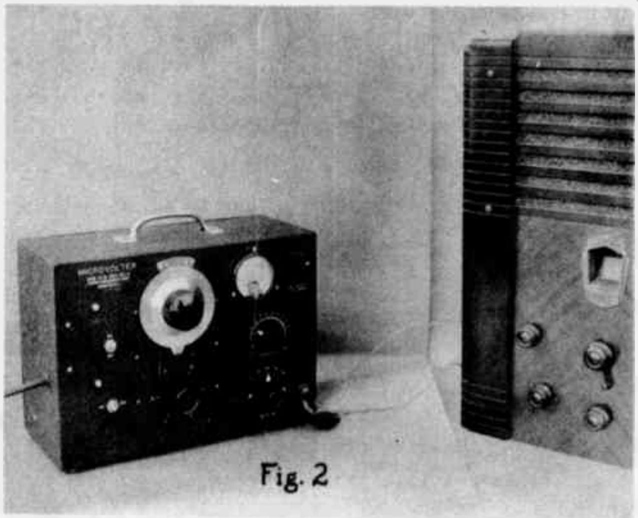


Fig. 2

with the slider about one inch from the condenser terminal. At 15,000 kilocycles, the nodal point is practically at the condenser terminal. In other words, at 15,000 kilocycles, the inductance present in the condenser itself, plus the mutual inductance in the leads, is sufficient to cause one hundred per cent error. It can be seen that it would be almost impossible to make any arrangement of this apparatus which would reduce the inductance enough to make the scheme useful at frequencies of the order of 10,000 to 15,000 kilocycles, and even at frequencies of the order of 1,000 kilocycles great care is required to avoid such errors. (The condenser used in the above test was a .002 molded type)

Remembering that stray voltages due to undesired inductance and capacity effects combine vectorially in quadrature, we can consider some specific cases, and see to what limits these undesired inductances and capacities must be held. Figure 3 is drawn to illustrate two of the most important sources of trouble which must be considered. The attenuator

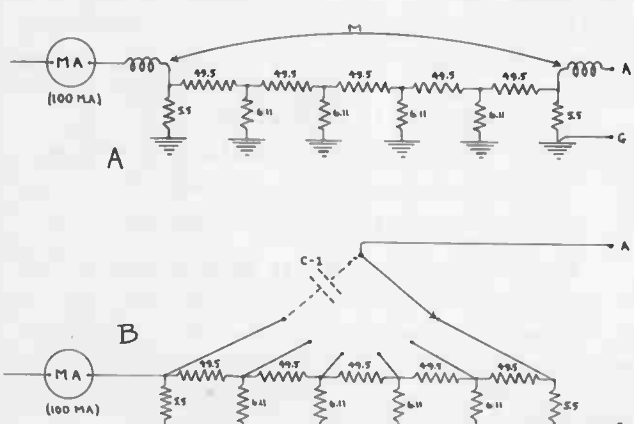


Fig. 3

shown has a five ohm impedance on each point, and, with 100 M.A. input, has 500,000 microvolts at its first point, and 5 microvolts at its output end.

Consider the mutual inductance indicated in figure 3(a). If we limit the voltage introduced into the output circuit by this mutual inductance to one half microvolt, which, combining in quadrature with the five microvolt output of the attenuator, would produce only one half of one per cent error, then the mutual reactance must not exceed 1/200,000 ohm. Since at 10,000 kilocycles one microhenry has a reactance of 62.5 ohms, this mutual inductance must not exceed one thirteen millionth part of a microhenry, if the above conditions are to be met at 10,000 kilocycles.

In practice, the only way in which such a low value of mutual inductance can be obtained is by means of very effective shielding between the attenuator input and output circuits.

Figure 3(b) is arranged to show another type of effect which will cause serious error unless properly considered in the design. The attenuator in this case is the same as in figure 3(a), but a switch is shown by which the output lead can be connected to any point on the attenuator. A condenser C-1 is shown to represent the capacity existing between the switch point which connects to the high end of the attenuator and the switch arm, which in the figure is shown connected to the output end of the attenuator. For the same conditions as above, the current flowing thru C-1 must not exceed one tenth microampere, and since the driving voltage is 500,000 microvolts, this means that its reactance must be at least 5,000,000 ohms, which at 10,000 Kcs. means a condenser not greater than 3/1000 of a micro-microfarad.

The permissible value of mutual inductance and capacity as computed above are of interest mainly to show the order of magnitude within which these factors must be held. It is of course impossible to measure one thirteen millionth of a microhenry by any of the usual methods of measuring inductance. Tests of the attenuator for "leakage" and accuracy usually show whether these values are within the desired limits.

Figure 4 shows two attenuators which differ only in the switching arrangement, the input being switched in figure 4(a), and the output in figure 4(b). As shown, there is little to chose between these two methods of switching.

The attenuators of figure 4 have a constant impedance of five ohms at any point, but give only five fixed values of output, namely 500,000, 50,000, 5,000, 500 and 50 microvolts. For practical use, this is not sufficient, and some scheme must be used to provide either continuous variation of output, or else many more steps. Two schemes for this purpose are in common use, and both methods have their own disadvantages as well as advantages. These two schemes are:

(a) Variation of the input current to the attenuator.
(b) Substitution of a calibrated potentiometer for one of the resistances of the attenuator.

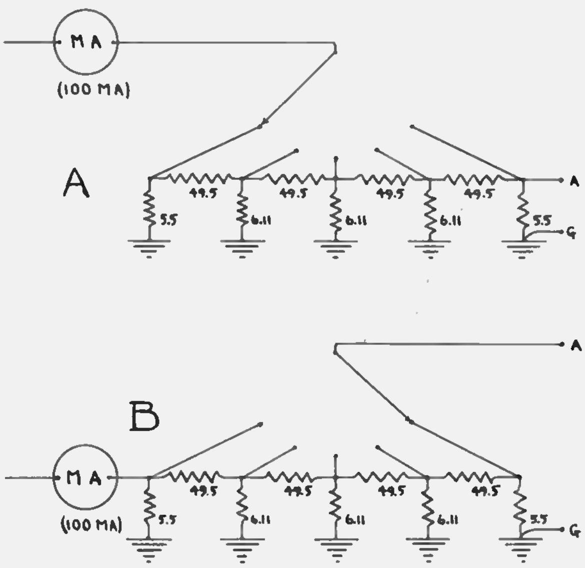


Fig. 4

Variation of the input current to the attenuator is almost certain to react on the frequency of the oscillator, unless a master oscillator-power amplifier type of circuit is used. This reaction is most serious when a heavy load must be drawn from the oscillator. Careful design of the oscillator circuit will reduce this reaction, but never entirely eliminate it.

Unfortunately, thermo ammeters and other types of meters suitable for measuring attenuator input have a deflection which varies with the square of the current. This means that about one third of full scale current is about the lowest value that can be read practically, and even at this value the accuracy of reading is very poor. It is therefore necessary to design an attenuator with voltage steps, having roughly a three to one ratio, instead of a ten to one ratio, thus making necessary many more steps in the attenuator, and also making necessary a multiplication of the meter reading by a value other than a multiple of ten, so that it can no longer be considered direct reading. It has been common practice to employ multiplying factors of 1, 2, 5, 10, 20, 50 etc., or 1, 3, 10, 30, 100 etc.

The use of a calibrated potentiometer overcomes the above disadvantages, since the input current need not be varied, and the potentiometer scale can be calibrated directly in microvolts, with multiplying factors of 1, 10, 100 etc. On the other hand, it introduces disadvantages of its own. It may be subject to wear, and change in calibration, so that it must be renewed after a period of service. Also, while suitable potentiometers of fairly high resistance are easily made, a low resistance potentiometer which is sufficiently free from inductance is quite expensive to make.

Practical attenuators involve various combinations of the above features, and may perhaps best be discussed by examining some specific designs, and discussing the features involved.

The diagram of figure 5 shows an attenuator quite similar to that shown in figure 3(b). It differs from that of figure 3(b) in the following particulars:

(1) It has one less section of attenuation
(2) It has a calibrated potentiometer for the final resistance
(3) The input is switched, instead of the output.

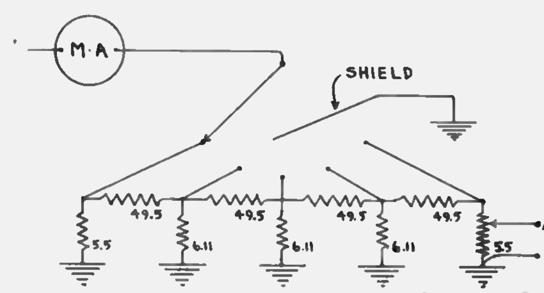


Fig. 5

Using a calibrated potentiometer for the output resistance of course makes it impossible to switch the output, and it has been necessary therefore to switch the input. As the attenuator has one less section than the one shown in figure 3, the effects of undesired inductance and capacity or only one tenth as serious. In the actual construction, it was found possible to use a conventional type switch, with only the precaution of a shield run thru a slot in the bakelite panel to lower capacity between the switch arm and the switch points.

The output potentiometer of this attenuator carries a dial which is calibrated directly in microvolts, from zero to fifty. The attenuator switch provides multiplying factors of 1, 10, 100, 1000 and 10,000. The output is therefore continuously variable from zero to 500,000 microvolts (one half volt), and the smallest scale division is one microvolt.

It will be noted that in this attenuator the output resistance is never over five ohms, but is always variable from zero to five ohms. There are some conditions, such as tests on highly regenerative receivers, where such a variable output resistance is undesirable.

Figure 6 shows a very different type of attenuator, and the photograph of figure 7 shows the actual appearance of three different attenuators. In this photograph, (a) is the attenuator of figure 6(b), (b) is the attenuator of figure 5 above, and (c) is the attenuator shown in figure 9 below, in one of its early experimental forms.

The arrangements shown in figure 6(a) and (b) differ only in the switching scheme. It will be

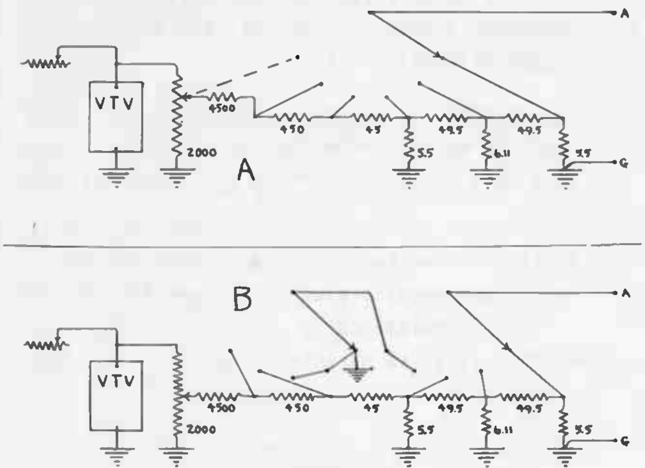


Fig. 6

noted that both switch arrangements give the same schematic connections, but the somewhat more complicated arrangement shown at (b) has to be used in practice to avoid trouble due to capacity across the switch, as illustrated in figure 3(b). This trouble is avoided by dividing the switch into two sections, so that the extreme difference in voltage levels does not occur in one section. Another method of

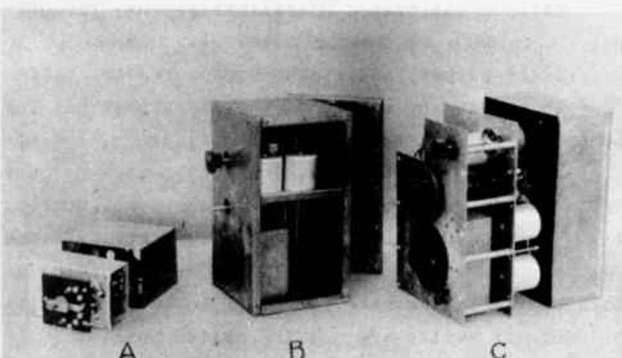


Fig. 7

eliminating this trouble would be very careful shielding of the switch, as has been done in the case of the attenuator shown in figure 7(c).

The attenuator of figure 6 has for its first resistance a 2,000 ohm calibrated potentiometer. A radio frequency voltage of one volt is maintained across the terminals of this potentiometer. The rheostat shown at the left is adjusted until the vacuum tube voltmeter shows just one volt across the potentiometer. The potentiometer dial carries a scale marked from zero to ten microvolts, with a division every one fifth microvolt, and the attenuator switch provides multiplying factors of 1, 10, 100, 1,000 and 10,000.

It should be especially noted that the input current to the attenuator shown in figure 6 is only ½ M.A., as compared with 100 M.A. in the attenuator of figure 3. Therefore, for the same permissable stray field in the output circuit, the undesired mutual inductance as shown in figure 3(a) can be 200 times as great. Stated in other words, the shielding problem is much less severe. What this means in actual structure can be readily seen by comparing the attenuator at (a) in figure 7, which has ½ M.A. input, with those at (b) and (c), which have 100 M.A. input.

In figure 6(a) a sixth point on the switch has been shown dotted, connected direct to the potentiometer slider. If this point were used, it would provide output up to 1 volt, instead of 1/10 volt. However, it would require much more shielding of the switch, and in addition, the extra capacity to ground from this point which the unavoidable leads would add would cause attenuator inaccuracies at the higher frequencies.

One very definite sacrifice has been made in the attenuator of figure 6 to make a simple construction possible. This is the use of high output resistance for the higher output voltages. It will be noted that the output resistance is five ohms on the first three points (Multiply by 1, 10 and 100 - that is, output voltages up to 1,000 microvolts) On the fourth point (output up to 10,000 microvolts) the resistance is 50 ohms, while on the final point it is 500 ohms.

In general, resistances of 50 and 500 ohms cannot be introduced into a receiver antenna circuit without causing appreciable error. The amount of this error will of course depend on the particular receiver. In one receiver recently tested, which was a broadcast receiver of conventional type, at 600 Kcs., fifty ohms in the antenna made a negligable difference, and even 500 ohms made a very slight difference. On the same receiver, at 1400 Kcs., 50 ohms made a slight difference, but 500 ohms caused an error of about 20 per cent.

The attenuator of figure 6 provides 5 ohm output resistance up to 1,000 microvolts, and all sensitivity readings will normally fall below this value. If an attenuator of this type is used for taking selectivity readings, higher values of output will be required, and a check must be made to see what effect the added resistance will have on the results, so that a correction can be made if necessary.

Of course, the ideal attenuator would have low output resistance, even for the highest output voltages. As mentioned above, this means a more elaborate shielding structure, and therefore a more expensive arrangement. Also, it requires more power from the oscillator, which also complicates the design problem. For these reasons, most commercial attenuators use higher output resistance for high output voltages, and the user must bear in mind the limitations thus imposed.

Figure 8 shows an attenuator designed for obtaining continuous variation of output by varying the input current, instead of using a calibrated potentiometer. It will be noted that a great many more

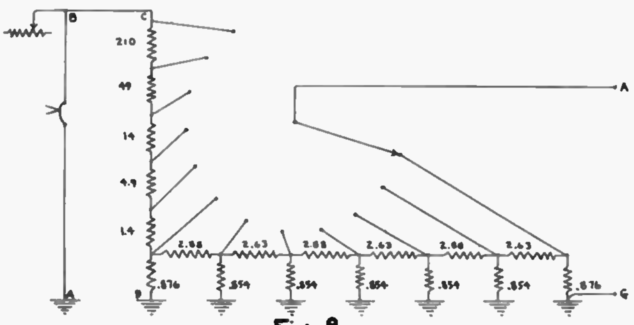


Fig. 8

steps are necessary in the attenuator. The thermo element used with this attenuator is a high resistance one, and is connected to measure the voltage input to the attenuator, instead of the current. For the design as shown, the maximum voltage at the input terminals is two volts, and the meter scale is calibrated from zero to five microvolts. Multiplier settings are 1, 3, 10, 30, 100 etc., up to 100,000, and then the final one is 400,000.

This arrangement is intermediate between those shown previously in the matter of output resistance. It does have high resistance for the highest output, but has only 7 ohms at 50,000 microvolts, and 21 ohms at 150,000 microvolts. It uses very low values of resistance, which are very difficult to make free from trouble due to series inductance. The diagram does not show the shielding necessary around the switch, which is of course necessary to prevent serious errors.

It might be mentioned that the attenuator of figures 6 and 8 show meters for measuring voltage input to the attenuator, while the one in figure 5 measures current input. Current measurement is usually more suitable for attenuators of low input resistance, while voltage measurement is often better when the input resistance is high. Both methods have possibilities of error, which must be reduced by careful design in particular cases. Referring to figure 8, any voltage induced in the loop ABCD will cause error, as it will introduce into the circuit voltage additional to that measured by the element. With a high resistance, low current input, it is usually easy to arrange the parts so that no appreciable magnetic field is picked up by the loop ABCD. With the method of measuring current input, as shown in figure 5, capacity to ground might cause error, as this would by-pass part of the current, and the entire current as measured would not pass thru the attenuator. With a high resistance attenuator, this error could easily be serious, but with an attenuator of five ohms input, as shown in figure 5, it would require a capacity over 150 micro-microfarads to cause error even at 20,000 kcs., and the capacity can very easily be kept below this value. (With a 200 ohm input attenuator, this permissable capacity would be reduced to less than 4 micro-microfarads, and this would be difficult to obtain.)

Figure 9 shows another arrangement, which combines some of the features of both methods of continuous variation. The right hand part of it is

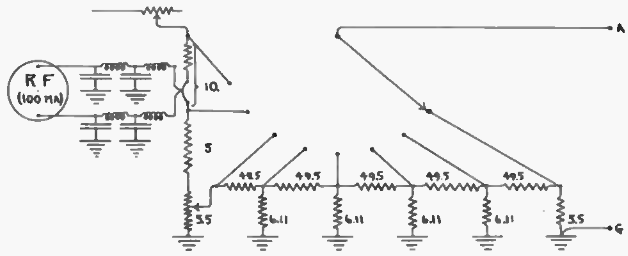


Fig. 9

similar to figure 5, except that it contains an additional step of attenuation, and the calibrated potentiometer is at the beginning instead of the end of the attenuator, thus making it necessary to switch the output instead of the input. This arrangement provides a fixed five ohm output resistance up to 50,000 microvolts, and a variable, zero to five ohms, from 50,000 to 500,000 microvolts. Above this, there are fixed taps at one volt and two volts, across ten and twenty ohms respectively, and intermediate values can be obtained by varying the current input. This arrangement provides output up to two volts across resistance low enough to be inserted directly in the usual antenna circuit, and does not require variation of the current input, except for values above one half volt. It does, of course, require extensive shielding, and draw heavy power from the oscillator. It might be mentioned that one advantage that goes with this is that a much more satisfactory type of meter can be used than where less power is available for the meter.

Figure 7(c) shows the appearance of an experimental attenuator of the kind shown in figure 9. It may be noticed that only a single switch is used, but that very elaborate shielding of the switch is necessary in this case.

Examples of attenuator design could be continued indefinitely, but it is believed that those described above are sufficient to illustrate many of the factors involved in practical designs.

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THE PHOTRONIC CELL & PHOTRONIC CONTROL

R. T. PIERCE*

The PHOTRONIC Photo-electric Cell represents the latest development in the photo-electric art. The position that it has in the art may best be shown by a brief review of the past development in photo-electric cells.

Photo-electric active materials have been known to scientists for over sixty years and a great deal of work was done originally on the selenium type of cell. This consisted of a plate of selenium which has the characteristic of changing its resistance under the action of light. A battery and meter, or relay, connected in series with this selenium cell allows the circuit to be adjusted so that the change in resistance under the action of light will change the current in this circuit enough to cause the instrument to deflect, or the relay to make contact. The selenium cell circuit, of course, is absolutely dependent upon the battery voltage and the current never goes to zero in this circuit, regardless of the condition of illumination on the cell. The selenium cell itself is quite sluggish in its operation and changes from time to time due to oxidation and other action which affects it chemically.

In 1885 a patent was issued which showed a method of holding a ship on its course by means of a light source and selenium cell used in conjunction with the compass needle. At that time no electric lights were available and this patent shows a kerosene lamp and a candle for the light source. In 1886 another patent was issued for controlling the level of water in a boiler using a selenium cell with a kerosene lamp and candle for the light source. It may, therefore, be seen that some of the early engineers and thinkers on photo-electric cells and photo-electric control were considerably ahead of their time.

Another type of cell that has been produced is the liquid type photo-voltaic cell. This is much the same as a wet battery having one electrode which is affected by light in such a manner that voltage is generated whenever this light hits the electrode. This is like any other battery in that the output changes as the electrolite ages and, under the action of generating voltage, gas is liberated which has been known to explode the cells with considerable violence.

With the advent of the vacuum tube various other types of photo-electric cells were produced in evacuated glass tubes requiring a voltage to produce a flow of electrons between the light-sensitive surface inside the tube and the collecting ring. Electrons were emitted from the light-sensitive surface which passed across the intervening space to the collector ring and this small current was passed on to the grid of a vacuum tube and the output amplified to obtain enough useful current for indication or control purposes.

There has been quite a bit of resistance toward the adoption of this type of photo-electric cell, by industrial establishments due to the fact that their experience with radio sets makes them somewhat hesitant about putting vacuum tubes on important control circuits.

THE PHOTRONIC CELL

About five years ago, the Research Laboratories of the Weston Electrical Instrument Corporation started investigating the possibility of providing a photo-electric cell or "electric eye" that would not require any auxiliary source of energy, and that would have sufficient power output to obviate the necessity of using amplifying circuits. That was the type of specification with which they started, and it must be admitted that at the start it seemed rather a hopeless task. However, their efforts were finally successful, and the result was the PHOTRONIC Photo-electric Cell.

The PHOTRONIC Cell consists essentially of a metal disc on the surface of which light-sensitive material is deposited. Contact fingers lead the current in and out of the cell by resting on the light-sensitive surface and also on the back of the plate on which it is deposited. Under the stimulation of light energy, electrons are emitted from the light-sensitive surface producing a current that flows into the plate beneath it, through the external circuit, and back to the cell again. It may be likened to a catalytic agent that transforms light energy into electrical energy without any deterioration of the cell.

*Weston Electrical Instrument Company

The cell output current starts from zero when the cell is in darkness, and increases, practically proportionally to the illumination, up to a value beyond the intensity of bright sunlight, when a low external circuit-resistance is used. This characteristic of the cell is very important both from the standpoint of light intensity measuring, and also from the standpoint of obtaining a control between quite narrow limits of illumination. The illumination-current characteristics of the cell are shown in Figure 1.

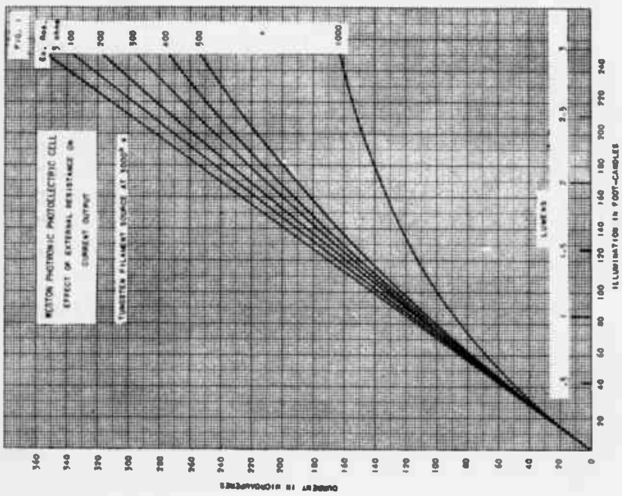

Fig. 1

Another important item in the specifications to be met by the cell was the spectral response characteristic. It was considered extremely important that the cell have a spectral response similar to that of the normal human eye. Such a cell would be much more useful for measurements of color, and for obtaining control by color change, than are other commonly available photo-electric cells. The PHOTRONIC Cell meets this requirement very well, as will be noted from Figure 2. Its peak value is a little more in the yellow than is that of the human eye, and it has up to 25 per cent response in the near ultra-violet and about a 10 per cent response in the near infra-red. Since it is above the human eye over the visible range, it is possible to equip it with a filter that will make it agree with the normal visibility curve.

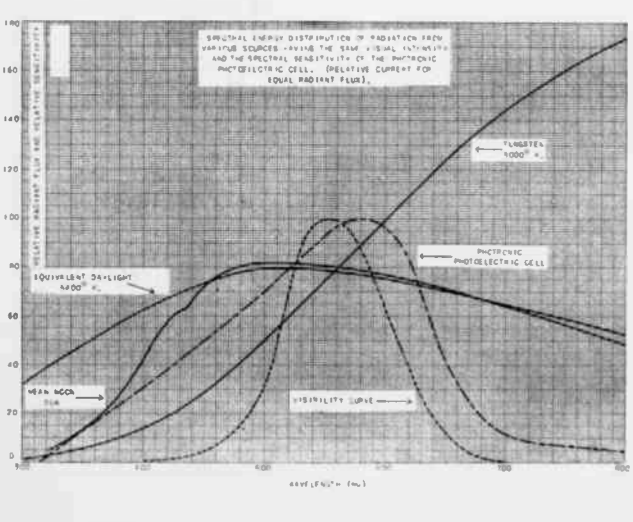

Fig. 2

Such a cell, in order to be useful in industry, should have a long life. The life of the PHOTRONIC Cell is practically unlimited, and its usefulness should, therefore, extend over a great many years.

The PHOTRONIC Cell is similar to a battery shunted by a high resistance. Due to the relatively low voltage generated, long leads may be used between the cell and the control relays without leakage affecting the accuracy of the control. This characteristic is very useful when installing control equipment in an industrial establishment. Also, like a battery, two or more cells may be connected in multiple to increase the output for a given light intensity.

In order to use this cell on measurement problems, whether they be the direct measurement of light intensity, or a measurement as applied to industrial process-control, it is only necessary to connect a microammeter, or relay, in series with the cell, no amplifying circuits being required as with most other photo-electric devices. In every other type of cell, except the photo-voltaic, an external source of electrical energy is required for purposes of excitation, and the accuracy of measurement, or control, is thus directly dependent on the constancy of the exciting power supply.

The PHOTRONIC Cell is almost instantaneous in its operation. One of the most interesting applications that has been made of this property, is the measure of the speed of rifle bullets by the Ordinance Department at Washington. Bullets are shot through several light beams, and the shadow of the bullet on the cell causes an indication on an oscillograph, which thus measures the relative speed of the bullet as it passes through these various beams. The cell has been used in sound-film work with a considerable degree of success, although, due to the fact that it is a good condenser, the output falls off at the higher frequencies. To compensate for this, changes in the conventional amplifying circuits are necessary and, in view of the number of these that are in use, it is not thought advisable to introduce the cell to this field at the present time. The quality of reproduction is very good, for the cell does not have the ground-noise that is present in the other types of cells, with which external excitation and high amplification are needed.

The circuits and circuit elements required for putting the cell to practical use are very simple. For measurement purposes it is only necessary to connect the cell directly to the terminals of a suitable microammeter. For control, the cell may be connected directly to the terminals of the sensitive relay. Usually, however, an additional and power relay is required for operating control devices. The power, controlled by the sensitive relay, and operating the power relay, may be supplied by batteries, or a transformer-rectifier unit, and thus requires nothing with which the average industrial electrical engineer is not thoroughly familiar.

MEASURING INSTRUMENTS

One of the first measuring instruments to be devised, including the PHOTRONIC Cell, was the illumination meter. Prior to its introduction, only the relatively crude comparison methods of the conventional photometers were available for illumination measurement. And, because of the complication of equipment, technique, and interpretation, the field of usefulness of such instruments was seriously limited by their cost, the need for skill on the part of the user, and in the interpretation of the data gathered. With the direct-reading illumination meter employing the PHOTRONIC Cell, however, these limitations are largely eliminated and the solution to a myriad of illumination problems, previously unattacked, made readily and economically available.

The Model 603 Illumination Meter, shown in Figure 3, was the first produced. It consists of two PHOTRONIC Cells connected in multiple, and mounted in a light target which is attached to the instrument by a 5 foot cord. Three ranges are obtainable in the simple instrument, and six ranges in some of the more elaborate instruments. The cells are some times equipped with filters which correct the spectral response to precisely that of the human eye, and enable accurate measurements to be made of any kind of light, such as that of neon, mercury vapor, or sodium vapor lamps, as well as tungsten, and other filament-type lamps.

Fig. 3

Another application of the photo cell for measurement is its employment for the predetermination of the proper exposure in photographic processes. All prior, and commonly employed, exposure meters relied on the judgment of the human eye to distinguish between various degrees of brightness, since it is the brightness of the object at which the camera is directed, that is, the light reflected from the object at which the camera is directed, that determines the rapidity of action on the sensitive surface of the photographic film, or plate, and hence the proper time of exposure. The human eye, however, like the other sense organs of the human body, is capable of perception of, and accommodation to, a tremendous range of the intensity of stimuli, and is similarly seriously limited in recognition of minor differences in the intensity of the stimuli. More specifically for instance, the normal eye is found to be quite capable of reading the larger sizes of news print type in bright moonlight, and in bright sunlight the range of light intensities of which is of the order of a million to one. It is to be expected, therefore, that the eye can hardly be expected to judge of minor differences of light intensity with sufficient precision to suit the relatively narrow requirements of light intensity of the photographic emulsions commonly used in photography.

When the development of the Exposure Meter was started, it was found that there were very little data available on the relative sensitivity of the film emulsions as expressed in terms of brightness. It was therefore necessary to obtain data on the sensitivity of emulsions before the relation between brightness measurements and exposure could be determined. This involved a lengthy investigation, and many films were exposed in order to determine the limits of "under exposure" and "over exposure." After this was once accomplished, it was a simple matter to make a calculator, or translator, that would interpret brightness measurements in terms of exposure time. In order that the instrument might measure brightness, however, it was necessary to restrict the exposure of the PHOTRONIC Cell to a 60° cone, since this is, roughly, the angle included by commonly used camera lenses.

Figure 4 shows the Model 617 Universal Exposure Meter. This meter has proven to be a major contribution to the photographic art, and is replacing the judgment and skill of many years of experience of some of the nation's outstanding photographers.

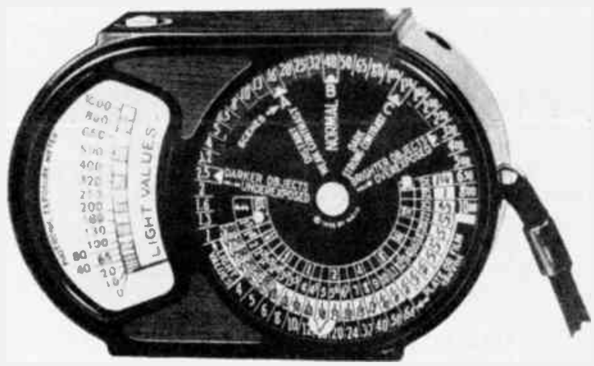

Fig. 4

A second type of Exposure Meter is provided for use with motion picture cameras, the scale of which is calibrated directly in terms of aperture for a given exposure, since in motion picture cameras the shutter operates at a substantially constant speed, thus requiring that the aperture be adjusted to accommodate the camera to the brightness to which it is exposed.

In various industrial processes, control may some times be accomplished by obtaining readings of changes in color or density of the material being processed. A simple combination of the PHOTRONIC Cell and an indicating instrument will give indications as to such changes in color or changes in density as a guide to process control, and many such devices are in use at the present time for such determinations.

PHOTO-ELECTRIC CONTROL

In controlling there are three basic principles that are involved. A light beam is a very useful link in mechanical motions, since it may be broken and re-established at will, for it has no inertia, nor does it mar delicate finishes. Furthermore, when using a cell that has a linear characteristic, it is possible to operate at such a point in its characteristic curve that control may be accomplished by slight variations in intensity. With a spectral response characteristic, such as that of the PHOTRONIC Cell, it is also possible to effect control by means of changes in color.

Thus, automatic control of processes may be obtained by:

(1) A cut-off of a light beam

(2) Change in light intensity

(3) Changes in color

One of the applications of a photo-electric cell that is usually thought of first, is the question of counting. Small pieces of material, such as paper, coming down a conveyor, that may be located in almost any position, cannot be counted readily by any mechanical means. A mechanical finger that might be placed in the path of the material, could not be operated from a small piece of paper, and the location of the other material might not be such that it could be accomplished without a very complicated arrangement. Also, highly finished materials that would be scratched by the mechanical finger may be counted by an interrupted light beam, and in no other manner. For the purpose of counting an especial system has been designed, and made commercially available. It consists of PHOTRONIC Cells in connection with suitable relays, and will operate the commonly-available electrically-driven counters.

Having a light responsive unit such as the PHOTRONIC Cell, it is natural that the question of illumination control should be considered. Street lights and sign lights have always been controlled by time switches. Time, however, does not take into consideration the changes in light intensity from day to day due to clouds, storms, and other natural conditions. Consequently, all too often, lights are not turned on when they are needed, nor are they turned off when they are not needed. It is natural, therefore, that an illumination-control device for street lighting, sign lighting, and also for interior lighting, in factories and office buildings should be considered. For this purpose an especial illumination control system, employing the PHOTRONIC Cell, has been developed, and has been applied quite successfully to many lighting control purposes. A large number of beacon lights on the American Airways are equipped with these units, so that the beacon lights are turned on whenever the daylight intensity drops low enough. Many lighthouses are equipped with these devices so that the ships may have proper guidance when daylight intensities are low. On such safety devices, it can be readily understood that automatic control in accordance with daylight intensity is the only type that is adequate. Similarly, street lighting circuits are being controlled by this type of equipment.

In this type of control equipment, the relay arrangement is slightly different from that used in counting. Counting equipment should operate quickly, but illumination control, where light intensity is changing slowly, requires a time delay so that when the lights are once turned on, they will stay on, and will not be effected by slight fluctuations due to passing clouds, and other atmospheric conditions. In order to accomplish this, a time-delay relay is placed in the circuit between the sensitive relay and the power relay.

A control that illustrates the principle of change in density is the smoke alarm. On this device a light beam is thrown across a smoke stack to a PHOTRONIC Cell, and when the smoke in the stack takes on a predetermined density indicating poor combustion, an alarm is given in the boiler room so that the attendant may correct the condition. By the use of such arrangements as this, not only is the industrial establishment assuring itself of the operation of its steam-raising equipment at greatest efficiency, but also it frees its community of the smoke nuisance otherwise difficult of riddance.

It may, therefore, be seen from the general discussion given above, that the combination of a PHOTRONIC Cell and two or three relays, make possible many types of control functions.

A new relay, that is far more sensitive than any other previously available, has recently been developed. This relay will make a positive contact at such low currents as 2.5 microamperes, and at 0.5 millivolt. It opens many new fields for the PHOTRONIC Cell. It is really a contact-making indicator so that indications of the conditions that exist even before contact is made, are evident from the scale of the device. The contacts will control a circuit of 5 watts at 110 volts. By using this new type of relay in conjuntion with a PHOTRONIC Cell, many many simple controls or alarm devices are possible.

Typical of the applications of the PHOTRONIC Cell are the following: Pieces of paper on a folding machine are being counted--this cannot be done in any other manner. Bags of sugar are delivered to the warehouse on a conveyor belt, and they do not necessarily place these on the same point on the belt every time. It is possible to count them by an interrupted light beam at a high degree of accuracy. Small pocket combs are being counted.

Another interesting application is on an elevator-leveling device. Through its use the elevator is brought quickly to the proper floor level with no special skill on the part of the operator.

In the Holland Tunnel is an installation for detecting any truck that attempts to go through there loaded too high. Frequently the main part of the load on the truck may be of the proper height, while some small piece projects above the rest of the load. This would jam in the tunnel, and might cause a traffic tie-up. A light beam is projected across the driveway, which is about 35 feet wide, on to a PHOTRONIC Cell in connection with an alarm circuit. When tested, it was possible to travel at 35 miles an hour, and detect the presence of a projecting unit that was only 1 inch wide. Since the detector is located at the toll booths, where the speed of the traffic is naturally only a few miles per hour, the system is extremely effective in detecting all projections above the allowable levels.

Another installation in the Holland Tunnel, is for detecting the presence of smoke. In this installation the PHOTRONIC Cell is about 5,000 feet away from the recorder that it operates. In this case, as in the case of many industrial applications, where it is necessary to locate the relays, or controlled equipment, remote from the source of control, the PHOTRONIC Cell showed itself to be especially effective.

PHOTRONIC Cells installed on weighing machines operate relays when a container is filled to the proper weight and the feed is automatically shut off.

In Detroit an automobile comes down the way to have the underside of the chassis sprayed. It is a completely assembled car, and care must be taken to see that the paint is not sprayed on the finished body. The car interrupts a light beam, and the PHOTRONIC Cell turns on a spraying machine, which covers the underside of the chassis, and yet, as soon as the car moves through this light beam, it stops immediately, and prevents spoiling the finish on the body.

In one automobile plant, there is a carrier for delivering the bodies to the assembly room. Oftentimes the chassis is not there for the body to be put on, and one will pass the point at which it is usually removed. These bodies used to go to the end of the conveyor and then would hit the wall, thus damaging them. A light beam was placed at a point just before the body would reach the wall, and if it were not removed before it got that far, the conveyor was stopped.

In a sewage plant a PHOTRONIC Cell and light beam were placed 5 feet below the surface of the water in the last settling tank. As the sludge level rose above this point, an air lift valve or pump was operated so that this sludge would be removed and placed back into the system. By using this automatic equipment, it was possible to operate this plant at a considerable overload with complete satisfaction.

In a film processing laboratory in the motion picture industry, the hypo solution that is used for fixing the film, becomes very rich in silver. For satisfactory operation, the silver should be removed and at the same time the amount of the reagent employed must be controlled so that the hypo is not seriously contaminated. This was accomplished by the PHOTRONIC Cell operating on the change in opacity of the liquid as the silver was precipitated.

Boiler-feed water has to be treated for the elimination of oxygen. This is done by the colorimetric method of hydrogen ion concentration control. The PHOTRONIC Cell watches the change in color, and adds just enough material to the boiler-feed water to eliminate the oxygen, and not enough to cause scale.

In roasting coffee, it is necessary that the degree of brownness be controlled very accurately. The PHOTRONIC Cell watches the change in color of the coffee bean, and when the proper brown has been reached, the alarm is given so that the operator may dump the roast. This is a very critical point, and considerable spoilage results when observation of the operator is depended upon completely. In order to obtain a high degree of sensitivity in this control, an interesting and effective expedient was resorted to in this installation. A coffee bean unroasted is green, and in its roasted condition is a deep brown. Brown color may be obtained by adding red to green. By placing a red filter over the cell, it was possible to control the amount of red that was added to the green to secure the desired brown with an unusually high degree of precision.

The PHOTRONIC Cell has opened many new fields of photo-electric measurement and control. The simplicity of the circuits involved, and the use of standard instruments and relays operating on well-known principles, largely eliminates the hesitancy of the industrial engineer toward the adaptation of photo-electric means to his control problems. New devices and methods will be developed, and the light beam will ultimately take its place with the gear, the lever, and the cam in the machinery used for the manufacturing processes of the world.

AUTOMOBILE RADIO RECEIVERS

The meeting of March 14, 1934, was devoted to a paper by Mr. J. T. Filgate, of the United American Bosch Company, Incorporated, on the subject of automobile radio receivers. Mr. Filgate discussed the problems of the design, installation, and operation of automobile radio receivers with especial emphasis on the causes and sources of ignition interference with radio reception, and those expedients which have been applied to the problem for the reduction, or elimination, of this type of interference. His paper included an unusually detailed analysis of those phases of automobile ignition phenomena, which give rise to interference, and pointed out in specific detail the peculiarly troublesome nature of this type of interference as well as the expedients, both useless and useful, which have been resorted to for its elimination. The influence of the type of location of the antenna on the intensity of the interference was brought out, and an explanation of the operating characteristics of several of the more unusual types of automobile receiving antenna was given. Of especial interest was a description of the influence of the bonding of the automobile body, and of the electrical wiring on the magnitude of the interference.

General discussion of the entire problem followed the paper, indicating a general appreciation of the part of the membership of the masterly treatment of the analysis given by Mr. Filgate.

REMOTE CONTROL OF RADIO RECEIVERS

A symposium on the subject of remote control of radio receivers constituted the meeting of April 11, 1934. Papers were read by Messrs. Virgil Graham, and Lee McCann, of the Stromber-Carlson Telephone Manufacturing Company, and by Mr. C. J. Franks of the Radio Frequency Laboratories. Mr. McCann's paper was devoted to a discussion of the type of equipment in which remote control is accomplished through the use of electro-mechanical means on centrally located selecting and amplifying equipment; while the papers of Messrs. Graham and Franks were devoted to a discussion of what has been characterized as remote tuning, in which selection, and high-frequency amplification, is accomplished at a point remote from the low-frequency amplification and reproducing equipment. Mr. McCann's paper was replete with photographic illustrations indicative of the especial adaptability of the systems described by him to incorporation in structural elements of homes. The papers of Messrs. Graham and Franks were especially detailed with respect to the problems met with in the incorporation of high-frequency transmission lines in the remote tuning type of equipment.

The discussion called attention to other types of remote control equipment, one of which was illustrated by a special demonstration by Mr. Franks.

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THE IMPORTANCE AND TECHNIQUE OF PERFORMANCE MEASUREMENTS ON RADIO-TELEPHONE TRANSMITTERS

W. C. LENT *

The transmission of intelligence by radio, whether that intelligence be in the form of an entertainment program or a definite message, depends upon some form of modulation of a so-called carrier wave. By modulation is meant the process by which the instantaneous character of the carrier is made to change in accord with the variations of the intelligence to be transmitted. The character of the modulated wave may be completely defined for a particular instant if the phase, frequency and amplitude at that instant are known. For radio-telephone transmission, at least, experience has seemed to indicate that the most suitable form of modulation is achieved by varying the instantaneous carrier amplitude in response to the variations of the signal energy to be transmitted. Experience has likewise indicated that it is desirable to maintain the carrier frequency and phase as nearly constant as possible during the modulation cycle.

Amplitude modulation of a constant carrier-wave by a complex signal wave results in the production of side-bands of variable width and amplitude. These side-bands are displaced from the carrier on both sides by a frequency interval equal to the modulation frequency at a particular instant. The amplitude of the side-bands is determined solely by the degree of modulation.

Since the intelligence-bearing portion of the modulated wave is not the carrier but the side-bands, it follows that the greater the side-band amplitude the greater is the ratio of useful power to the total output power. For a completely modulated wave, that is one modulated 100 per cent, two-thirds of the total power is concentrated at the carrier frequency with the remaining one-third in the side-bands. The signal energy varies between this one-third maximum and zero directly as the square of the modulation degree.

If the side-band power is calculated as a function of modulation degree it is found that the following relations hold:

MOD. DEGREE	% TOTAL POWER CARRIER	% TOTAL POWER SIDE-BANDS
0	100	0
25	97	3
50	89	11
75	78	22
100	66.6	33.3

In a wave modulated 25% only 3% of the total power is useful signal producing power. A 100% modulated wave delivers 11 times as much signal energy as one modulated 25%. Certain general rules may be set down as follows:

1. Doubling the modulation degree for a given carrier power doubles the signal level without increasing the nuisance field.

2. Doubling the carrier power while holding the modulation degree constant results in a signal level increase of only 41.4 per cent and in an equal increase of the nuisance field.

3. To obtain a signal level increase equivalent to that obtained by doubling the modulation degree, holding carrier power constant, requires a carrier increase to a value four times the original, maintaining the same modulation degree. In other words, doubling the modulation degree pays the same dividend in signal level as a four to one carrier power increase.

Thus it appears that both the relative and absolute efficiencies of radio telephone transmitters are directly dependent on the modulation degree and will be the maximum obtainable for the condition of complete modulation.

Were efficiency the only condition with which the station engineers are concerned it would be a simple matter to adjust input signal level to a value sufficient for complete modulation and let matters take their natural course. Unfortunately

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the processes whereby modulated radio frequency energy is generated and amplified involve the use of elements which inherently are non-linear. Such non-linearity results in the production of distortion of the original signal wave which appears in the form of measurable harmonics of the signal frequencies. For high quality transmission these harmonics must be reduced to limits at which the distortion is not perceptible to the ear. In the processes under consideration this distortion is primarily an amplitude function, that is, the harmonic amplitudes depend on the length of the portion of the non-linear characteristic traversed during the modulation cycle. This, in turn, is directly a function of the modulation degree and we are thus brought to the term "modulation capability". As defined, the "modulation capability" of a transmission system is that modulation degree which can be attained without exceeding the distortion limit which has been determined to be acceptable for a particular class of service. "Modulation capability" is not a term which can be indiscriminately applied without specific stipulation of the acceptable distortion limit for the class of service in question.

Useful modulation and the only modulation in which operators of high quality radio-telephone stations can be interested, is that modulation which does not exceed the capability of the transmission system. What that capability must be can only be determined after the permissible amount of distortion has been ascertained. This capability as determined will not represent efficient use of channels or transmission plants unless complete modulation can be attained. At present, the minimum capability acceptable to the Federal Communications Commission is that in which a modulation degree of 75% can be attained without the production of combined harmonics in excess of 10%. It is certain that with increasing activity in the field of high fidelity this requirement will be made more stringent with respect to the allowable harmonic content, and perhaps, in the interest of efficiency to the absolute modulation degree attainable as well.

The acceptability of the service of any station depends, aside from purely artistic considerations, upon the ratio of the wanted signal output to the total extraneous noise output at the receiving loudspeaker. Measurements seem to indicate that this ratio must be at least 1000 to 1 for high quality service. Extraneous noise originates at several points in the transmission system. Switching transients, shot-effect in vacuum tubes and thermal agitation in associated tube circuits, insufficient filtering of power supplies and modulation by mechanical shock all combine to cause noise modulation products to appear in the output wave. At the receiving end of the system local interferences from various discharge devices, loose and vibrating contacts between metallic surfaces and poor filtering of receiver power supplies combine with interchannel interference and heterodyning effects to increase the apparent carrier noise on the wanted signal. The transmission engineer can control the absolute level of the noise originating in the transmission circuits by proper design and treatment at the source. Furthermore, the relative noise level may be reduced by maintaining the modulation as nearly complete as possible. While control at the source of the noise originating in the transmission medium and the receiving end of the system is denied, inso-far as absolute noise level is concerned, the relative level may be reduced by consistent maintenance of high modulation degree at the transmitter.

It would appear, then, that the case is clear for the requirement that the modulation capability of transmission systems be raised to the absolute physical maximum as regards modulation degree and that under this condition of complete modulation the total harmonic content be within the limits found to be acceptable for the particular class of service. High quality broadcasting would seem to require that total distortion be less than 5%. In no case should the practice of reducing distortion at the expense of absolute modulation degree be followed.

That the need exists for improvement all along the line is amply indicated by the results of measurements made on 69 stations East of the Mississippi by our engineers last year. In most cases the figures represent actual capability of the transmission system defined as that modulation degree which could be obtained without noticeable shift of the mean carrier amplitude during the modulation cycle. The stations are rated in five classes as follows:

CAPABILITY RATING	% IN EACH CLASS
Less than 25%	4.3
25 to 50%	34.8
50 to 75%	47.7
75 to 100%	10.2
100%	2.9

Included in this group are fifteen 100 watt, eight 250 watt, twelve 500 watt, sixteen 1 kilowatt, seven 5 kilowatt, one 10 kilowatt, three 25 kilowatt and seven 50 kilowatt stations.

The average capability for the whole group is 60% with the average within each power class about the same throughout.

It is not and cannot be the purpose of a paper of this nature to treat rigorously with the exact technique involved in the making of performance measurements on transmission systems. Rather it is desired to review the current methods in a general way and to indicate the possible inaccuracies of each together with the requirements which must be met if reliable results are to be obtained.

1. Modulation Measurement:

The measurement of the degree of amplitude modulation involves not the measurement of a physical quantity but rather a ratio. This ratio exists between two quantities in the same dimension and hence may be determined at any absolute level.

Methods of modulation measurement fall into two classes depending upon whether measurement is made before or after demodulation of the envelope of the output wave.

Into the first class, that is where measurement is made prior to demodulation, fall the methods involving the cathode-ray oscillograph, direct-reading peak voltmeter and the thermo-galvanometer.

The cathode-ray method permits the observation of a trapezoid the shape and dimensions of which furnish some information. This method has two inherent drawbacks. Accurate scaling of image dimensions is difficult if not impossible. In a pattern 1 square inch in area the normal double trace width can be responsible for an error of approximately 12%. Furthermore, as the operator proceeds in measurement from the condition of no modulation to the desired test condition with modulation it is impossible to detect a shift of the boundaries of the unmodulated pattern. Since the width of this pattern is the reference level of the unmodulated carrier, it is essential that that width remain constant during the modulation cycle. The process of measuring the long side, subtracting the length of the short side and dividing by 4 is fallacious in that it yields absolutely no information as to the symmetry of the modulation. Performing this process on a pattern obtained under the condition of 25% positive and 75% negative modulation yields the mathematical value of 50% which in the absence of evidence to the contrary is taken to mean that a symmetrical modulation of 50% is actually being obtained.

The only way in which such a method can be trusted to yield reliable indications is to use an auxiliary carrier-shift indicator. Pursuing the use of the method beyond the point where carrier-shift becomes evident is useless.

The direct-reading peak voltmeter used either before or after demodulation possesses the same inherent fault. It has an advantage over the oscillographic methods in that it is not subject to scaling and trace-width errors.

Perhaps the worst method from the standpoint of possible error is that in which a thermo-galvanometer is used as a direct indicator of modulation degree. True, on purely sinusoidal test tone and in the complete absence of carrier-shift the indications thus obtained will furnish valuable information simply and quickly.

The type of modulation here discussed is an amplitude function. Maximum amplitudes must be determined if proper evaluation of modulation degree is to result. It is obvious that this method fails in the presence of complex wave modulation since only effective and not maximum values are indicated by a thermal instrument. The blind use of such a method with complex wave modulation and in the absence of carrier-shift indication can lead only to results erroneous in the extreme.

Methods of measurement which involve the process of demodulation possess all of the foregoing possibilities for error plus that resulting from non-linearity in the demodulator.

Perhaps the most satisfactory and accurate modulation measuring equipment is based on the principle of measurement after demodulation and incorporates a carrier-shift indicating device together with some form of peak voltmeter--preferably direct reading.

2. Distortion Measurement:

The determination of the modulation capability of the transmission system necessitates the simultaneous measurement of both modulation degree and harmonic content. The methods of modulation measurement have been outlined.

Total harmonic content may be evaluated as a lumped quantity directly or by the mathematical combination of the various harmonic amplitudes measured separately.

For general use in which quantitative results are sufficient the method of determination of distortion as a lumped factor is most satisfactory. Briefly, the method consists of segregating the harmonics from the fundamental by means of differential filter systems and evaluating the total harmonic root-mean-square amplitude in terms of the fundamental amplitude. The result multiplied by 100 yields the harmonic content as a percentage directly. Granted that the filter systems are correctly designed as to cut-off, image impedance and working level it is still necessary to be certain that the demodulator preceding the differential system is strictly linear and that the amplifier interposed between the filters and the output indicator is linear with respect to both amplitude and frequency. The output indicator must be a strictly square-law device if the true root-mean-square value of the harmonics is to be indicated. Furthermore, the source of test frequency must produce an absolutely pure wave.

This method has the disadvantage that it is limited to measurements at a single frequency. Unless the elements in the transmission system remain constant with frequency, measurements of capability at one modulation frequency do not necessarily indicate the true capability at any other. If, however, the system will modulate consistently in the absence of carrier-shift over the entire signal-frequency range, it is reasonably safe to rely on single frequency distortion measurements as a true indication of capability.

For rigorous measurements of distortion some form of wave analyzer is necessary. Separate harmonic amplitudes can be measured and combined mathematically as the square root of the sum of the squares of each to yield the true value of the combined distortion. Of course, the same requirements are placed on auxiliary equipment as in the case of the first method.

A carrier-shift indicator, while not capable of measuring distortion, is nevertheless the most reliable and fool-proof tool available for capability determinations. Carrier-shift is absolute evidence that non-linearity and hence distortion is present in the system. The converse is not necessarily true because a perfectly linear system will still pass freely all distortion present in the source.

Unfortunately the correlation of carrier-shift with quantitative distortion is not as simple as it would appear. A given carrier-shift at one modulation degree represents an entirely different distortion condition than does the same amount of carrier-shift at some other modulation level. If a carrier-shift indicator is to yield quantitative results it will be necessary to provide a complete family of characteristics of distortion versus carrier-shift for many convenient values of modulation degree. Measurements made so far indicate that this is entirely practicable. In general, the lower the absolute modulation level the greater the distortion for given amounts of carrier-shift.

Fourier analyses of oscillographic traces are limited for quantitative purposes by the same inherent inaccuracies indicated in the discussion of oscillographic devices as modulation indicators. If anything, errors are even more serious from these sources in distortion measurement than in the case of modulation measurement.

3. Carrier Noise Measurements:

Carrier noise can be measured directly as a modulation degree of small magnitude or in terms of a ratio to a given reference modulation level.

The measurement of noise directly and quantitatively involves tremendous difficulties if gross inaccuracies are to be avoided. When it is realized that a noise level 60 Decibels down from a 100% modulation level corresponds to an absolute noise modulation of only 1/10 of 1% the problems involved in such methods of measurement will be better recognized. Furthermore, noise in the measuring equipment itself can, without careful design, easily be greater than the noise which it is desired to measure. It is not felt that direct measurement of carrier-noise is practical.

Relative or comparison methods have the advantage that resistance attenuators, which can be made exceedingly accurate, can be used. They form the only variable element in a properly designed system. Errors from noise in the measuring set-up cancel out in the process of comparison measurements. Furthermore, the range of an attenuator can be carried to limits permitting measurement of magnitudes far beyond the capability of direct methods. It is a simple matter to convert a ratio in terms of Decibels to the actual noise modulation level if desired.

If such a measuring system is to be used certain precautions in design are necessary. The system as a whole must have a high, stable and controlled gain. The frequency response must be uniform and comparatively wide--preferably from about 25 to 10,000 cycles per second--due to the fact that tube noises and circuit noises have components lying in the upper speech frequency range. The variable element, the attenuator, must be capable of variation in small steps--say 1 Decibel. The output indicator must possess a strictly square-law characteristic if the true root-mean-square value of the noise modulation products is to be accurately indicated.

If high quality broadcasting is to progress as much in the future as it has in the past a rigorous routine of performance measurements must be a definite part of the duties of the operating staff. New wide range receivers, when and if they are made available to the public are bound to cause adverse criticism of many of the services now being accepted as satisfactory. Not only must technical equipment of suitable design be provided the operating engineer but some judgment be used in the initial selection of operators. This will require a more careful attention to the capabilities of persons supervising operation at transmitting plants. Too many now so employed are content to throw the burden of monitoring programs on the control room staff. It seems that a transmitter operator should be held responsible for overmodulation and poor quality originating beyond the control of the studio personnel.

DISCUSSION

At the conclusion of the reading of the paper the meeting was thrown open to discussion by President Houck.

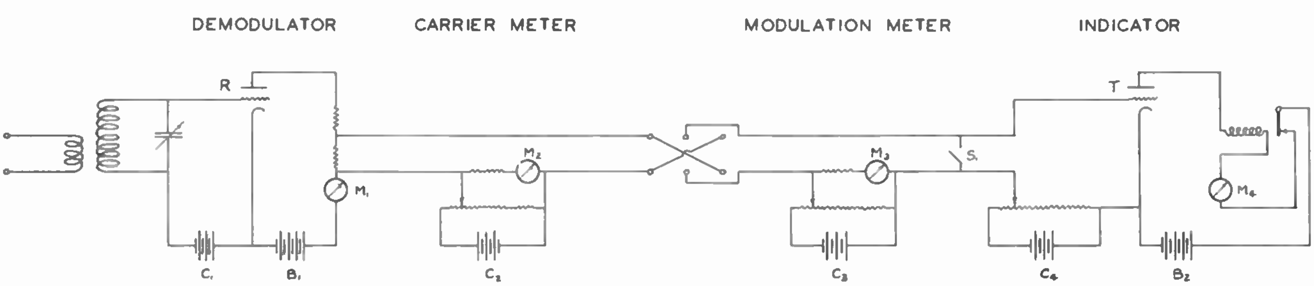
Inquiry was immediately directed toward the details of the apparatus suited to making the measurements the need for which was the point of Mr. Lent's paper. In reply to a specific request as to the nature of the apparatus used for measurement of carrier magnitude shift, Mr. Lent showed the following circuit arrangements and discussed their functions and usefulness.

The modulation meter consists essentially of a linear rectifier and appropriate circuits and apparatus for the measurement of the ratio of that portion of the rectifier output which is proportional to the carrier to that portion of the rectifier output which is proportional to the low-frequency variation of the high-frequency envelope resulting from the process of modulation. It differs largely from circuits and apparatus commonly employed for this purpose in the use of a triode rectifier, and in the use of a thyratron as the indicator of balance in the measuring system.

It was pointed out by Mr. Lent that while the high internal impedance of the conventional triode--as against that of a diode of the same dimensions more commonly used for this purpose-- requires that an extremely high resistance be employed for the plate load of the triode, with the consequent problems incidental to the maintenance of the insulation resistance at appropriately high values, the advantages to be gained through the isolation of the plate, or measuring circuit from the grid, or signal circuit, well justified the employment of the necessary precautionary measures in the design and the operation of the device.

The operation of the thyratron indicator circuits was explained as follows: Since the problem of the measurement of the modulation degree is essentially that of the balancing of the several potentiometer voltages with those available at the rectifier output, the problem reduces itself to one of determining a voltage balance. In the past simple indicating instruments have been used for this purpose, and have brought with their use several unavoidable limitations on the precision of the determination of balance. Outstanding amongst these, is the fact that where an instrument of sufficiently low current sensibility to avoid the need for frequent replacement is used, the determination of balance is necessarily poor, and, in any event, conventional meters all suffer from so high a degree of inertia that they are useful only when a repetitious form of modulation is employed, such as simple tone modulation.

Where the degree of transient modulation, such as speech and music, is to be measured, conventional measuring instruments, per se, are completely unsuitable and an indicator of negligible inertia is essential. Such a low inertia indicator is provided in the thyratron. When once the potentiometer, C_4 is so set that plate-circuit current barely does not flow, only an extremely minute reduction of the thyratron grid bias, for the barest instant of time, is required to initiate plate current after which it continues to flow until manually or automatically interrupted, and the previous critical conditions reestablished.



Provision for the preliminary adjustment of the thyratron operating conditions is provided through the use of the isolating switch, S_1, which when closed provides for the adjustment of the battery, C_4, to the lowest value of negative bias on the thyratron grid at which no plate current flows. The relay in the plate circuit provides a closed circuit for the plate current that will later flow and make itself evident by the meter, M_4, and the operation of the relay.

Once this adjustment has been made, the balance of the local d-c voltage available from C_2 for the determination of the relative value of the carrier amplitude, and the balance of the d-c voltage available across C_3 for the determination of the relative value of the modulation amplitude, may be determined with unusually high precision.

The scales of meters M_2 and M_3 are so chosen as to give a direct reading of the per cent modulation on M_3 .

Mr. Lent pointed out that not only did the equipment provide for the measurement of the degree of modulation, but provided for the determination of frequency with which the modulation exceeded any value for which the apparatus might be set. The thought was contributed by the membership that by the addition of a simple electrical counter to the plate circuit of the thyratron, the system might then integrate the number of such excesses of modulation, and by the further addition of an electric clock movement carrying appropriate recording discs it might time such excesses.

Considerable discussion of the degree with which so essentially a power series device as the electron tube could approach strict linearity brought out from Mr. Lent the fact that, in so far as his measurements were able to reveal it, there was no departure from strict linearity in the triode rectifier included in the equipment.

The discussion soon turned to the matter of high fidelity transmission in connection with which it was brought out that broad frequency response is only a part of the high-fidelity program, and that of almost equal importance is the matter of dynamic range. It was recalled that during the Bell Laboratories demonstrations last spring it was shown that a full orchestra has a dynamic range of at least 70 db. As against this, the average broadcasting station can transmit a dynamic range of only about 20 db.

The studio operator, in a modern broadcasting plant, is the man who is charged with the responsibility for seeing that the actual dynamic range that goes out from the studio to the station does not exceed the capabilities of the station. To do this successfully requires rehearsal of the program, so that the points of greatest intensity or highest level may be noted on the cue sheet, and proper notes made as to how the level is to be controlled during the program. The level must be cut down gradually and well in advance of the loud passages, if the audience is to get as much of the dynamic range as the apparatus will permit without overloading and distortion either at the transmitter or receiver.

At the upper limit of the dynamic range there is the modulation capability of the transmitter. At the lower limit there is the noise level. If these were, let us say, 100 db apart, so as to leave a factor of safety over anything that even the full orchestra can do, the measurements outlined in Mr. Lent's paper would be far less important. The fact is, however, that the dynamic range of the transmitter is less than one-third of what it should be, even when all the equipment is in the best possible order. Additionally, the modulation capability tends to fall, and the noise level tends to rise, unless the station is expertly maintained, and this requires frequent and precise measurement of each of these quantities, with the most accurate and convenient apparatus that can be secured.

Nothing can be more detrimental to the full enjoyment of the best programs than to have overmodulation and distortion occur, even if only for an instant, at those very points in the score that represent the climatic heights of the music. To have the quieter and more contemplative passages obscured below the noise level is only a slightly lesser evil. Expert handling of the studio gain controls. based on a complete and current knowledge of the capabilities of the station, is therefore a primary requisite in the high-fidelity campaign.

Mr. J. H. Miller

The discussion has brought out the fact that the majority of rehearsed programs will have a cue sheet correctly marked with the position of the gain control for the various parts of the program. The gain control position is presumably set for that value which gives maximum modulation without exceeding 100% at any time due to orchestral crashes or other peaks.

At the same time it must be realized that many programs can not be rehearsed. This obviously applies to running accounts of sports events and to certain studio features. It is here that the volume level indicator, integrating over certain time periods, becomes of value.

The device described by the speaker which indicates that modulation has exceeded 100%, indicates only after the event occurs; the commercial volume indicators indicate continuously.

A considerable gain in the available control is now had through the use of level indicators with varying speeds. Through the use of a pair of these instruments, one slow speed unit with a period of about 1.5 seconds and the other a high speed unit with a period of about 0.3 second, a reasonably good control may be had over a program. The instrument with the longer period is usually kept as high as possible so that the maximum use of the facilities may be had. The high speed instrument which will indicate peaks must be kept below the overload point, usually arranged well up on the scale, and by thus watching both instruments a reasonably good monitoring job may be done. It would seem that with these two instruments, a cue sheet, and a device giving a record of overload peaks, we would have the best possible arrangement of equipment to give a high quality program and at the same time make use of the available facilities to the utmost.

It is believed that the speaker's statement that absolutely no overloads should even be tolerated would result in operating a station at too low a modulation level. If we will allow an occasional modulation peak of over 100% we will immediately arrive at a reasonable compromise between the maximum use of facilities and a program of general high quality and such a reasonable compromise would seem to be the optimum condition.

Mr. W. C. Lent

I have read Mr. Miller's discussion and it occurs to me that he missed somewhat the purpose for which the modulation measuring system discussed was intended.

That apparatus is not intended for everyday monitoring purposes but rather to enable the supervisory staff to check the operation of a station in the hands of the operators. For station use an entirely different type of modulation meter is more suitable. This instrument actually indicates modulation degree directly and continuously during the program. Obviously this meets the objection of Mr. Miller. The system under discussion was not presented as one suitable for monitoring purposes which fact apparently Mr. Miller did not understand.

Mr. E. A. Tubbs

It seemed to be the consensus of opinion that a transmitter should be allowed to pass beyond its modulation capability on occasional signal peaks, the point at issue being to what extent it should be allowed to so overmodulate.

From personal observation I have found that a station, which may seem to have no noticeable distortion when properly tuned in, may at the same time be causing serious interference with a weak signal on an adjacent channel. It seems to me that this should be given serious consideration when deciding how far and how often a transmitter be allowed to overmodulate. If this adjacent channel interference is kept to a low value then I do not believe a station need worry about the amount of distortion its signal is undergoing.

This seems to me to be doubly important in as much as the lack of fidelity with which a station reproduces its program is mainly harmful to itself; but when it creates adjacent channel interference it is in effect destroying the fidelity of stations on neighboring channels, for listeners located near the offending transmitter.

While it may be true that there are very few receivers capable of receiving a weak signal on a channel adjacent to that of a powerful local station, yet such receivers can be built and I believe there would be more creative effort expended in this direction if the prevalence of adjacent channel interference were given such more serious consideration as would minimize or eliminate it.

JOINT MEETING

ROCHESTER, NEW YORK
November 12, 13, 14, 1934

RADIO CLUB OF AMERICA

INSTITUTE OF RADIO ENGINEERS

ENGINEERING SECTION R. M. A.

Program of the joint meeting will be forwarded to the membership shortly. In the meantime hotel and other reservations for the meeting should be made by addressing

MR. O. L. ANGEVINE

Rochester Engineering Society

HOTEL SAGAMORE ROCHESTER, N. Y.

Year Book

In view of the fact that 1934 marks the twenty-fifth year of the Club's activities, the Board has determined to commemorate the occasion by a "Twenty-Five Year Book".

That publication will supplement the material of the usual form of year book by a detailed and intimate history of the development of the radio art and industry as participated in by the club members. A questionnaire devised to serve as a basis for such an historical volume has already been distributed to the membership and all are urged to supply the desired data at once. In addition to this data any photographs which will more graphically indicate the part played by the membership in the development of radio are urgently solicited.

June Meeting

The meeting of June 13th was devoted to a paper by Mr. W.G.H. Finch on the adaptation of the conventional Teletype printers to the radio system of the International News Service and on the compact radio printer which he has developed for mobile work.

The membership will be glad to know that one of its number, in the person of Mr. Finch, was appointed to a place on the engineering staff of the Federal Communications Commission and that while the taking over of his new duties has made impossible the completion of his paper for publication as originally scheduled, it is expected that it will appear in the Proceedings shortly.

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ALL WAVE RECEIVER PROBLEMS

BY

MURRAY G. CLAY*

Delivered before the Radio Club of America
October 10, 1934

CIRCUIT DESIGN PROBLEMS

The important requirement that radio receivers designed for use in the high frequency ranges provide the same effectiveness of performance as has become commonplace in the receivers designed for use in the normal broadcasting range requires something more than the extension into the higher frequency ranges of the conventional circuit design methods which have made the broadcast receiver possible. It has, on the other hand, required a careful re-examination of the principles underlying those methods of design and the working out of new and especial applications of those fundamentals to the specific conditions encountered in the high frequency ranges.

The mere fact that receivers of this type are necessarily of the multirange type introduces the complicated problem of switching of high frequency circuits completely lacking in broadcast receivers: it introduces the extremely serious problem of adapting the highly standardized components of the relatively low frequency broadcast receiver to the high frequency field and introduces the unusually troublesome problem of providing for the use of a single antenna over so wide a frequency range as to make the conventional receiving antenna a negligible portion of a wave-length long at one end of the range and perhaps more than a complete wave length at the other.

Early attempts to provide economically usable multirange receivers in this field completely evaded many of these problems by the omission of all selection and amplification between the antenna and first detector. The extremely unfavorable image response, the relatively low signal to noise ratio and the usually low sensitivity and effective selectivity of such arrangements quickly proved them inadequate and pointed for the need in the high frequency ranges of circuit elements analogous in their performance characteristics to those employed in the broadcast range.

Thus, experience has established the fact that present tubes can be made to function as radio frequency amplifiers with a much more favorable signal-to-tube-hiss ratio than can be attained in first detector or converter circuits and a marked increase in usable receiver sensitivity is therefore obtainable when one or more high gain radio frequency stages are arranged to supply an augmented signal voltage to the converter grid thus "swamping" the noise contributed by the latter tube. Image ratio is improved, as radio frequency tuned circuits are cascaded, proportionally to Q_n where "n" is the number of tuned circuits and "Q" is the figure of merit of each, assuming they are alike. At the same time, where reasonable shielding is employed, one or more radio frequency tuned stages afford acceptable freedom from interference caused by strong signals or electrical noises getting directly into the intermediate frequency amplifier. Radiation is likewise materially reduced since tuned circuits and tubes, preceding the oscillator section of the circuit, act as effective "buffers".

The importance of antenna coupling circuit gain cannot be over emphasized since, in addition to all of the above advantages, this factor determines the ultimate usable sensitivity of the receiver in favorable locations. While tube noise in a radio frequency amplifier stage may be markedly less than that contributed by a converter stage, this noise limits the amount of usable amplification, of any type, which can follow this tube. Antenna circuit gain, then, is of unique importance and every effort to maximize it without complication, is worth while.

INTERSTAGE COUPLING SYSTEMS

Early study of these factors led, immediately, to the measurement of the relative figures of merit, "Q", of some twenty coils, of suitable physical sizes, designed to operate between 8 and 20 mega-

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cycles in conjunction with standard broadcast type tuning condensers.

APPROXIMATE VALUES OF "Q" FOR VARIOUS PRACTICAL COILS AT EIGHT MEGACYCLES

Coil #	"Q"	Dia.	Turns	Wire #	T / inch	Comments
1.	101	1.0	5.0	18	20	
2.	116	1.0	5.5	18	10	
3.	358	1.25	15.0	18	12	Small C to tune
4.	68	1.0	5.5	30	16	
5.	116	1.0	5.5	24	16	
6.	130	1.0	6.0	20	10	
7.	141	1.25	6.0	20	10	
8.	60	1.25	6.0	30	10	
9.	124	1.0	6.0	16	10	
10.	113	1.0	6.0	14	10	
11.	84	1.0	6.0	24	10	
12.	131	1.0	6.0	18	10	Chosen for tests
12.	145	1.0	6.0	18	10	Tested at 9MC.
13.	141	1.25	5.5	18	12	
14.	135	1.25	5.5	18	8	
15.	113	1.25	5.5	18	6	

The table gives the results of some of the more significant of these measurements obtained with each coil placed in a two inch square aluminum can of suitable length.

From these data the best practical coil was selected and tested in a tuned radio frequency stage using various coupling means. As might have been expected, discouragingly low gains resulted from the use of the conventional, tuned, inductively coupled step-up transformer in the plate circuit. This led to the coupling arrangements shown in Fig. 1 whereby a markedly higher tuned impedance was included in the plate circuit of the amplifier tube.

This circuit, including the necessary filtering, while sufficiently stable, varied widely in gain from 6 down to little more than 1. The fact that the gain was maximum when the capacity was nearly minimum substantiated the fact that, as has been generally recognized, the use of broadcast capacity range tuning condensers with the necessarily minute inductances to cover the high frequency ranges, is a serious limitation to the maintenance of the amplification at a high value throughout the frequency range.

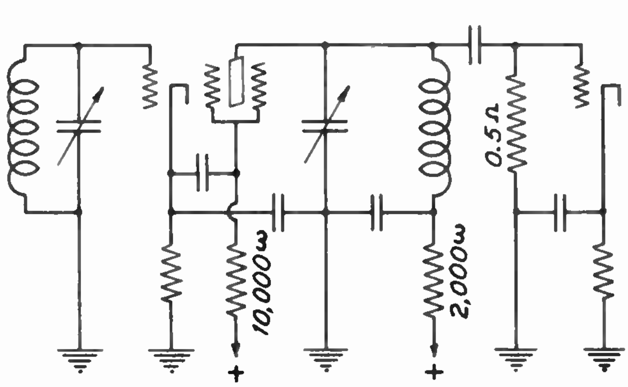


Fig. 1

With these facts in mind, after it had been determined that anything as small as a 100uuf variable condenser would not afford broadcast band coverage, serious consideration was given to the possibility of using larger inductances with the customary broadcast type ganged tuning condenser. A few tests indicated that a marked increase in inductance would be necessary to attain a reasonable gain from amplifier stages operating at these high frequencies. At the same time, the equally great need for improvement in antenna circuit gain was kept in mind.

The use of a small fixed condenser in series with the usual tuning capacitor was considered, but not tried because of the inevitable crowding at the high frequency end of the dial, the increased losses to be expected unless a fixed air condenser were to be used, and the limitation in tuning range. An inductively coupled tuning transformer, as indicated in Fig. 2, was next considered, tried, and found to function fairly satisfactorily, though with little improvement in gain in return for the increased switching complication, bulk and cost.

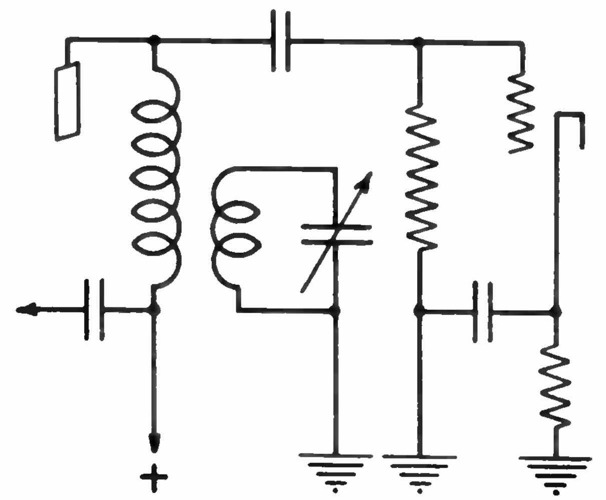


Fig. 2

Theoretical considerations indicate that, in an ideal system, connecting a given condenser, "C", across a few turns, "n", of a given coil having "N" total turns, is equivalent to putting a capacity,

$$C\frac{n^2}{N^2}$$

across the whole coil. Thus, if such a "tuning auto-transformer" consisted of a condenser of 350uu fd capacity connected across half of a coil of 2.5 micro-henries total inductance, as indicated in Fig. 3, this combination would be approximately equiva-

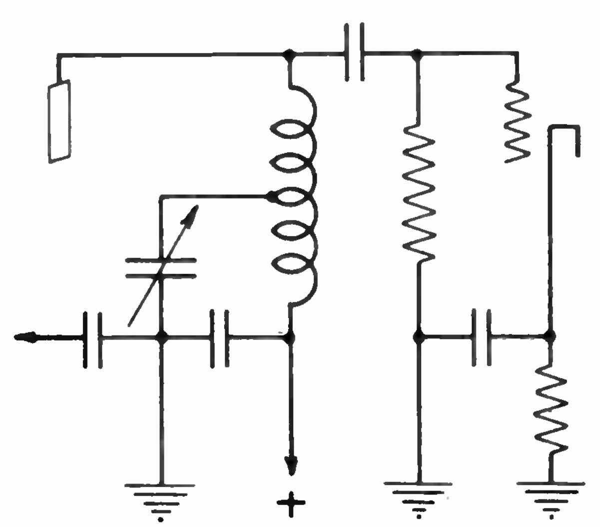


Fig. 3

lent to 350uufd connected across an entire coil of 0.65 micro-henries and would tune to about 10 megacycles in each case. First tests with this type of system used in a tuned radio frequency stage indicated that while the theoretical maximum increase in gain of four to one was not attainable, it can be approached sufficiently close to afford a marked improvement in performance.

Considerable subsequent experience with this system, used as an r-f coupling means covering a frequency range from 8 to 20 MC has shown it to be very effective. Gains as high as 35 were found possible using completely practical circuit elements with a comparative small change of gain with frequency over the band of any one coil. Variations of gain over a single band of less than two to one was easily attainable and in an exceptional case a gain of only 10 per cent was obtained. Frequency ratios of 2.25 to 2.75 are attainable in practical arrangements with tuning curves substantially the same as those resulting from the use of the same elements in conventional type of circuits.

It is to be noted, however, that the capacity resident in the circuit wiring and the dielectric losses in that capacity become of especial importance in this type of circuit and must be kept at the irreducible minimum for best results.

ANTENNA COUPLING SYSTEMS

Returning to consideration of the problem of high antenna gain, using broadcast type tuning condensers, a study was made of the possibility of using the same center-tapped tuning arrangement in order to maximize the resonant rise of the antenna voltage into the grid of the r-f tube and to eliminate tracking difficulties when used in conjunction with this type r-f coupling system. Of the many possible, two (fortunately the simplest) coupling means were found to be the best.

Fig. 4 indicates one simple antenna coupling method, in which the small series antenna condenser may have a capacity between 10 and 15uufds, while the alternative is to connect the antenna to the top of the coil through a capacity of about 3uufds. Subsequent experience has shown that while the results from the two systems are quite similar, practical switching considerations may give one some preference over the other. Using either connection, gains ranging from 3 to 6 have been experienced using a 400 ohm dummy antenna. In general, modification of existing excellent practical receiver circuits to incorporate this principle has resulted in over 50 per cent improvement in antenna circuit gain.

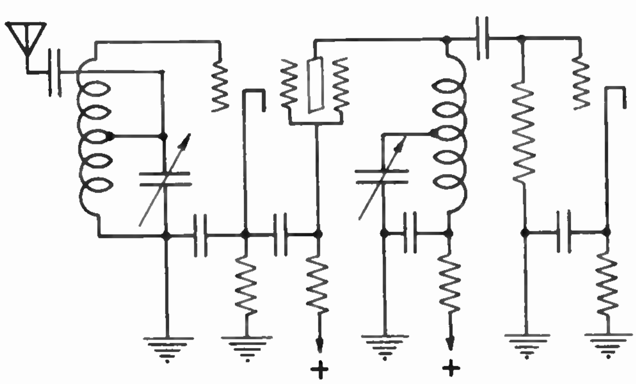


Fig. 4

In the combination of this antenna coupling system with the similar r-f system oscillation occurred on resonance in spite of all the usual filtering precautions which had been incorporated in the set-up.

This instability defied all attempts at its elimination, though thorough shielding, considerable component separation, and separate battery power supplies were tried. A study of the circuit then revealed the possibility that since, at these high frequencies, the coil and condenser leads constitute a considerable part of the total inductance of the tank circuit, the very short, low impedance rotor shaft connecting the usable sections of the ganged condenser might well provide sufficient coupling between the grid and plate circuits, as indicated in Fig. 5. to cause the instability noted. Checking this theory by using separate tuning condensers substantiated this fact.

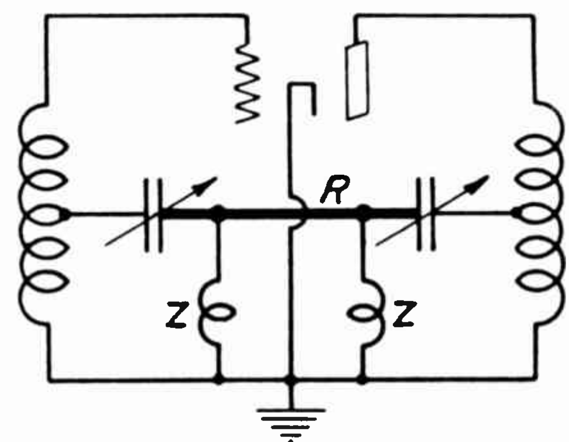


Fig. 5

The elimination of this coupling by the centralization of the several groundings in the circuit at one point in the variable condenser structure as shown in Fig. 6 resulted in complete stability at all frequencies.

In fact, it was rather astonishing to find how little coil shielding and part or circuit separation were needed to provide complete stability in such a high gain system after this cause, which has proba-

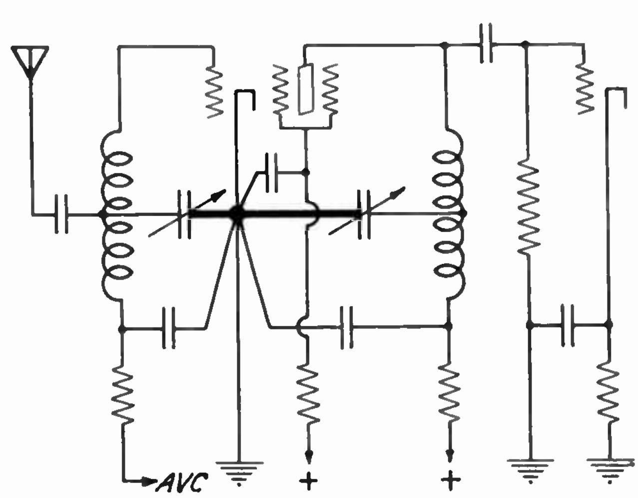


Fig. 6

bly always existed in less aggravated form, was eliminated. Marked improvement in sensitivity and signal to this ratio was then obtained, with complete stability, using the entire system as indicated in Fig. 6.

COIL SWITCHING SYSTEMS

Experience with the practical embodiment of these principles has led to simple efficient coil switching arrangements. Careful measurements have shown a negligible difference in efficiency between parallel and series coil switching systems so that different physical set-ups may give either a 5 per cent advantage over the other. For that reason, both arrangements have been used with success, choice depending on convenience. A simple series coil switching arrangement has been devised, embodying all the desirable features which have been mentioned, as indicated in Fig.7, while Fig.8 indicates an equivalent circuit utilizing the parallel coil arrangement. It will be noted that in either circuit the maximum benefit is derived from the grid and plate "boosting coils" on the highest frequency band and that as more inductance is switched into each circuit for the lower frequency bands, the effect of these coils becomes less and less until, on the broadcast band, their effect is negligible. In this manner it is possible to partially counteract other factors which lessen receiving efficiency on the higher frequencies where it is desirable to concentrate the greatest gain.

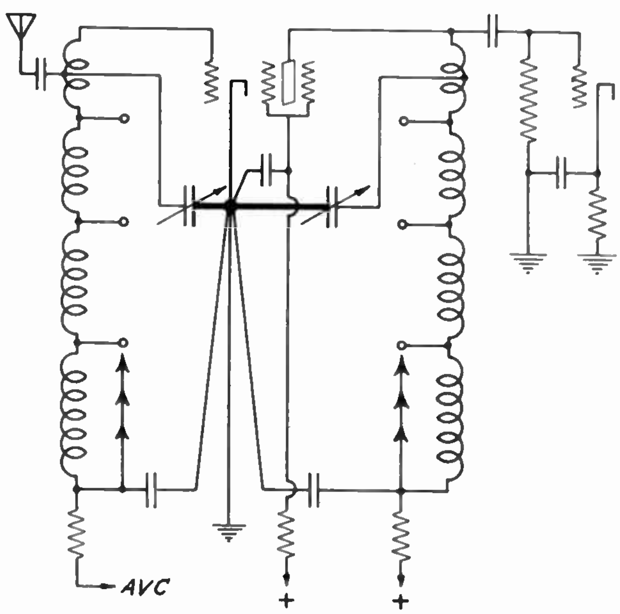


Fig. 7

In order to facilitate tracking and compensate for slight inductance variation in the manufacture of these coil systems, individual trimmers may easily be connected across each of the series coils shown in Fig. 7. Padders may be conveniently added, to either circuit, if desired, by providing an extra set of contacts on the coil switch. However, padders have not been found generally necessary or desirable. Due to the use of a small condenser in series with the antenna lead, wide changes of antenna characteristics have a negligible effect on the tracking of the antenna and r-f circuits.

In general it may be said that the circuit arrangements, the development of which has been here described, provide highly effective means for securing high radio frequency stage gain as well as antenna circuit gain and thus provide performance characteristics for short wave receivers comparable with those commonly accepted in broadcast receiver practise. No especial or "tricky" expedients need be resorted to to attain amplifier stability and the resultant mechanical and electrical structure need include nothing but the simplest of coil and switching gear construction.

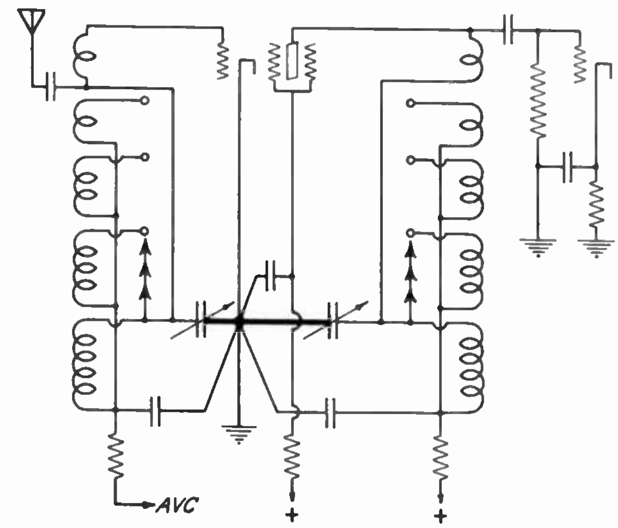


Fig. 8

ACOUSTIC FEED BACK PROBLEMS

The broad subject of regenerative electrical feed back in radio equipment has been of interest and importance ever since the introduction of the vacuum tube amplifier and detector provided such degrees of amplification as made possible sustained oscillation through the ever present feed back couplings inherent in the equipment. The possibilities of the utilization or elimination of this phenomena were early recognized and the means for its control or neutralization thoroughly investigated and established.

On the introduction of the loud speaking type of reproducer into the radio system an extension of the feed back principle including, in addition to the electrical couplings, a mechanical or acoustic coupling resulting from the mounting of the loud speaker in or near the mechanical structure of the amplifier system gave rise to acoustic regeneration resulting in what is termed "acoustic howling" and requiring new analysis of the problem and new methods and expedients for its elimination. While the presence or absence of purely electrical feed back coupling may markedly influence the over-all stability of an amplifying system including an acoustic feed back element, the methods for the analysis and elimination of the purely electrical feed back are so well established and generally understood that no further reference to them need be made in this discussion.

The development in the last few years of extremely high gain receivers in ever decreasing physical size and the therefore unavoidable increase in acoustic couplings between the several parts of the system has made essential the careful analysis of the effect of such couplings on the over-all stability and the devising of means for its reduction or elimination. It is, therefore, the purpose of this portion of the paper to review work done in this field and to suggest methods for the experimental analysis of this phenomena and to suggest such modifications in the electrical and mechanical structure of conventional forms of receiving equipment as will offset the influence of such acoustic couplings as are unavoidable.

AUDIO MICROPHONICS

Of all acoustic feed back trouble audio microphonics are, perhaps, the most widely known and understood. They result when sound energy from the loud speaker causes one or more of the audio tubes or the detector to vibrate. The condition necessary for this feed back to cause howling is fulfilled when the audio amplification between the detector and the loud speaker is equal to or exceeds the acoustic attenuation between the loud speaker and the detector. The tendency toward microphonic howling is therefore, a function of the audio gain, loud speaker efficiency, acoustic attenuation between speaker and detector and any mechanical resonances in the tube or the speaker.

While this variety of howl is no longer of great importance in practise as the result of the general use of relatively low audio gains and ruggedly constructed tubes certain familiar facts pertaintaining fo this problem are pertinent to later discussion and will, therefore, be reviewed at this point. Reference is here made to Fig. 9 which represents graphically an audio system including the loud speaker and acoustic feed back link, and in which it is assumed that the only effective acoustic link extends to the detector.

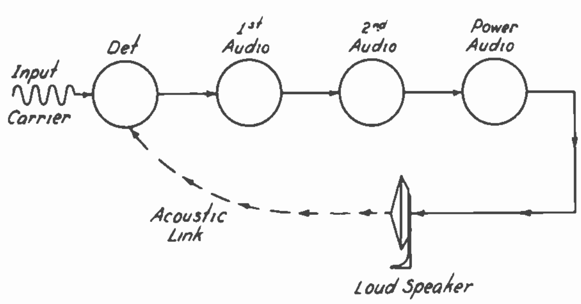


Fig. 9

In the case of the grid leak detector, the microphonic howl occurs when little or no carrier and hence little resulting biasing voltage is impressed on the grid since under those conditions is the tube most effective as an amplifier and highly effective for the modulation of its plate current by the vibration of the grid. In the case of the grid bias type of detector circuit the converse is true since the detector tube acts as an effective audio amplifier only when the applied carrier offsets the effect of the normal grid bias and thus acoustic howling results only in the presence of a relatively strong carrier.

The early investigations into this phenomena suggested simple means of analysis which are still of great utility. Briefly, after disconnecting the loud speaker from the receiver circuits, maintaining its physical connection with the amplifier, however, the speaker is excited by a conventional audio oscillator of variable and known frequency and voltage and the output signal at the receiver terminals across a load equal to the speaker impedance resulting from this mechanical excitation carefully measured over a wide range of frequency. Hence, and from the above criterion that, to avoid acoustic howls, the acoustic attenuation must exceed the audio amplification it follows that at those frequencies at which the receiver output voltage exceeds the speaker exciting voltage, the system may howl, depending upon the phase relations between the voltages involved. And the amount by which these voltages differ is a direct measure of the tendency to howl or the margin of stability against howling.

For the elimination of howling the phase of the speaker connections should be so chosen to suppress the howl corresponding to the greatest of the several possible tendencies to howl as indicated by the above mentioned voltage differences and the remaining tendencies toward howl provided against by the usual expedients of flexible detector sockets, weighting the detector tube to lower its natural period, damping the detector tube vibration with rubber or felt pads, and reducing the acoustic coupling through the further removal or acoustic insulation of the loud speaker.

It should be noted at this point that the type of howl here discussed is not to be confused with the "motor-boating" sometimes quite similar in sound which, however, results from poor regulation of the power supply unit and overloading of the tubes, commonly the power tube. In practise it will be found that this type of howl may be differentiated from the normal acoustic howl by the measurement of the tendency to howl as outlined above or, as an alternative, the replacement of the loud speaker by headphones, properly loading the output tube, of course, and observing whether or not the howl persists with the negligible acoustic coupling provided by headphones.

It should be further noted that in every case where excitation of the receiver by a carrier is resorted to, the carrier must be completely free of modulation since the presence of modulation probably through something of the nature of super-regenerative action, markedly reduces the tendency toward howling.

HIGH FREQUENCY MICROPHONICS

Like the audio frequency microphonics discussed in the previous section, radio frequency microphonics are not unique in the modern high gain receiver but their troublesome influence has become more evident with the introduction of the superheterodyne and its higher sensitivity and selectivity. Like the audio frequency microphonics they require that a carrier be present since, without it the mechanical and relatively low frequency vibration of tubes and circuit elements can provide no excitation of such a frequency as will be passed along by the R.F. and I.F. stages to the second detector tube. The consequent possible confusion between A.F. and H.F. microphonics can, therefore, be avoided only by preliminary search for the seat of the feed back in the A.F. system by its isolation from the H.F. system and only after any A.F. microphonics have been eliminated can the further investigation into the H.F. stages of the system be effectively carried on.

In general, there are two distinct types of microphonics in the high frequency stages: those arising from the vibration of circuit elements such as variable condensers, padders, coils and connecting leads, etc.; and those arising from the vibration of the tube elements. The former of these, for reasons that will be obvious from their further discussion in a later section can give rise to microphonics only when the system including them is detuned from the carrier exciting them. The former for reasons perhaps obvious from the discussion of A.F. microphonics, can give rise to microphonics only when the system is tuned to the frequency of the carrier and, indeed, are most troublesome when the receiver is precisely tuned to the carrier. Thus, there is provided a simple and easily effected means for localizing the seat of the H.F. microphonics as between the tubes and the circuit elements of the receiver, and the further analysis may then follow.

"ON RESONANCE" HOWLS

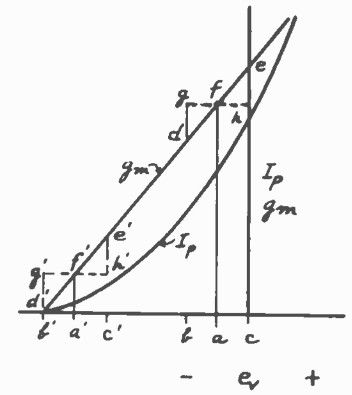
For the further consideration of H.F. microphonics arising in the tube elements more detailed consideration must be given to the phenomena resulting from the vibration of the tube elements when the tube is acting as an amplifier of high frequency voltages. For this reason it is to be noted that the amplifying properties of any tube reside in the fact that as the voltage on the grid is varied the field about it and within the tube structure varies so as to repress or encourage the flow of electrons from the cathode to the plate. If, however, the potential on the grid is maintained at a fixed value and its position relative to the cathode and plate is varied, a similar change in the field about it occurs and gives rise to a change in the amplifying properties of the tube, much as if a change of voltage on the grid had taken place. Obviously then, when the grid is caused to vibrate freely, a cyclic change in the amplification results and any carrier present on the grid of the tube appears in the plate circuit modulated at the frequency of the grid vibration.

It is this mechanically modulated carrier, when applied to the second detector that provides the audio current which makes the acoustic howl possible. And since it is the absolute amplitude of the modulation of the carrier and not merely the ratio of this amplitude to the mean carrier amplitude, i.e., the percentage modulation, that determines the audio output of the second detector, it is usually necessary that the system be precisely tuned to the carrier so that a definite threshold level of carrier at the second detector be exceeded if the acoustic coupling is to be effective in creating howling. The inclusion of the AVC in the receiver in no way modifies this requirement since first, any practical A.V.C. system is by no means completely effective in maintaining the second detector input independent of the applied carrier amplitude and secondly, the very employment of AVC circuits further augments the tendency toward howling at high carrier levels.

This will be more easily evident from Fig.10 if it is borne in mind that the mechanical vibration of the grid through a definite amplitude performs the same electrical function as the application to the grid of a definite A.C. voltage of the frequency of the vibration. In Fig. 10 is shown the characteristic curve of a conventional tube in which the plate current is proportional to the square of the grid voltage.

The amplification possible with such a tube when used in conventional circuits is proportional to rate of change of plate current with grid voltage change, i.e., the mutual conductance, G_m of the tube, which in the square law tube is directly proportional to the grid voltage as shown in the mutual-conductance curve of Fig. 10.

Fig. 10



$$I_p = Ke_v^2$$

$$\frac{dI_p}{de_v} = 2Ke_v$$

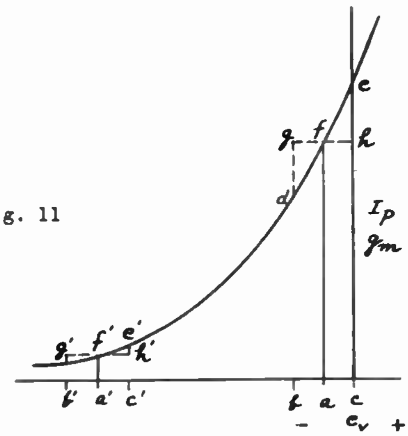
It will be evident that if the grid voltage at a definite carrier level adjusts itself to the value a and the vibration of the grid results in an equivalent grid voltage change of bc, there results an amplitude of modulation he and a percentage modulation of he divided by fa which in this case is of the order of sixteen per cent.

Where, however, a high carrier level operating through the AVC mechanism drives the grid bias to the point a' and the vibrating grid provides the same equivalent grid voltage swing as before and as indicated at b'c' there results the same amplitude of modulation as before, as indicated at e'h' but, however, with a greatly increased percentage of modulation as given as the quotient of e'h' divided by f'a' in this case about 100%. And since the AVC functions as to maintain the voltage on the second detector grid approximately of the same value and largely independent of the carrier input, the audio output of the second detector and hence the tendency to howl is much increased by the action of the AVC.

This condition does not, however, obtain where the exponential type of amplifier tube is used since it is the primary characteristic of this tube that the amplification made available through its use varies exponentially with the control grid bias and thus provides a substantially constant percentage of modulation for a given equivalent grid voltage change generated by the vibration of the grid. This will be more clearly evident from Fig. 11, which shows the characteristics of the exponential type tube, and in which the notation is substantially the same as in the previous figure. And while the commonly available remote cut off tubes are probably not strictly exponential in their characteristics, but are, for the most part, largely constituted of tube elements giving two separate square law characteristics, the lower effective amplification constant of this type of tube at highly negative grid biases provides definite usefulness in offsetting the tendency toward microphonic howls through their employment.

It is to be noted in connection with all work in the tracking down of "on resonance" howls and their relation to tube characteristics, that the investigator must assure himself that all variable tuning elements are in thorough alignment, lest the

Fig. 11



$$I_p = Ke^{ae_v}$$

$$g_m = \frac{dI_p}{de_v} = Kae^{ae_v}$$

lack of that alignment hide the presence of howling tendencies otherwise discoverable or, on the other hand, the misalignment make possible howls due to causes other than mechanical feed back to the tubes.

For the treatment of "on resonance" howls, substantially the same remedial measures recommended for audio howls are of usefulness with the addition that the employment of exponential or remote cut-off tubes in the offending stages is especially helpful.

"OFF RESONANCE" HOWLS

Where the offending element in the receiver giving rise to acoustic coupling with the loud speaker is one of the tuning elements, a distinctly different type of phenomena from that observed in the case of the speaker-to-tube coupling results. This will be evident from consideration of what occurs when, for instance, the plates of the variable condenser are caused to vibrate through their mechanical coupling to the loud speaker. If they are well centered with respect to one another and vibrate at the frequency of the emission from the loud speaker and are, at the same time so adjusted as to precisely tune the system to the frequency of the carrier being received, their tendency to modulate the carrier will be very little because of the "flatness" of the resonance curve of the circuit of which they are a part. But even though the resonance curve were extremely sharply peaked, such modulation as occurs would not be of the same frequency as that of their mechanical vibration since they would, in fact, throughout a single cycle of movement of the plates, either side of their centre and normal position, detune the system not once, but twice and thus modulate the carrier, at twice the mechanical exciting frequency, if at all. Thus no acoustic feed back can result under these conditions.

If, however, the tuning system is deliberately detuned and sufficiently high carrier levels are applied to the input, a distinctly different condition results. In this case it is evident that as, for instance, the condenser plates are vibrated and the frequency to which they tune the system varies between the upper and lower limits either side of the normal resonant period, only once per cycles of the mechanical vibration does the tuned system approach resonance with the applied carrier and thus the carrier is modulated at the frequency of the mechanical vibration and howling can occur. There is, fortunately, a definite phase relation between the mechanical excitation applied to the tuning system and the resultant modulation which immediately suggests the possibility of the repression of this type of acoustic howl through the reversal of the loud speaker connections. And, indeed, such a reversal does accomplish the expected repression only, however, to bring it into being again when the tuning system is so adjusted as to set it off of resonance on the other side of the carrier and thus really provides nothing but an effective method of recognizing the seat of the coupling. Since, for a given condition of speaker connection this type of howl occurs at only one side of resonance it has been termed, not only "off resonance" howl but "assymetric" and "adjacent channel" howl as well.

It is to be noted that howls from such causes as are here discussed cannot occur on both sides of resonance and that evidence apparently to the contrary indicating symmetrical off-resonance howls is merely indicative of the fact that at least two separate couplings are present and giving rise to the howls and that the search for their location must include more than one of the circuit tuning elements.

The influence of the value of the "Q" on this type of howl should be especially noted as suggested by Fig. 12 in which the resonance curves of circuits having Qs of the ratio of two to one are shown. Thus for a vibrational change of the resonant frequency of the circuit of a value ef or its equal, bc, will result in the case of the relatively low Q and non-selective circuit shown at S in a modulation of the carrier proportional to km while in the case of the more selective circuit shown at T it will result in a modulation indicated by gh with the consequently greater tendency to howling.

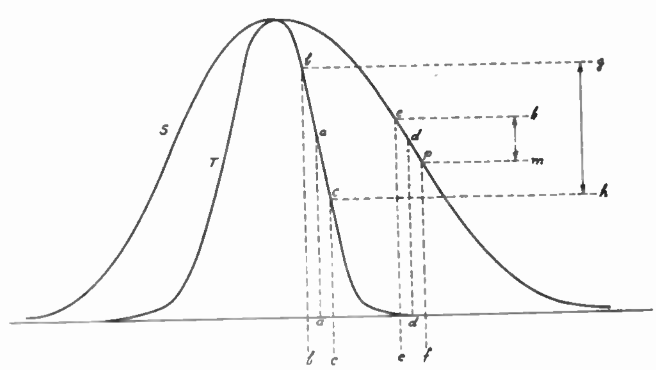


Fig. 12

It will be noted also that the amount by which the tuning system must be off resonance to offer the maximum slope of the resonance curve and hence maximum tendency to howl is also a function of the value of Q, being greater the smaller the value of Q.

It is obvious then that those factors which make for greater selectivity and image suppression as well as all of those performance requirements tending to increase the Q of the radio circuits unavoidably increase the tendency toward off-resonance howls.

Since practical experience shows that the most fertile source of mechanical modulation in the tuning systems resides in the variable condensers, it is of value to give further detailed consideration to this one of the circuit elements. It is generally appreciated that where a metal plate constituted a capacity with two other parallel adjacent and symmetrically placed plates, minor motions of the central plate in a direction normal to the surface of the plates gives little change of capacity, since the increase of capacity brought about by the approach of the moving plate to one of the stationary plates is largely offset by the reduction of the capacity with respect to the other of the stationary plates. And, so in the case of a variable condenser made up of a series of carefully aligned plates, little difficulty resulting from carrier modulation by the axial motion of the plates relative to one another need be anticipated except, perhaps, in the case of the end plate for there must, in any conventional structure, be plates at the end of the structure that are not symmetrically located with respect to two others. To meet this latter condition, special proportions in the end plate spacing as referred to below may be employed but, insofar as mechanical modulation due to uncentralized intermediate plates is concerned, there is only one rational solution and that lies in careful alignment since the art of condenser production has so far advanced that there is left no excuse whatsoever for this type of irregularity.

As for the influence of the end plates there is, of course, no condition under which such an assymetrically located plate will contribute more to the howling tendency than any of the intermediate plates. Careful theoretical investigation indicates, however, that if the end plate is removed from its neighboring plate by a distance approximately forty per cent greater than the normal plate spacing the tendency to howl is minimized.

Aside from these detailed specific points there are several generalizations worthy of formulation and note. Amongst these is the fact that in general, the more compact the variable condenser, the greater is the tendency toward the generation of howl. There is in general no single setting of the variable condenser that will always accompany the greatest tendency to howl; if the tendency to howl is completely resident in the variable condenser that position which provides the highest selectivity will most easily provide a howl: if, on the other hand, some element providing a large portion of the residual capacity, such as padders, etc., is the vibrating member, the greatest tendency to howl will occur with the condenser plates unmeshed: while if the vibrating member is a portion of the inductive element the howl will occur most readily at that setting of the condenser which provides the highest selectivity and since this latter condition is usually realized at the low frequency end of the range it may be said that, in general rotor plate and inductive microphonics occur at the low frequency end while microphonics due to trimmers, padders, connecting leads, etc., occur at the high frequency end of the range; and that a complete check for microphonics can be made only by a careful investigation throughout the range.

There is another interesting point to be noted in this connection and that is summarized by pointing out that, unlike the case of A.F. howling, any stage of the high frequency system is equally susceptible to howl excitation since the phenomena is one of modulation of the already present and successively amplified carrier.

Further, there are several interesting points with respect to the choice of plate thickness that merit consideration.

In general, condenser plates should be thick rather than thin. The stiffness increases approximately as the cube of the thickness and the mass only as the first power. This raises the natural period of the plate and with given energy, the amplitude of vibration decreases directly with frequency. Since the amplitude of condenser vibration controls the per cent of microphonic modulation, the tendency toward acoustic howl decreases as the natural mechanical resonance is increased. Condensers with plates whose mechanical resonance is above 1200 cycles have seldom been found to give rise to acoustic howl.

Such specific expedients for the elimination of microphonics originating in the variable condenser as have, in practise been found useful are the mechanical insulation of the condenser through rubber or felt mountings: the acoustical isolation of the condenser from the speaker through its further removal from the speaker and the elimination of such coupling elements as large dials, knobs, etc.; increased spacing of condenser plates, particularly the end plates; the use of soft metals such as aluminum for the condenser plates and the damping of especially troublesome individual stator or rotor plates with rubber or felt where necessary: the weighting of the entire condenser assembly where evidence of its vibration as a whole is shown as the cause of the howl.

INTERMEDIATE FREQUENCY MICROPHONICS

While the fundamental causes of microphonic howls originating in the I.F. system of the receiver are identical with those inherent in the R.F. system as discussed above, the somewhat different mechanical and electrical properties of the I.F. system provide certain important differences in the tendency to howling. Primarily this is due to the fact that the Q of the I.F. tuned systems is, in general, likely to be markedly higher than that of the R.F. tuned systems and hence greater care to guard against acoustic couplings is necessary. Furthermore the higher value of Q results in a smaller difference between the point of howl and the resonant frequency and thus provides a clue for differentiating between off-resonance howls originating in the R.F. and the I.F. systems. And since, in general the capacity load on the I.F. transformers is less than that in the R.F. stages, the effect of the vibration of connecting wires, etc., is more important and justifies closer scrutiny of these elements in running down the source of acoustic coupling. Loose compression type condensers of the type commonly used in the I.F. stages are a frequent source of acoustic coupling since this type is often set in a wide-open position: and where air condensers are used for this purpose they are more likely to lack the nice plate alignment commonly found in the variable air condenser.

But aside from these specific mechanical conditions, for the elimination of which the same expedients as were recommended in the previous section will serve, there is nothing essentially different to be looked for in the I.F. stages.

OSCILLATOR MICROPHONICS

Insofar as the oscillator circuits are concerned, the modulating influence of the vibration of the oscillator tube will be found to be identical with that of the vibration of any other of the high frequency tubes in the system and will manifest itself by the usual on-resonance howl, and in this the oscillator tube is no more susceptible than any other tube in the receiver. However, this effect is negligible as compared with that resulting from the frequency shift due to the vibration of the oscillator tube which gives rise to assymmetric microphonics.

Assymmetric howls caused by oscillator vibration possess all the characteristics of i-f microphonics except that they occur on weaker signals, or reduced amplification, and becomes more prominent as the oscillator frequency is increased. Thus oscillator microphonics are factors of major importance at short waves.

The greater tendency toward howling due to oscillator vibration will be obvious from consideration of two pertinent factors. First, since the signal delivered to the I.F. amplifier is of a frequency which is the difference between the oscillator frequency and the signal carrier frequency, the percentage change of this beat frequency will be greater than the percentage change in the frequency of the oscillator itself due to vibration by the ratio between the oscillator and I.F. frequencies. And secondly, since the selectivity of the entire I.F. system is operative against the vibrationally varied beat frequency, the I.F. system acts much as if the beat frequency were maintained constant and the resonant frequency of each of the tuned circuits constituting the I.F. system simultaneously and widely varied. Such a combination of conditions obviously provides most effectively for the modulation of any carrier applied to the system by any mechanical vibration which is allowed to effect the oscillator tube or any portion of the oscillator system. Especially severe is this condition in the high frequency ranges where the ratio of the oscillator frequency to the I.F. may be of the order of forty to one and the tendency toward microphonics proportionately great.

Against these unusually severe conditions little more by way of recommendations for their elimination can be made than were made in the preceeding sections. It is usually well worthwhile, however, to provide for the mounting of the oscillator coil not less than one quarter inch from any metal and, where space is available, at even a greater distance. The oscillator section of the variable condenser should be at one end rather than at an intermediate point in the assembly. And, no attempt should be made to align the capacity of the oscillator section with the capacities of the R.F. sections by bending any of the oscillator section plates, especially does this apply to the end plate. If equalization must be accomplished by this crude means, the necessary adjustments should be made on the other sections of the condenser using the oscillator section as the standard.

The author is pleased to acknowledge that the major portion of the section of this paper devoted to the subject of microphonics is the result of an intensive study of the subject made by Mr. David Grimes who is to be commended for his orderly analysis of the several phases of the problem and the compilation of the mass of useful data gathered in his investigation of it.

NOVEMBER MEETING

The November meeting held on November 15th, was devoted to a joint paper by Mr. A.B. Chamberlain and Mr. W.B. Lodge of the Columbia Broadcasting System, on "Broadcast Antenna Systems." In this paper was disclosed a host of data gathered in the investigation of the operating properties of a wide range of antennas as now used in the Columbia System. In these is included various types of vertical and flat-top wire antennas, mast antennas, and combinations of antennas for securing unusual and especially desirable field patterns. This paper will be published in the November issue of the PROCEEDINGS.

DECEMBER MEETING

Mr. Joseph Chambers of WLW will describe some of the especial features of the antenna systems and transmission lines of the 500-kilowatt installation in Cincinnati. In view of the especial characteristics made necessary by the high power employed at WLW, it is anticipated that this paper will be especially interesting.

ROCHESTER JOINT MEETING

The Radio Club of America held a joint meeting with the Institute of Radio Engineers on the evening of November 12, 1934, at Rochester, New York, in connection with the now traditional Fall Meeting of the Institute. Vice President R.H. Langley presided, and called attention to the fact that the Club is three years older than the Institute and that this year marks its twenty-fifth anniversary.

The paper of the evening, by I. Wolff, E.G. Linder and R.A. Braden, all of R.C.A. Victor Company, was presented in a most interesting way by Dr. Wolff. It dealt with the transmission and reception of centimeter waves and included a most informing and convincing demonstration. The structure of the special oscillator tube, designed for a wavelength of 9 centimeters, and including a circuit resonant at that wavelength as part of the structure of the tube, was explained in detail with the aid of slides that showed the construction. The tube is based on the basic fact, which seems to underlie chemistry and magnetism as well as electrical science, that an electron in motion is deflected by a magnetic field.

Although this same tube can be arranged for use as a detector of the centimeter waves, the receiver used in the demonstration employed a crystal detector with an audio amplifier and loud speaker. The transmitting tube with its radiating circuit is mounted in a 6 foot parabolic reflector, and a similar pick-up reflector has the receiving resonant circuit and crystal detector. Modulation from a phonograph pickup was obtained first by changing the plate voltage in the transmitting tube, and later by an artificial Heavyside Layer consisting of large argon and mercury tubes whose ionization was varied in accordance with the music. The unmodulated beam from the transmitter was modulated by being sent through this region of ionized gas, and also by refraction and reflection from the gas. The beam was also reflected from small metallic sheets and from the rear wall of the lecture room. Fading effects were accurately reproduced by moving a reflecting surface or the artificial Heavyside Layer. It was obvious from the applause that the paper had been thoroughly enjoyed by Club and Institute members alike.

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11 West 42nd Street + + New York City

The Radio Club of America, Inc.

11 West 42nd Street - New York City

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PROCEEDINGS of the RADIO CLUB OF AMERICA

Volume 11 November, 1934 No. 6

THE BROADCAST ANTENNA

BY

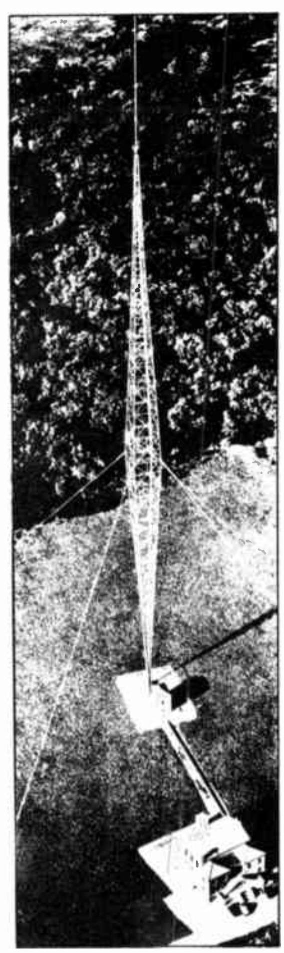
A. B. CHAMBERLAIN and WILLIAM B. LODGE*

Delivered before the Radio Club of America
November 15, 1934

The design of a broadcast antenna is governed largely by commercial considerations. A broadcast transmitter whose income is derived from advertising sales must usually concentrate on its primary service area. This may be defined as the area in which any home having an average receiver, and desiring to listen to the program transmitted can receive a satisfactory signal, free of noise and fading. Uncertain, irregular reception at long distance has some prestige value, but it seldom contributes to the income of the station, since, in general, the broadcast receiver is now used to bring entertainment into the home with little interest being shown in the fading noisier signal of distant transmitters.

In broadcasting, the power output of the transmitter is limited, both by economic and regulatory factors. The problem, then, is to concentrate all of the available radio frequency power on the local service area. Only a small fraction of this power is now directed at the population it is desired to serve.

Increasing the signal intensity in a horizontal direction is only part of the problem in the case of higher powered transmitters, nighttime fading starts many miles before the field intensity becomes too weak for a satisfactory signal. Radiation above the horizon is completely lost during the day-



light hours, this power being robbed from the useful signal. At nighttime, however, conditions are changed so that some of the radiation at angles between 40° and 70° above the horizon is returned to earth 40 to 150 miles from the transmitter. This reflected signal is of negative value since its variation in phase and amplitude causes it to interfere with the signal arriving directly from the transmitting antenna. At night the local service area ends at a point where the reflected signal approaches the directly transmitted signal in intensity.

It is fair to say that many experienced antenna engineers doubt the probability of redirecting this high angle radiation and concentrating it along the horizon. Such a development in broadcast antennas would increase the commercial value of the radio station far more than a simple increase in transmitted power. We believe the limit has by no means been reached in the improvement of transmitting antennas.

Before reviewing what has already been accomplished in this line, let us see what happens to a watt of carrier power leaving the last amplifier in a broadcast transmitter. The output of a radio station is usually fed to the antenna over a transmission line. The r-f power is then dissipated as follows:

* Columbia Broadcasting System

1. Radiation
   - (a) Along the horizon to potential receiving sites. Usually the zone between 0 and 5 degrees above the horizon includes all such sites.
   - (b) Towards the ground; partially reflected back at angles above the horizon and partially dissipated in the earth.
   - (c) Into space; either lost or reflected back to potential receiver sites: causes fading; gives long distance reception.
2. Transmission line loss
3. Coupling equipment loss
   - (a) Resistance in units.
   - (b) Transfer of r-f energy to non-radiating surfaces.
4. Resistive loss of ground system
5. Dielectric loss at base of antenna
6. Resistive loss of actual antenna
7. Power picked up by nearby metallic objects (towers, guys, power wires, building framework, etc.) and either dissipated or reradiated.

There are other losses which absorb smaller percentages of the radio frequency power. However, of the divisions of power listed above, only the first one - radiation along the horizon - is of value to the broadcaster.

There is little accurate quantitative data available as to the strength of the signal radiated above the horizon by different types of antennas.[1,2] In the case of the vertical radiator there have been extensive mathematical discussions giving the signal distribution in a vertical plane.[3,4,5,6] Attempts have been made to fix the exact height of a vertical radiator to minimize sky-wave signal at points which would otherwise remain outside of the primary service area of the station. To make actual check of this theory in practice requires airplane measurements above the antenna or an extensive study of the reflected sky wave at various distances from the transmitter.

In tne case of the lower powered stations on shared basis', the fading is not usually a consideration of importance. In these stations, therefore, the first consideration is to obtain the maximum signal at one mile. In the case of the high power, clear channel stations, however, the usable signal extends out into the zone which is affected by fading. At these stations it is important to consider both efficiency and fading characteristics of the antenna.

CLARIFICATION OF ANTENNA TERMS

Before going further, it would be well to clarify certain terms commonly used in connection with certain antennas. The terms to be clarified are:

1. "Antenna Efficiency".
2. "Antenna Length".

One of the most commonly used terms in connection with antennas is "radiation efficiency" or "antenna efficiency". And there are almost as many definitions for it as there are engineers in the field. Theoretically, this efficiency should represent the ratio of the radiated power to the input power of the antenna. However, it would be necessary to integrate the power streaming away from the antenna toward every point in space to determine such a figure. As this is not a practical procedure, measurements of field strength are usually made at convenient points on the ground near the antenna. Among American engineers the field intensity at one mile is the value usually determined, but in the formulation of a figure of merit many different arbitrary standards have been used as a basis of comparison. Thus, 100. 123, 187, 194 and 265 millivolts per meter at one mile for 1 kw have at various times all been called "100% efficiency".

In addition, since the field strength of any station is proportional to the square of the power, some engineers square the "field strength efficiency" and obtain "power efficiency". Actually, whatever the means used to express it, "antenna efficiency" can only tell the engineer the signal in millivolts per meter at one mile at the earth's surface for a certain antenna input. Efficiency ratings at present are very ambiguous, and since there appears to be no one fundamental value upon which to base efficiency ratings, it is hoped that the method of rating antennas simply in terms of signal output will be adopted generally by engineers.

The Engineering Department of the Federal Communications Commission has arbitrarily defined the definition of "antenna efficiency" as follows:[7] "The antenna efficiency equals 100% if the effective field intensity of the station at one mile, per 1 kw antenna input power, is equal to 265 millivolts per meter".

$$A_{eff} = \frac{F^2 \times 100}{265^2 \times P}$$

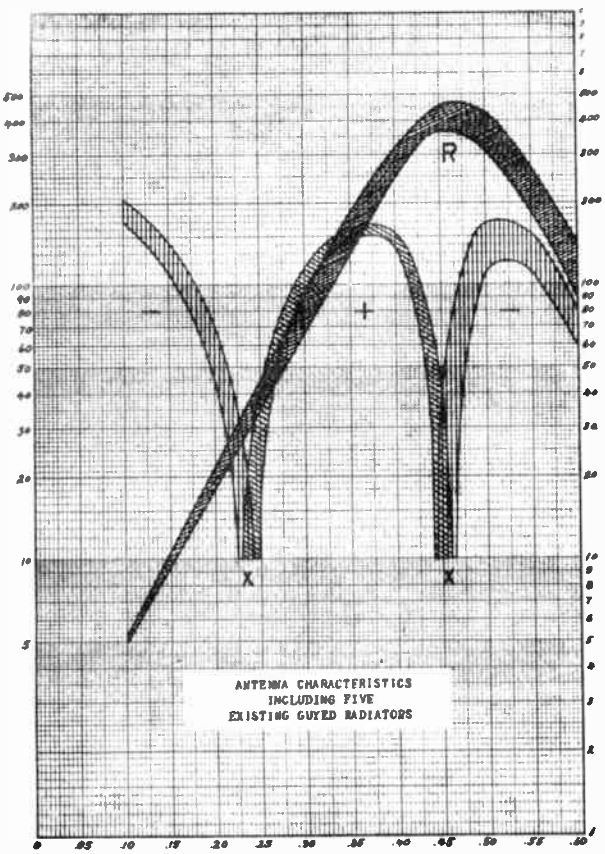
F = Effective field intensity at 1 mile.
P = Antenna input power (in kw)

The root mean square value of all field intensities at one mile from the antenna in the horizontal plane without attenuation is termed the "effective field intensity" (F).

Early antennas were described in terms of their fundamental wavelength. This was the longest wavelength at which the antenna had zero reactance and the mode of operation of the antenna was given by the ratio of operating wavelength to fundamental wavelength.

Theoretically, a vertical wire antenna will have zero reactance whenever it is any number of quarter wavelengths long; that is it will then have zero reactance when its length is one quarter, one half, three quarters, etc., of a wavelength. Practical measurements indicate however, that a vertical wire antenna for zero reactance is about 4% shorter than the above values - that is, the velocity of propogation along the wire is apparently .96 of the free space velocity.

Referring to the impedance characteristics of the guyed mast, Figure 1, a "half wave" antenna of this type is physically only .45 of a wavelength high. Following the above method of argument, it would be said that the propogation velocity of this type of antenna is only 90% of the theoretical velocity. Further, if reference is made to Figure 2, which gives the impedance characteristics of the self-supporting vertical radiator, it could be argued that the velocity of propogation is 66% of the theoretical velocity. In fact, a well-known radio laboratory in this country has released data to that effect, and has stated that a vertical tower, one half wavelength in physical height, is actually three quarters of a wavelength long electrically.



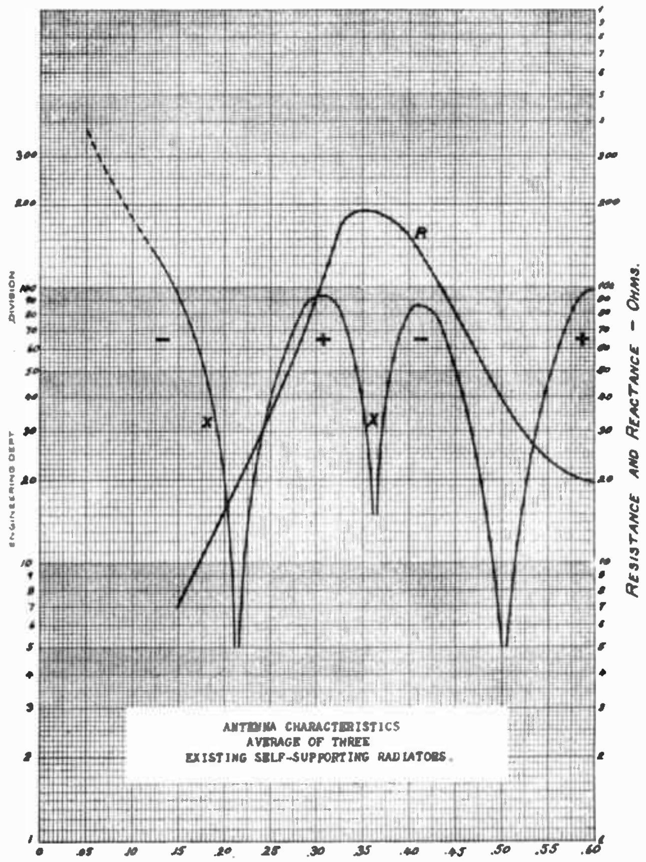
PHYSICAL HEIGHT - FRACTION OF A WAVELENGTH
Figure 1

This method of reasoning is based upon the assumption of a sinusidal distribution of current on the antenna. This assumption is not justified in practice.[8] It also assumes there is no lumped capacitance in the antenna itself. When the equivalent circuit of the antenna is considered, it may be seen that this circuit can be tuned to zero reactance at almost any frequency by the choice of a proper shunt condenser. This is exactly what occurs in the case of the self-supporting tower.

If a four to five hundred foot guyed mast and a similar self-supporting tower of a conventional design were measured, it is estimated that the capacitance to ground of the first thirty or forty feet of the wide base tower would exceed that of the other structure by at least 800 mmf. (including insulators). Taking a .47 guyed mast, operating at 1080 kc, as an example, Figure 1 shows its impedance to be 400-j60. If an 800 mmf. condenser were shunted across the base insulator, the measured impedance at 1080 kc would be.

$$z = \frac{-j178\ (400 - j60)}{(400 - j60) - j178} = 63 - j133$$

The impedance of this mast was thus made similar to that of a wide base self-supporting tower which is given in Figure 2. Yet no one would argue that the velocity of propogation on the mast itself had changed.



PHYSICAL HEIGHT - FRACTION OF A WAVELENGTH
Figure 2

It is our belief in connection with vertical antennas involving structures whose entire length is not of uniform cross section that:

1. The terms "electrical length" and "velocity of propogation" have no significant value.
2. Engineers working with these antennas should standardize on physical height (in fractions of a free space wavelength) as a method of describing the antenna dimensions.

EVOLUTION OF BROADCAST ANTENNAE

The early broadcast antenna consisted of a pair of steel, or wooden masts supporting an antenna structure usually consisting of a vertical wire, or a vertical wire and horizontal section consisting of a flat top, or cage. This latter type was known as the "T" or "L" type antenna. It is interesting to note that more than 70% of the broadcast stations in operation today employ this older type of antenna Most of these have a natural wavelength less than 1/4 of the operating wavelength. Later, it was realized that some gain would be made if larger antennas were employed. Therefore, the same type of antenna was used, but the natural wavelength was increased to 1/3 or 3/8 wave.[9] In 1924, some attention was given to this problem by Dr. Stuart Ballantine [10,11] which later resulted in the so-called .58

wave guyed vertical mast antenna. In 1931 two such antennas, designed and fabricated by the Blaw-Knox Company[12], were erected at CBS stations WNAC-WAAB, Boston; and WABC, New York. The electrical gains expected from these antennas have been realized in practice, and during the past few years a great deal of study has been given the subject of obtaining a clear picture of the electrical properties of this type of radiator. During the past two years attention has also been given to the self supporting type which has made an appearance in the field. It can be shown that the guyed type of vertical mast antenna, or the self supported type of single mast antenna, is superior to the older conventional antennas, electrically, physically and economically.

A study of Figure 3 indicates the evolution of broadcast antenna efficiencies, and shows, conclusively, that the higher mast type antenna is much better than the older conventional types from an electrical viewpoint. Early broadcast antennas resulted in an effective field intensity of as little as 100 millivolts per meter at one mile, per 1 kw antenna input power. The average of fourteen conventional type antennas, recently measured, shows that with 1 kw antenna input power, the effective field intensity is 169 millivolts per meter at one mile. The average of five self supported type mast antennas, from .20 to .35 wavelengths high, shows an average field intensity of 204 millivolts per meter, and measurements made at eight .58 wavelength antennas shows the average field intensity to be 247 millivolts per meter at one mile. There is one guyed type vertical mast antenna which has an effective field intensity as high as 280 millivolts per meter. Figure 4 indicates this same story in terms of comparative increase in power. It should be emphasized that these results are based on actual measurements and show the higher antennas to be far more efficient electrically than the older conventional types.

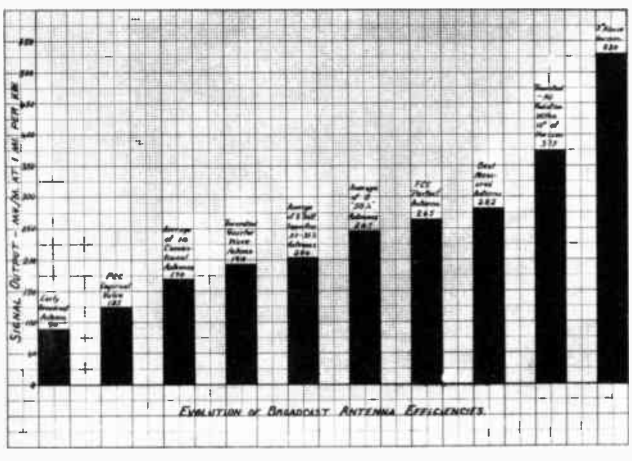


Figure 3

Either the self-supporting or the guyed radiator can be erected at less cost than a two tower conventional antenna of the same height. It has far greater reliability of operation, has a lower maintenance cost, and the horizontal polar diagram, or distribution of power, is not influenced by nearby towers, which tend not only to distort the field pattern, but also to lower the antenna efficiency. The self-supporting antenna has a very practical application in the cases of broadcast transmitters which are located in tall buildings. It is sometimes difficult to erect the older type of antenna system because of physical limitations. However, if an adequate ground is provided a moderately efficient antenna system can be erected atop a tall building through the use of a vertical mast antenna.

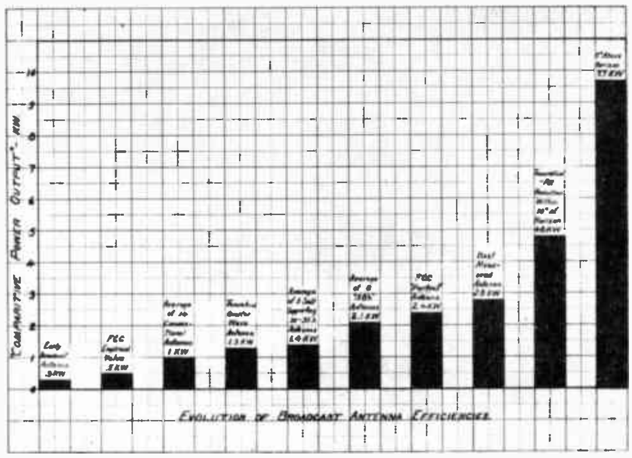


Figure 4

During the past, broadcast engineers have been very hesitant in adopting the vertical mast antenna, particularly the wide base, self-supporting type, because of a "bugaboo" which has existed with regard to high base capacitance being directly related to loss in efficiency. Station engineers have also been given to understand that it is next to impossible to couple their transmitter into such a structure because of the high reactance between it and ground.

Within the past three years a number of self-supporting towers have been erected, usually with strict limitations on insulator capacitance and tower capacitance to ground. The fear of base capacitance has continued up to the present and measurements of the signal output of these antennas operated at higher frequencies led to the conclusion that they would be unsatisfactory for use in heights greater than .35 or .40 of a wavelength. In 1934, Mr. John Byrne, working as a consultant for the International-Stacey Corporation, made measurements which showed that the loss of efficiency was due to dielectric losses in the earth near the tower. With this fact established, it was simply a matter of reducing the r-f voltage gradient in the soil at the tower base.

Self-suporting vertical antennas have appeared having the first 80 or 90 feet constructed of wood, and the radiating section above insulated from the wood. Various other modifications of the self-supporting structure were developed, including all-steel designs which insert the base insulators 20 to 30 feet above ground. Another satisfactory solution has been to construct a well grounded copper mat or ground screen beneath the tower, thus greatly reducing the high voltage normally impressed across the earth at the tower base. At station WDOD, Chattanooga, Tenn., measurements were made recently which substantiate this point. The station uses a self-supporting tower .42 wavelengths high, and has installed two sets of insulators, one just above ground, and the other set approximately 20 feet above ground. It has been found, by making antenna resistance and impedance measurements, and also field intensity measurements at one mile, that the same antenna efficiency can be obtained using either the higher set of insulators or the lower set of insulators. However, if the lower set of insulators is used, it becomes necessary to install a ground screen to reduce the dielectric losses at the base of the antenna.

From present data on broadcast antennas, we feel we can safely say that the same antenna efficiency can be obtained with nearly any type of vertical mast antenna of a given height, providing the necessary precautions are taken in design and erection, and providing the proper ground system is used.

In the design of a ground system, it is necessary that the voltage between the base of the antenna and the ground be accurately estimated in order to reduce the losses previously discussed. This information is also necessary in order that adequate insulation may be provided. If impedance measurements have been made on antennas of the same general shape as the one contemplated, the voltage may be determined quite accurately in advance. Contrary to popular belief, the highest voltages encountered do not always occur in high resistance antennas. For instance, a vertical radiator, now in the course of construction which will be .47 wavelengths high, will have a resistance of 400 ohms at the operating frequency. Its unmodulated 50 kw carrier voltage, at the base of the antenna, will be 4500 volts r.m.s., ±10%. There is another 50 kw station in service, whose antenna resistance is 15 ohms. The impedance of this particular antenna is 140 ohms, so the base voltage is approximately 8000 volts r.m.s. Figures 1 and 2 show the average resistance and impedance characteristics of self-supporting and guyed vertical radiators, and may be used in estimating base losses, insulation requirements, antenna current, antenna loading and lighting circuit requirements.

THE GROUND SYSTEM

The proper design of the ground system for use with broadcast antennas has always been important, but the design of a ground system for use with high antennas becomes particularly important, because factors other than ground loss resistance must be taken into account. The radial ground system used with a high antenna should be large in diameter, having a radius equal to at least 1/2 wavelength. It is also important that a large amount of copper be used. Figure 5 shows the empirical relationship between antenna efficiency and ground radius, based on measurements of 35 stations. The reasons for a large ground system are to reduce to an absolute minimun, ground resistance loss, dielectric losses, and the absorption of radiation directed towards the ground. Figure 5 is based on data obtained from various sources, but it is considered to be indicative of the advantages of increasing the ground radius.

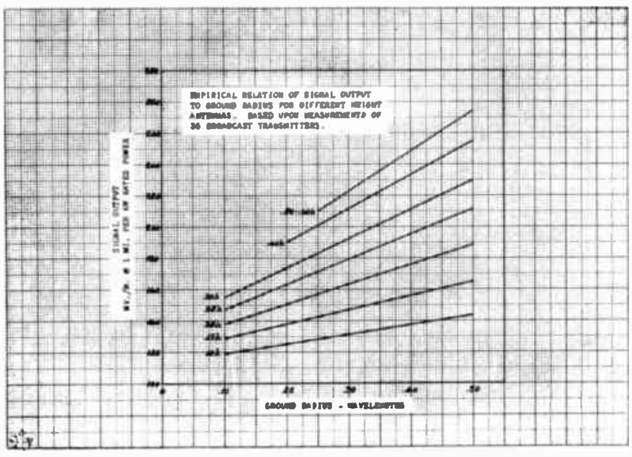


Figure 5

This paper is primarily concerned with the transmitting antenna and its associated ground system. However, the choice of a station location is definitely related to the antenna design. A site surrounded by soil of the highest possible conductivity is advisable. Actually, the final choice of a station location is governed by considerations of population distribution and coverage, so the engineer must often design an efficient radiating system above soil of poor conductivity.

If the transmitting antenna is located over poor soil, the ground current will tend to avoid the high resistance earth path and will remain on the lower resistance copper wires of the ground system. This is shown graphically in Figure 6, the current on a single ground radial at two 50 kw transmitters. The antenna and ground systems of the two stations are practically identical, but WSM at Nashville was fortunate in obtaining a station location where the soil conductivity is considerably higher than at any which were avilable to WABC.

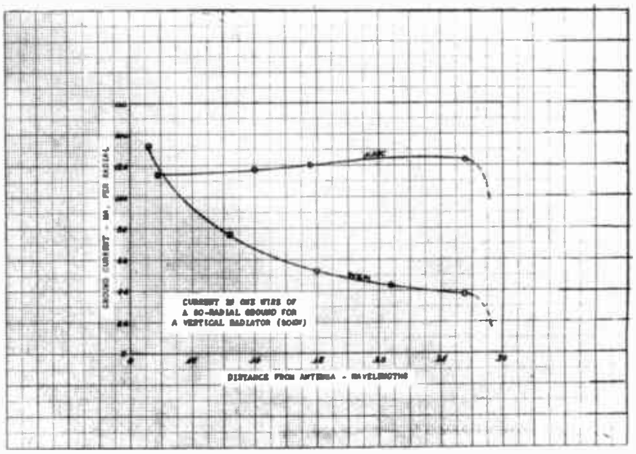


Figure 6

The effect of soil conditions also is evident in the vertical pattern of the higher types of antennas. Figure 7 shows the calculated and measured vertical polar diagram for a guyed mast antenna located in a part of Pennsylvania of comparatively low conductivity. Several studies have shown that the current distribution on an actual antenna structure departs from the assumptions which give the desirable calculated pattern of Figure 7. Earth of high conductivity is also assumed in the calculations. Skywave and fading measurements indicate that two other guyed mast antennas[14], which are surrounded by

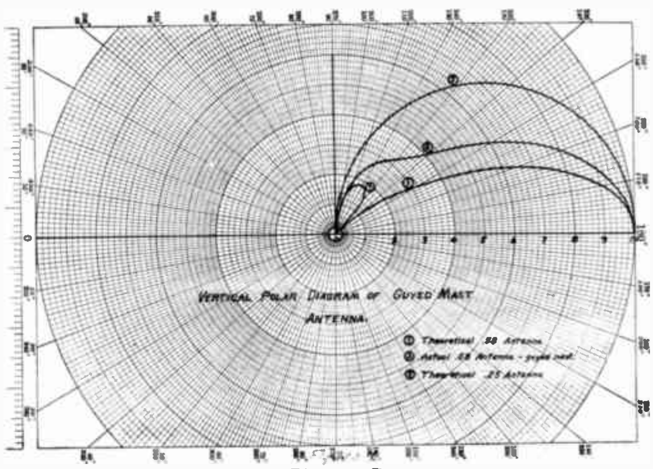


Figure 7

high conductivity soil more nearly approach the theoretical vertical pattern than similar antennas with less favorable ground conditions.

An extensive ground system increases the efficiency of any antenna, but is absolutely essential at stations which are forced to locate in areas of low conductivity.

DIRECTIONAL ANTENNAS

Directional antenna systems have been installed at stations, WJSV, Alexandria, Virginia; and at WKRC, Cincinnati, Ohio. These antennas are in operation at the present time. The purpose of employing such systems is to fulfill specific interference-reduction requirements consistent with the rendering of maximum public service.

The WJSV antenna system consists of two vertical conductors suspended between two 150 foot steel towers, insulated at their bases. The antennas are 3/8 wave apart ($\theta = 135°$) and the current in the West antenna leads the current in the East antenna by 1/8 wave ($\theta = 45°$). The 10 kw WJSV transmitter is located about four hundred feet from the antenna system and power is transmitted from it to the antenna by a conventional 600-ohm two-conductor open-wire line. Proper phasing is obtained by using transmission lines to each element of such length as to obtain the desired phase difference. The field intensity distribution in a horizontal plane is a flattened cardioid, with the minimum signal in an easterly direction. The horizontal space pattern - at one mile - is: E = 10 mv/m: N = 580 mv/m: W = 500 mv/m: S = 550 mv/m: with 10 kw antenna input. Figure 8 shows the horizontal polar diagrams of the

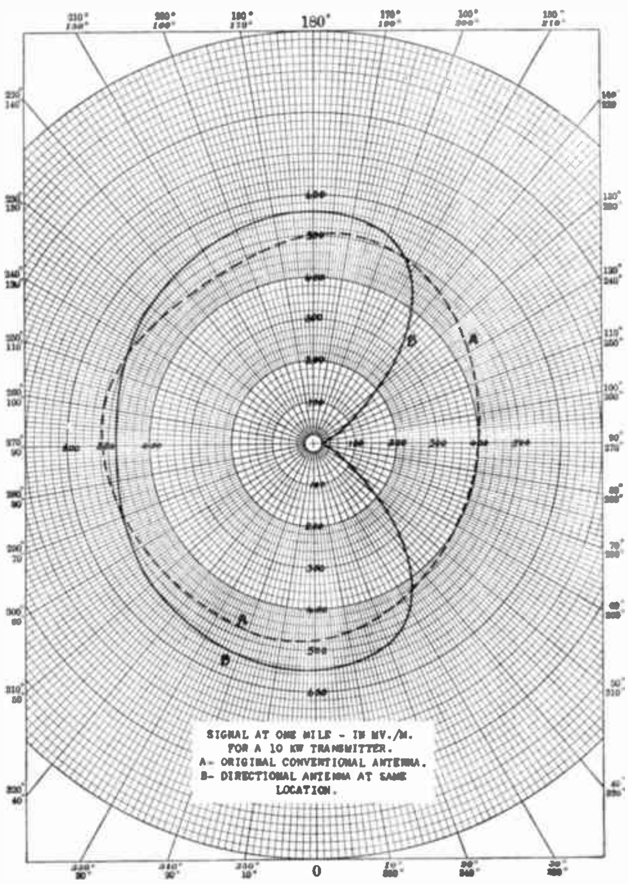


Figure 8

WJSV conventional antenna versus the directional antenna which is now in use. Optimum reduction of signal was desired, in this case, at a point one mile east of the station. It can be seen that a reduction in signal intensity of approximately 50:1 has been obtained. In order to determine the stability of the system, an accurate automatic signal intensity recorder was installed one mile east of the WJSV antenna system which records the signal strength of this station continuously. This equipment has been in operation twenty-four hours a day since July, 1933. The automatic field intensity receiver is a.-c. operated and is specially designed so that its sensitivity is independent of variations in ambient temperature, humidity and line voltage.

The records obtained during the past year indicate that the stability of the WJSV antenna system is entirely satisfactory. Inasmuch as the cardioid pattern is the most difficult to maintain, experience at WJSV has definitely shown that it is unnecessary to employ special means of maintaining stability.

The directional antenna system at station WKRC is erected on the roof of the Hotel Alms, Cincinnati, Ohio, and consists of two self-supporting towers 154 ft. high and 1/8 wave apart. (Space, $\theta = 45°$). The current in the North antenna leads the current in the South antenna by 140° (θ - 140°). Figure 9 shows the skeleton block schematic diagram of this system. In order to obtain the proper phasing, an artificial line is used because of the relatively short transmission lines erected on the roof. The lines themselves, being approximately 110 feet long, each have an electrical length of approximately 30°. The additional 80° is obtained by properly adjusting the artificial line. With this arrangement, the field intensity distribution in a horizontal plane satisfies the general requirements of this case. A considerable amount of data concerning the operating characteristics of directional antennas was obtained during the design, construction and adjusting of the two systems described above.

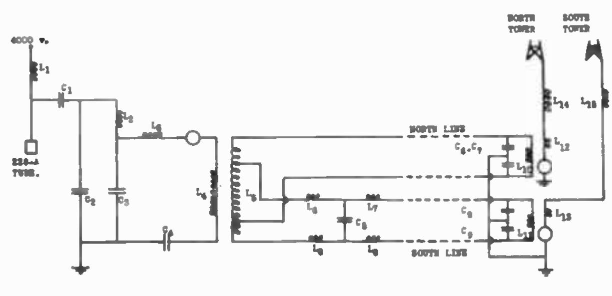


Figure 9

Schematic diagram of the WKRC directional directional antenna system.

Directional antennas are being used at regional stations to cut down interference in the areas served by other stations occupying the same channel, and it now appears that the use of this type of antenna will become more general in the future. Providing a directional antenna is properly designed and installed, it is possible to accurately predict its space pattern in advance[15,16]. However, it is not always possible to predetermine the efficiency of such an antenna system. Directional antennas should find wide application when and if synchronized station operation on a large scale becomes feasible.

TOWER LIGHTING AND PAINTING REQUIREMENTS

The marking of aircraft obstructions is set forth in detail in Aeronautics Bulletin No. 16, copies of which may be obtained without charge upon request from the Aeronautics Branch, Department of Commerce, Washington, D.C. The general requirements, as they affect radio stations, are summarized below.

Skeleton towers should be painted throughout their height with either alternate bands of chrome yellow or international orange (yellow No.4 and orange yellow No.5, respectively, of Color Card No. 3-1) and black, or alternate bands of international orange and white, terminating with either chrome yellow or international orange bands at both top and bottom, depending on color combination used. The width of the chrome yellow or international orange bands should be one-seventh the height of the structure for all structures less than 250 feet in height and from 30 to 40 feet for structures over 250 feet in height. The black or white bands should be one-half the width of the chrome yellow or international orange bands.

For night marking an aircraft hazard, a red obstruction light consisting of a 100-watt lamp in a red waterproof globe should be mounted at the top of structure.

For radio towers, or towers having a network of wires between the towers, additional fixed red lights consisting of 50-watt lamps in waterproof globes should be mounted on diagonal corners at the one-third and two-thirds points and so arranged as to be visible from any angle of approach.

Some areas which present a hazard to flying a civil airway, may require obstruction marking for night flying by use of lights of the high-intensity fix projector type. The high intensity fixed projectors should be 24-inch parabolic units using 1,000-watt lamps with lamp changers, should be pointed so as to envelop and outline the areas over which flying should be restricted, and should be elevated so that the luminous beams of light will intersect at the height of the obstructions to be cleared. In addition, such hazardous flying areas should be marked with one or more certified landmark beacons as conditions may require to give pilots a long range warning. Such beacons should be similar to the 300-millimeter airways electric code beacons of the double-Fresnel lens type with two 500-watt lamps and aviation red color shades, showing not less than 6 flashes per minute and having a luminous period of not less than 35 per cent. As an alternate system of marking such hazardous flying areas, certified 24-inch rotating landmark beacons equipped with 1,000 watt lamps and lamp changers and with red cover glasses and making 6 revolutions per minute may be used.

All lights marking hazardous flying areas should be exhibited from sunset to sunrise.

At the present time each radio station is being treated as a special case. The regulations outlined above are for advisory use only. In a new installation, it is now necessary to submit to the Federal Communications Commission (through Herbert L. Pettey, Secretary of the Commission), the plans for lighting and painting of radio towers. The submitted plans may be approved, or they may be returned with additional requirements which must be fulfilled.

Due to the tower obstruction lighting requirements, it is necessary to furnish up to 2 kw power for lighting the obstruction lamps. In the case of vertical radiators, these lamps are located on a structure at r-f potential, and means must be provided for isolating the lighting circuits on the tower from ground. Figures 10 and 11 indicate four methods which have been successfully used for this purpose. Use of the insulated generator as shown in Figure 10 is no longer necessary, since there are commercially available satisfactory chokes capable of isolating potentials up to 10,000 volts at broadcast frequencies.

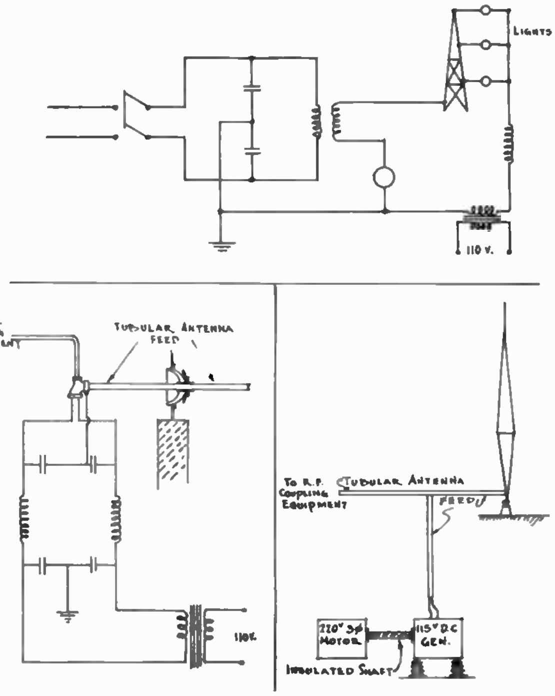


Figure 10

Three methods which have been successfully employed to transmit one or more kilowatts of lighting power to mast aircraft obstruction lights.

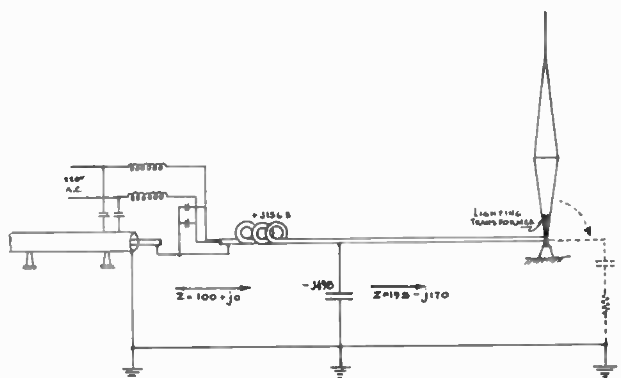


Figure 11

Method used to transmit lighting current to guyed vertical radiator at WLW. Note simplicity of method employed for coupling power from concentric transmission line to the antenna.

ANTENNA SYSTEM COSTS

Indicative of present costs of antenna systems employing either the guyed type or self supported type of tower, are Figures 12 and 13. The costs indicated on these graphs are based on average conditions in the field. In order to allow a station engineer to more intelligently figure the costs involved in the design of a complete antenna system, the following items are listed which should be included in his estimate:

Structural Steel
Insulators
Ground System
Foundations
Erection
Obstruction Lighting Equipment
Painting
Freight
Insurance
Engineering Expense

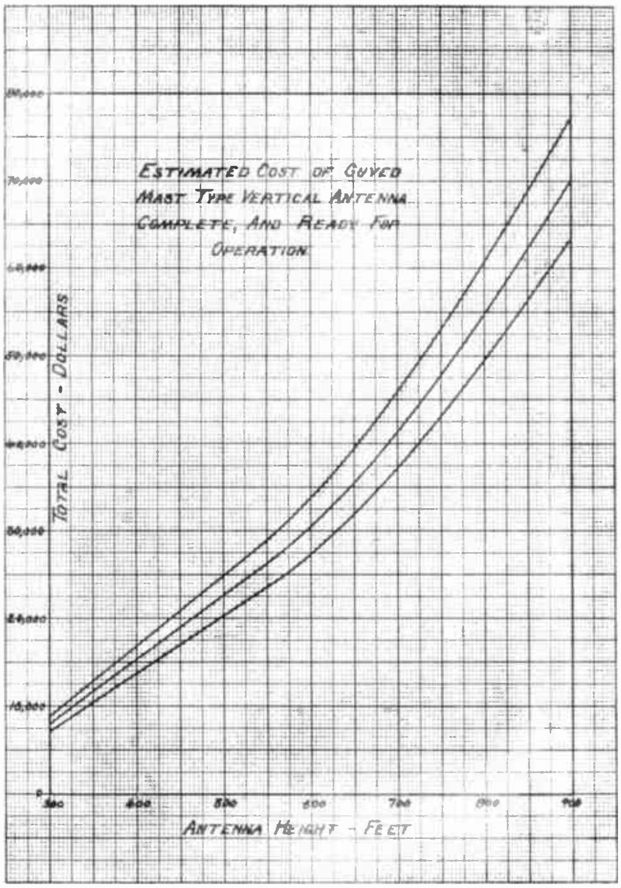


Figure 12

These items are included in the estimates shown in Figures 12 and 13. In a particular installation, one or more of the items indicated above might run considerably more than the average. This is particularly true in the cases of foundations, erection, obstruction lighting and freight.

It should be stated here that the most economical method of increasing the general efficiency of a broadcast station can usually be obtained by improving the antenna system. As an example of antenna economics, consider the following case. A 1 kw regional station desires to improve its coverage. It applies to the F.C.C. for a power increase to 2½ kw. If the application is granted, the average station incurs the following expenses:

Litigation	$ 1,500.
New Transmitting equipment complete with tubes and power supply	15,000.
Miscellaneous Modifications	1,000.
Installation of Equipment	1,500.
	$19,000.

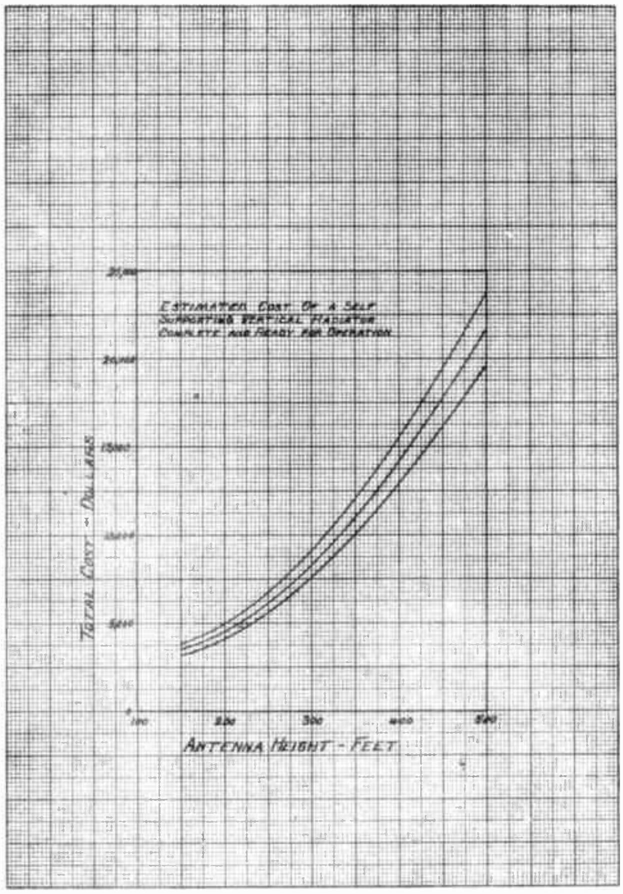


Figure 13

The annual operating expenses of the station will be increased by $3,000, plus amortization and interest on the above capital expenditure. The license for the increased power probably allows only daytime operation at 2½ kw.

That is the dollars and cents story of the power increase, but what does the millivolt and listener side of the picture show? The average 1 kw regional station in this country now has a signal output of 125 mv/m at one mile. For 2.5 kw this signal is increased to 198 mv/m - a 4 db gain in signal. A similar gain in signal can be made by the installation of a .25 wave self-supporting vertical radiator, whose cost will depend on the operating frequency. The estimated complete installed cost is as follows:

1500 kc	$ 4,200.
1000 kc	6,000.
600 kc	14,000.

The station would have saved money by purchasing a new antenna rather than higher powered transmitting equipment. A $3,000 a year saving would have been made since no increase in operating costs would be included with the new antenna. And the signal increase would have been full time rather than part time.

This example should indicate the economic importance of proper antenna design and the necessity of a careful balance of antenna cost with respect to the complete transmitting plant.

SUMMARY

1. The two-tower construction, which has been used in the past, is definitely outmoded by the single vertical radiator.
2. Use of a vertical radiator in place of an older conventional antenna will, in the average case, produce a signal increase equivalent to doubling the power of the transmitter. The exact signal gain will depend upon the efficiency of the existing antenna.
3. The self-supporting radiator may be used effectively at heights of approximately 1/2 a wavelength. It can confidently be expected to perform with approximately the same efficiency as the guyed type of antenna.
4. If the self-supporting tower antenna is used, precautions must be taken to prevent excessive dielectric losses in the soil near the tower base. A high base capacitance, of itself, does not contribute to low efficiency.
5. The ground system should be radial in nature and should consist of maximum amount of copper extending to the maximum radius consistent with economical considerations. In practice, this should mean a radius of at least 1/2 wavelength and a number of radials at least equal to 120.
6. To increase the non-fading area of a transmitter as much as possible, is an important factor of design which is not yet completely solved. There have been a number of studies which indicate that the optimum height, from the fading viewpoint, ranges between .45 and .60 of a physical wave length with the types of structures so far placed in service. The exact height cannot be determined until further work has been completed.
7. The vertical radiator is especially well adapted to use in directive antenna systems. Published theoretical methods of pattern calculation agree with field results. However, the efficiency of directive antennas has not yet been reduced to a mathematical process, and its determination must be based upon the engineer's experience.

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Annual Meeting

As announced, the 25th Annual Meeting was held at Columbia University on the night of December 3. The purpose of the annual meeting in the general discussion of the business affairs of the Club and, more specifically, preparation of nominations for the several offices as well as for the directorate of the Club. This meeting was, as is all too usual, only meagerly attended, although it was characterised by an extremely open and frank discussion of the potential nominees whose names were offered. As a result of the extended discussion and the subsequent balloting, the roster of nominees as given below was determined upon:

For President:
Harry W. Houck
Ralph J. Langley

For Vice President:
F.X. Rettenmeyer

For Treasurer:
Joseph Stantley

For Recording Secretary:
Keith Henney

For Corresponding Secretary:
Fred Klingenschmitt

For Members of the Board of Directors:

E.H. Armstrong	H.M Lewis
G.E. Burghard	R.H. McMann
A.B. Chamberlain	John Miller
C.L. Farrand	Fred Muller
L.C.F. Horle	L.W. Rosenthal
Frank King	C.R Runyon

W.A. Winterbottom

In connection with the nomination of candidates for office in the Club, the attention of the membership is called to the following excerpt from Article 7, Section 1, of the Constitution of the Club:

"Members unable to attend in person the Annual Meeting at which the above nominations are called for may obtain from the Corresponding Secretary a prescribed blank form on which they may nominate candidates for any or all of the above offices. Such nominations must be in the possession of the Corresponding Secretary within ten days after the Annual Meeting at which nominations were called for. Nominations made in this way must be three in number for any nominee to have him considered as such provided he has not already been nominated at the prescribed meeting."

Mr. Houck Withdraws

The Editor of the PROCEEDINGS is in receipt of the following letter from President Houck:

December 6, 1934.

Editor Proc. of the Radio Club of America,
90 West Street,
New York, New York.

Dear Sir:

I shall be grateful if you will find space in the forthcoming issue of the PROCEEDINGS of the Radio Club of America for this work of explanation for the withdrawal of my name from the roster of nominees for election to office in the Club.

Let me say first that I am thoroughly appreciative of the compliment which the Club pays me in nominating me for a second term as President, and I can assure the Club that my interest and active participation in the affairs of the Club will continue, to whatever extent I may be called upon to assist.

I feel, however, that in view of Mr. Langley's nomination for the presidency, and the good effect upon the management of the Club by the precedent of single term presidency of the Club, set by Mr. Sadenwater in 1931, and followed by all of the Club presidents since that time, that I can best serve the Club by this action.

I, therefore, addressed the Secretary, asking that my name be withdrawn, and I urge that all possible support be given to Mr. Langley in the coming election.

Respectfully yours,

Harry W. Houck

Proceedings

of the

Radio Club of America

Incorporated

June, 1935

Volume 12, No. 1

RADIO CLUB OF AMERICA, Inc.
11 West 42nd Street + + New York City

The Radio Club of America, Inc.

11 West 42nd Street - New York City

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PROCEEDINGS
of the
RADIO CLUB OF AMERICA

Volume 2 June, 1935 No. 1

MISCELLANEOUS APPLICATIONS OF VACUUM TUBES

BY

F. H. SHEPARD, JR.*

Delivered before the Radio Club of America
May 9, 1935

Introduction

It is generally accepted that a great number of seemingly impossible functions can be performed by the use of vacuum tubes. If enough tubes are used and complicated enough circuits are devised, almost anything can be done.

The object of this paper is to show a few of the numerous applications where ordinary vacuum tubes can be used in comparatively simple circuits to accomplish things that have heretofore required special tubes, expensive apparatus and comparatively complicated circuits. It is also the object of this paper to show that circuits can be devised in which variations of supply voltages and of tube characteristics will have little or no effect on the operation of the circuits.

1. A NEW LIGHT-RATIO INDICATOR CIRCUIT

In a great number of conventional phototube bridge circuits, it is possible to balance out the effects of light-source variations at one particular balance point. To the best of the Author's knowledge, there are no available circuits that can be used to indicate directly a ratio of the intensities of two light sources. A device which does this, is useful for making comparisons of the light transmitted or reflected by different specimens. For instance, in color matching the unknown samples can be set up so as to transmit or reflect light directly or indirectly to each phototube. A color filter is then placed between the light source and the samples. If a variation in the ratio is indicated by the device, it is obvious that one sample transmits or reflects more of the color in question. In this way, by putting various color filters in front of the light source we can match the samples in any desired light bands.

Because this method indicates changes in the ratio of light, we can use it successfully to compare the color of extremely small objects with a larger standard sample by mounting the small specimen against a black background.

Figure #1 Shows a circuit that will give a definite output versus light-ratio curve regardless of the actual intensities of the lights in question provided, of course, the intensities are greater than a certain threshold value. This value is determined primarily by the leak-

LIGHT-RATIO INDICATOR CIRCUIT

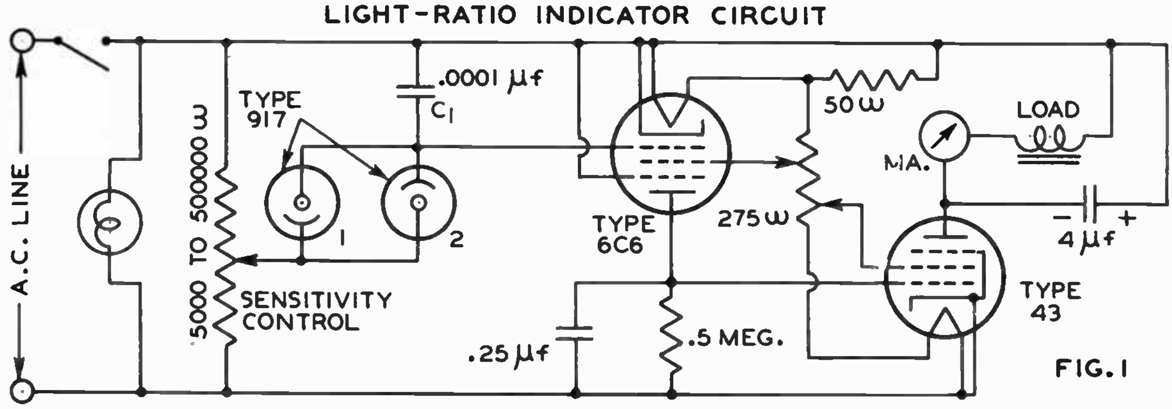


FIG. 1

*Research & Development Laboratory
Radiotron Division
RCA Manufacturing Company, Inc.

age currents of the phototubes as affected by dirt and moisture, of the amplifier tube, as well as the leakage current of C_1 and the grid current of the first amplifier tube.

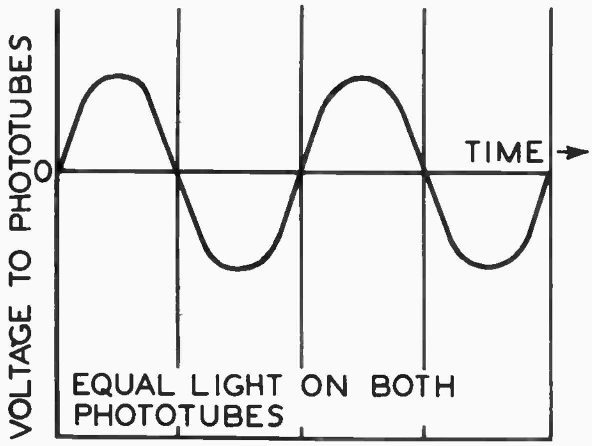


FIG.2A

Figure 2_a shows the a-c voltage which is applied across the phototubes when the voltage across C_1, in figure #1 is zero. The anode voltage on each photocell is positive for one-half the cycle and for an equal amount of time. Figure $\#2_b$ shows the currents drawn by the phototubes. The distance of the curve above the axis indicates the current drawn by photocell #1; the distance of the curve below the axis indicates the current drawn by photocell #2; the area under the curves above the axis is a function of current multiplied by time and is a measure of the charge fed into the condenser C_1 by photocell #1. Likewise the

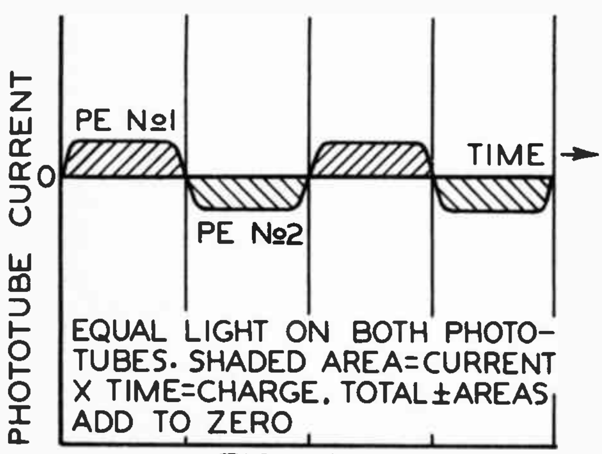


FIG.2B

area under the curve below the axis is a measure of the charge fed into C_1 by photocell #2. The average current going into or out of C_1 over a period of time must be zero. The areas included by the curve above or below the axis, therefore, should be equal. Now, suppose we double the light on photocell #1; this will cause the distance of the curve above the axis to be doubled, the distance of the curve below the axis remains unchanged. See Figure $\#2_c$. Momentarily, the area under the curve above the axis is increased and condenser C_1 charges, See Fig. $\#2_d$; this in turn causes the time that the anode voltage on photocell #2 is positive to be increased. Condenser C_1 thus charges until the areas above and below the curves are again equal. See Figure $\#2_e$. This takes place, when the a-c supply voltage to the phototubes is fixed, with a definite voltage across C_1 for every light ratio between the two photocells. The voltage across C_1 for any given light ratio will be directly proportional to the a-c voltage supplied to the phototubes. Thus the d-c voltage change across C_1, when the a-c supply voltage to the phototube is known, can be taken as a measure of the light ratio. This d-c voltage is indicated by the d-c amplifier shown, see Figure #1.

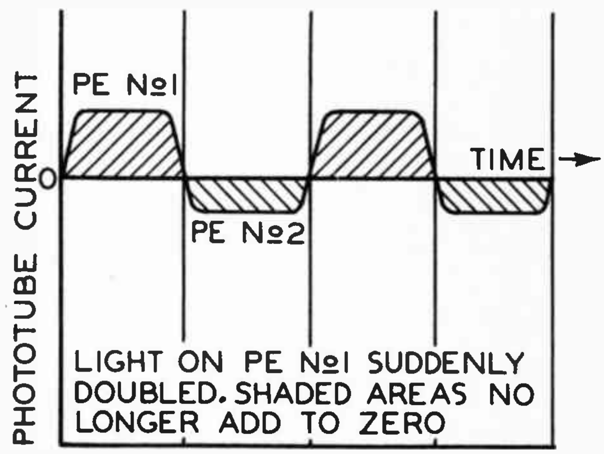


FIG. 2C

2. AN A-C OPERATED D-C AMPLIFIER

The use of an a-c supply voltage for the plates of a d-c amplifier has a definite advantage over conventional d-c amplifier circuits in that it eliminates the necessity of cascading "B" supply voltages or of using bucking or batteries between stages.

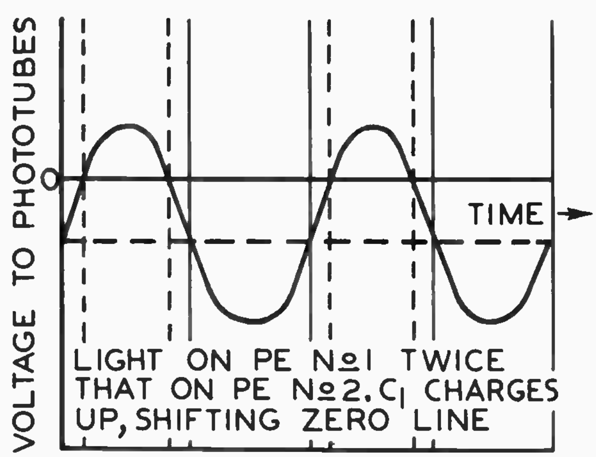


FIG.2D

Referring to the d-c amplifier used in figure #1 to indicate the d-c potential accumulated across C_1 by the phototubes as explained above, we can see that the plate supply is the a-c line. In considering the operation of this circuit, it is well to remember that the d-c potential between the two sides of the a-c line is zero. The tubes in this circuit can conduct only during the part of the a-c cycle that their own anodes are positive with respect to their own cathodes. Because of the direction of current flow thru the tubes, the drops in the load resistors are such that the plates of the tubes assume potentials negative with respect to the plate supply. Now, since the d-c value of the supply voltage is zero, the plates of the tubes in these circuits assume average potentials that are actually negative with respect to their own cathodes. The tube currents, and hence the drop across the plate load, is controlled as in conventional circuits by the grid bias. Because of this, the plate of the first amplifier stage, shown in figure #1, can be connected directly to the grid of the second stage to supply d-c. bias and signal to the output stage. The output will be a rectified pulsating d-c current which is smoothed out by the indicated electrolytic condenser and passed thru the output load.

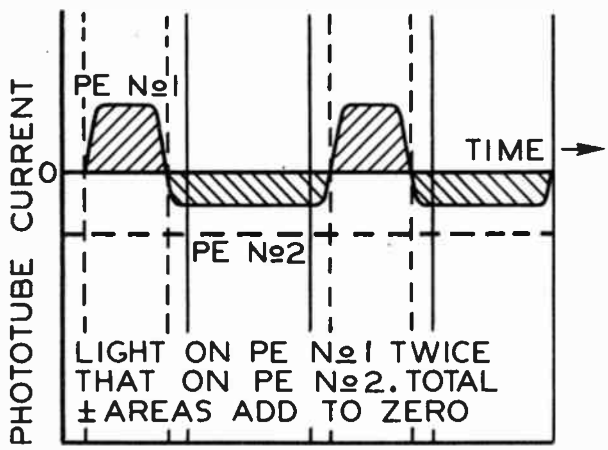


FIG. 2E

3. A SENSITIVE RELAXATION-TYPE CURRENT AMPLIFIER

Figure #3 shows what the author believes to be a new type of current-amplifier circuit. This circuit can be used to amplify the extremely small currents of a phototube which is receiving very small amounts of light. The circuit consists primarily of a phototube, a relaxation oscillator the frequency of relaxation of which is controlled by the light or current thru the phototube, a diode rectifier connected to the relaxation oscillator in such a way that the voltage it develops indicates the frequency of relaxation, and a power output tube controlled by the diode output voltage. The operation of the circuit is as follows:- The oscillator circuit oscillates violently, builds up a negative charge on its grid and suddenly blocks or stops oscillating. The oscillator will not again start oscillating until the charge on C_1 (see figure #3) has leaked off thru the phototube to such an extent that the oscillator will again conduct. Thus the time between the bursts of the oscillations of this oscillator are directly controlled by the rate at which C_1 is discharged thru the phototube.

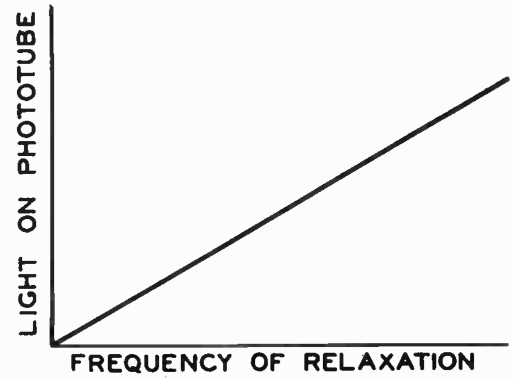


FIG. 4A

Figure #4a shows the linear relation between the frequency of relaxation and the light on the phototube. Figure #4b shows how the voltage developed by the above mentioned diode and fed as grid voltage to the power tube varies with the frequency of relaxation. When the frequency of relaxation is very low, the pulses of rectified current fed to the grid of the power tube are not numerous enough to cause any appreciably voltage drop across the power-tube grid resistor. As the frequency of these bursts of oscillation increases, more and more rectified current will be fed thru the power-tube resistor; this results in an increasing bias to the grid of the power tube. This bias starts out increasing directly with the frequency of relaxation and then it approaches the peak of the a-c voltage of the oscillator.

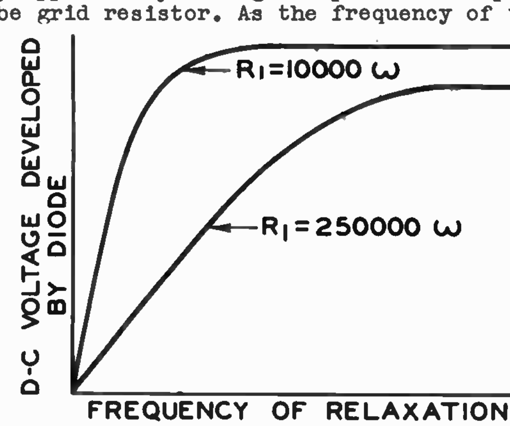


FIG. 4B

RELAXATION-TYPE PHOTO-AMPLIFIER CIRCUIT
FOR MEASUREMENTS OF SLOW LIGHT VARIATIONS

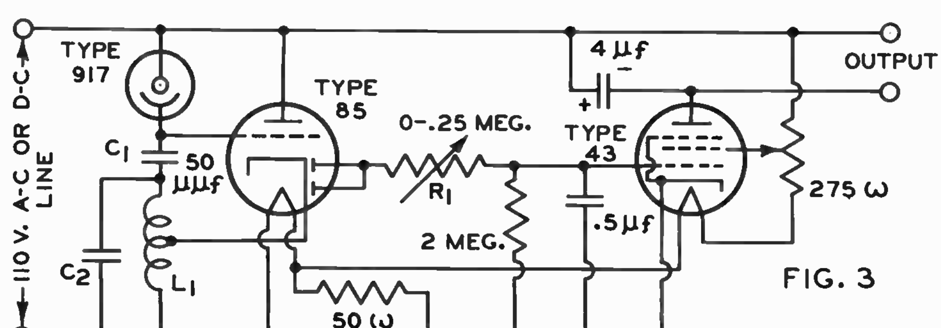


FIG. 3

See figure #4b. As R_1 is increased the current or charge fed to the grid of the power tube per individual burst of oscillation will be reduced. The upper curve in figure #4b is shown for a low value of R_1 while the lower curve is shown for a higher value of R_1. As the voltages on the grid of the power tube varies the power tube current also varies; thus, the output of the amplifier is a function of the power tube grid voltage which in turn is a function of the frequency of relaxation of the oscillator, which in turn is a function of the phototube current, which is in turn a function of the light on the phototube. Thus we can see that the output of this power tube is a function of the light on the phototube.

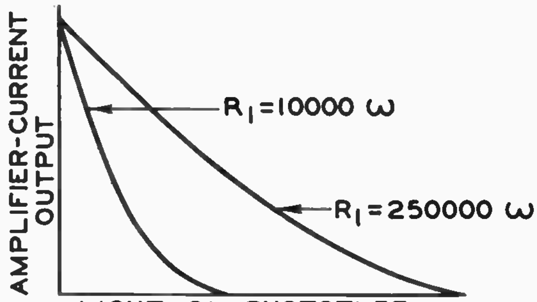


FIG. 4C

This is illustrated by figure #4c which shows two curves for two values of R_1. Theoretically, it is possible to reduce R_1 to a sufficiently low value to get full swing of the output for a definite percentage variation of a very small amount of light. The advantage of this circuit lies in the fact that the oscillator is completely inactive, that is, its plate current is completely cut off during the part of the cycle that the phototube is discharging condenser C_1. Because of this, any gas which may be in the tube can not be bombarded or ionized by electrons in the tube and hence cannot contribute any gas current to influence the accuracy of this device. In conventional circuits, gas current is one of the greatest limitations in making small current measurements. Using a demonstration model of this circuit in air, currents as low as one-thousandth of a microampere will operate a 20 milliampere relay. If the apparatus is placed in a dessicator much lower currents can be measured.

4. A CAPACITY-OPERATED RELAY

Another interesting vacuum-tube application is illustrated by the capacity operated relay circuit shown in Figure #5a. The sensitive part of this circuit consists of an oscillator, the feed-back, and hence the intensity of oscillation, of which is controlled by the antenna capacity to ground. The feed-back of this oscillator is a function of the difference in the ratio between the two parts of the oscillator coil and the ratio between C_1 and C_3. As C_1 or C_3 is varied, the feed-back is varied smoothly from a negative value thru zero to some maximum positive value. As this takes place, the intensity of oscillation varies smoothly from zero to some maximum value. In the circuit illustrated, a diode-triode combination type of tube is used as the oscillator. The diode is grounded for the oscillator frequency thru the 0.1 ufd condenser shown. Thus, the diode will develop a negative d-c voltage equal to the peak oscillator voltage present between the cathode of the oscillator and ground. This negative d-c voltage is fed to the grid of a power-output tube which can be used to operate a milliammeter or a relay. Figure #5b shows how the output current varies with the antenna-to-ground capacity (C_1) for various values of C_3. This circuit finds its usefulness in operating or initiating an advertising display at the approach of a customer, a burglar alarm, a door opener, etc. In a demonstration model of this circuit, the antenna-to-ground capacity is about 50 uuf. A person holding his hand about five feet away from the antenna will cause a two or three-milliampere output variation by wiggling his forefinger.

5. A ROADSIDE PHOTOELECTRIC TRAFFIC-SPEED INDICATOR

A simple roadside photoelectric speed indicator is shown in Figure #6. This circuit contains all of the necessary parts to indicate the speed of passing vehicles. A relay operates to indicate when a passing vehicle exceeds a predetermined speed, and the meter swings to indicate the speed of each passing vehicle. Briefly, the operation of the circuit is as follows: A car first interrupts the light to the upper photocell causing the grid of the first section of the tube type 19 to be made negative by the photoelectric current of the lower cell. This negative grid potential causes the plate current of the first section of the 19 to be cut off, the IR drop in the plate load resistor to decrease, and the plate potential to increase. The grid of the second

A CAPACITY-OPERATED RELAY

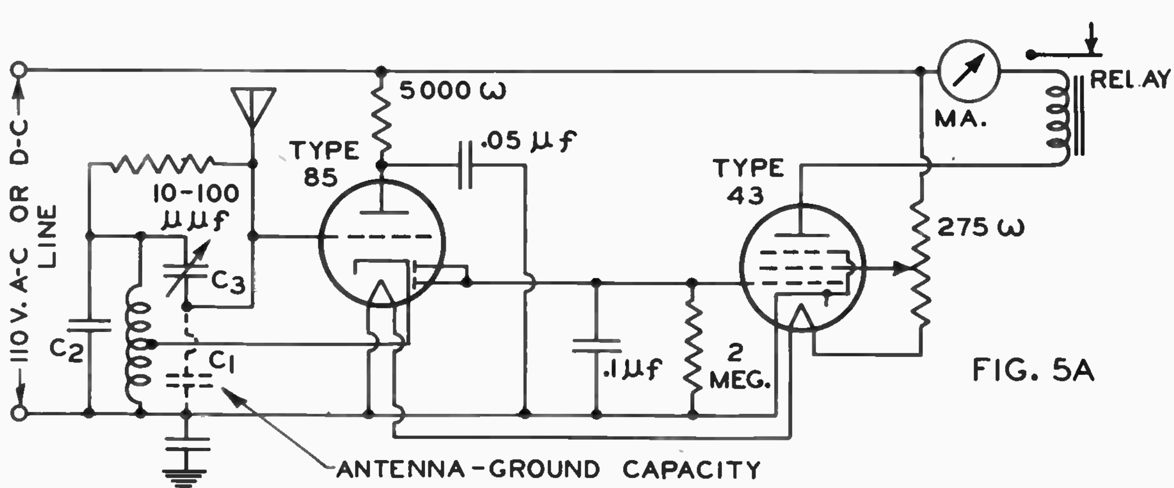


FIG. 5A

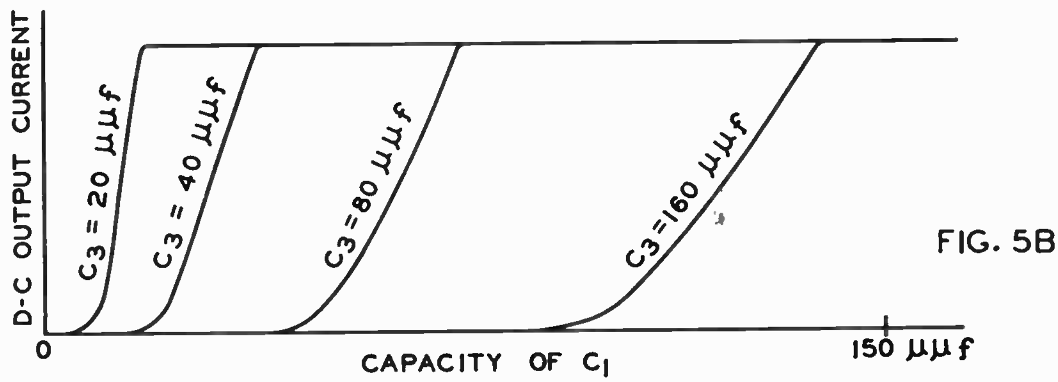


FIG. 5B

triode unit is pulled slightly positive by the increased positive voltage of the plate of the first tube section acting thru the condenser C_1 and register R_1. During this interval, condenser C_1 is being charged thru the plate-load resistor of the first triode unit, resistor R_1, and the electron current to the grid of the second triode unit. During this same interval because of the positive voltage on the grid of the second triode unit, its plate current rises and causes the relay to pull over. After an interval of time, the length of which depends on the speed of the passing vehicle, the light to the lower photocell will be interrupted. Thus, the light on both photocells is interrupted and the grid of the first triode unit is pulled slightly positive or to zero bias by the current thru the 10 megohm resistor. The positive or zero biased grid will cause the plate current of the first triode unit to rise, and its potential to drop to its former normal zero-bias value. Thru the coupling action of condenser C_1 and resistor R_1, the grid of the second triode unit will be driven negative below its normal bias by an amount proportional to the charge accumulated by C_1 during the interval between the interruptions of the lights to the two photocells. Since this negative voltage is proportional to the elapsed time between interruptions of light to the phototubes, it is, therefore, an inverse function of the speed of the passing vehicle. Thus, the reading of the output meter in the plate circuit of the second triode unit can be calibrated directly in miles per hour. If the elapsed time is long enough, that is, if the speed of the passing vehicle is slow enough, the charge accumulated will be sufficiently great to cause the grid of the second triode unit to be driven sufficiently negative to reduce its plate current below the point that will cause the relay to be released thus indicating the passage of a slow moving vehicle. If the passing vehicle had exceeded a certain speed, the relay would not have released. The relay used is of the usual magnetic type having considerable back-lash between its pull-over and release values. Because of this, the relay will stay either open or closed on the normal output current of the second triode unit.

A SIMPLE ROADSIDE PHOTOELECTRIC SPEED INDICATOR

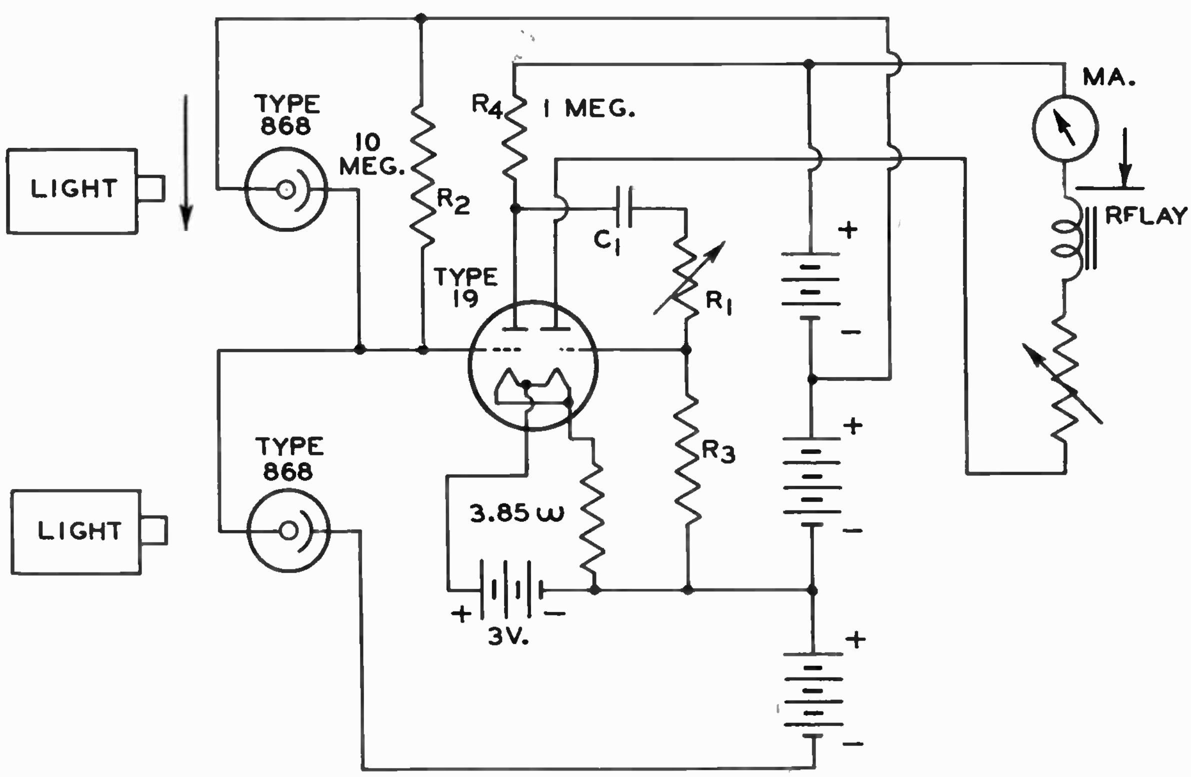


FIG.6

A PRACTICAL ROADSIDE PHOTOELECTRIC SPEED INDICATOR

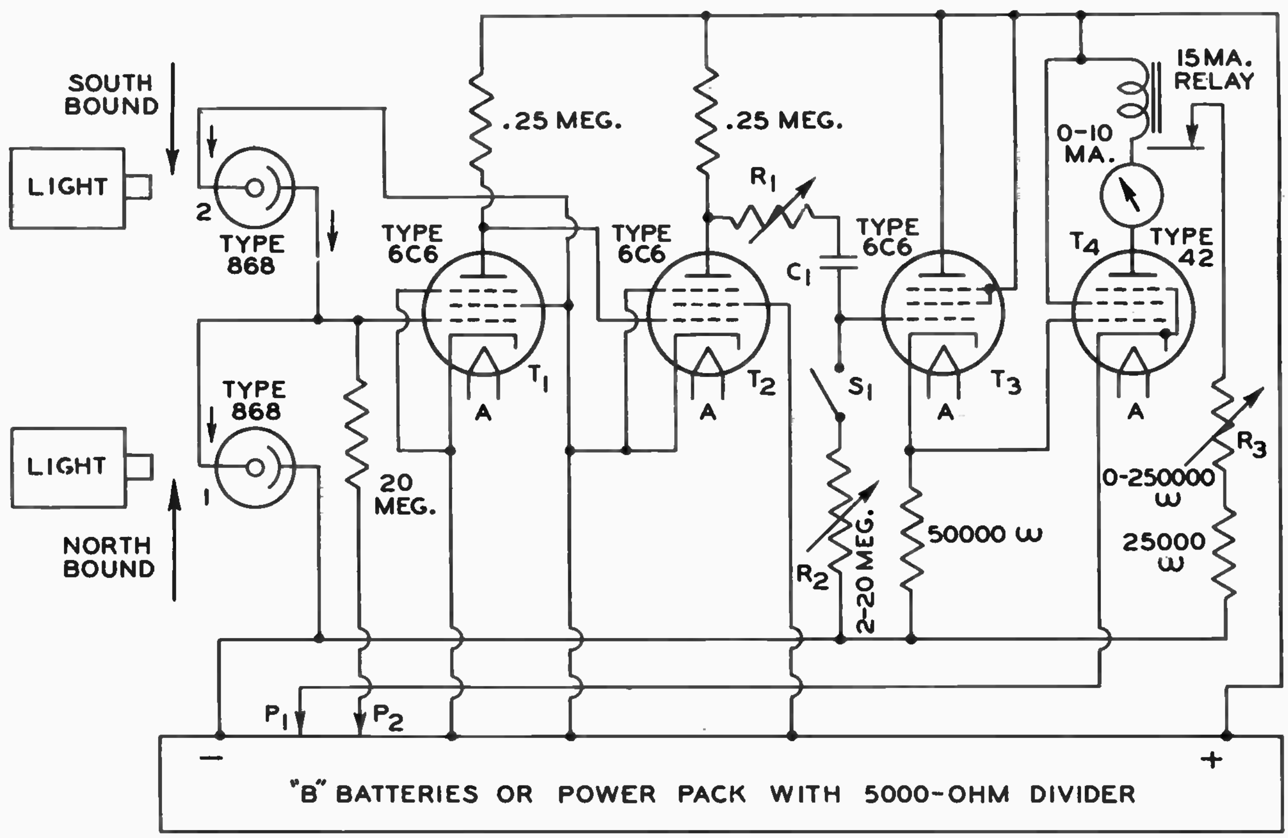


FIG. 7

Any charge accumulated by C_1 during the passage of a vehicle will leak off thru the grid-leak resistor R_3 and the device will be ready for the next passing vehicle. If the circuit constants of the device are properly set, it will be ready for the next vehicle even before the complete passage of the last vehicle.

Once adjusted and placed in operation, the device should require no attention other than an occasional check of battery voltages and of calibration.

Figure #7 shows a practical roadside-speed-indicator circuit. This circuit is more sensitive (that is, it operates on smaller amounts of light) than the circuit shown in figure #6 and is more easily adjusted. It will also indicate with equal accuracy the speed of vehicles passing in either direction. The operation of the circuit shown in figure #7 is traced thru step by step in the chart.

6. A HIGH-GAIN NON-MOTORBOATING AUDIO AMPLIFIER

All who have had occasion to build high-gain audio amplifiers have been confronted with difficulties from motorboating due to coupling from the output stages back thru the "B" supply to the input stages. This difficulty is often overcome by the use of very large filter condensers or by the use of separate "B" supplies for the various stages.

Figure #8 shows a stable high-gain resistance-capacity-coupled audio amplifier that has a voltage gain of more than 1,000,000 and which can be operated from a single poorly filtered, poorly regulated "B" supply. The stability of this amplifier circuit is due to the fact that it has a sharp low-frequency cut-off so that its individual "B" supply filters can effectively by-pass all the frequencies that are amplified, and so that the amplifier will not amplify any frequencies passed thru the "B" supply.

Figure #9_a shows the frequency characteristics of the amplifier as shown in figure #8. It is apparent that this amplifier has a sharp low-frequency cut-off. The sharp low-frequency cut-off of this amplifier is obtained by the combined effects of degeneration in the self-bias resistors, in the series-screen resistors, and in the coupling resistors. Figure 9_b illustrates how the gain of an amplifier stage varies with frequency when all of the supply voltages are fed from fixed voltage supplies except for the grid bias which is obtained from a self-bias resistor. As the frequency is lowered, the gain approaches a minimum which is the amount obtained when the self-bias resistor is totally unby-passed. As the frequency is increased the gain approaches a maximum. This is the value of gain that should be obtained using an infinite by-pass condenser or a fixed bias. This same curve is indicative of what happens when all the supply voltages are fixed, except for the screen voltage which is fed to the tube thru a series resistor which is by-passed to ground. The effect of the blocking-condenser-resistor combination is shown in figure #9_c. When the combined effects of the self bias, the series screen, and coupling resistors are properly matched, a frequency-response curve such as that shown in Figure #9_a will be obtained. When circuit time constants are considered, the tube impedances,

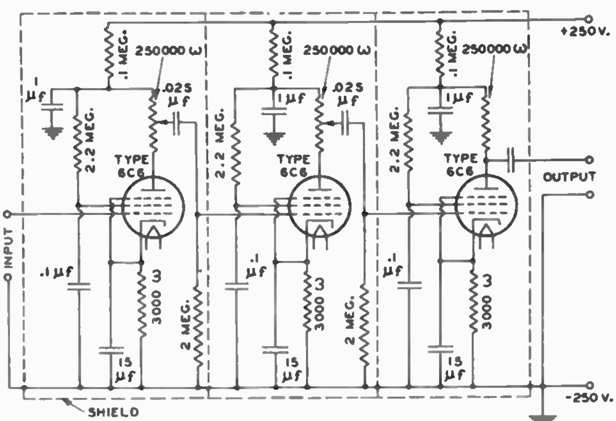
OPERATING DETAILS OF CIRCUIT SHOWN IN FIG. 7

POSITION	CAR	PE No. 1	PE No. 2	GRID OF T_1	GRID OF T_2	PLATE OF T_2	CONDENSER C_1	GRIDS OF T_3 & T_4	OUTPUT CURRENT	RELAY Makes On 15 Ma. Releases Below 5 Ma.
1	North Bound Slow	Light On	Light On	Cut Off	Zero Bias	Definite Low Voltage	Normal Initial Charge	Normal Bias	10 Ma.	Maintains Position
2	North Bound Slow	Light Off	Light On	Zero Bias	Cut Off	Approaches Supply Voltage	Charging	Drawing Current	50 Ma.	Closed
3	North Bound Slow	Light Off	Light Off	Cut Off	Zero Bias	Definite Low Voltage	Normal Charge + Accumulated Charge	Increased Negative Bias	Below 5 Ma.	Open
4	North Bound Slow	Light On	Light Off	Cut Off	Zero Bias	Definite Low Voltage	Accumulated Charge Leaking Off	Bias Returning To Normal	Approaching 10 Ma.	Maintains Position
5	North Bound Slow	Light On	Light On	Cut Off	Zero Bias	Definite Low Voltage	Normal Initial Charge	Normal Bias	10 Ma.	Maintains Position
1	North Bound Fast	Light On	Light On	Cut Off	Zero Bias	Definite Low Voltage	Normal Initial Charge	Normal Bias	10 Ma.	Maintains Position
2	North Bound Fast	Light Off	Light On	Zero Bias	Cut Off	Approaches Supply Voltage	Charging	Drawing Current	50 Ma.	Closed
3	North Bound Fast	Light Off	Light Off	Cut Off	Zero Bias	Definite Low Voltage	Normal Charge + Accumulated Charge	Increased Negative Bias	Between 5 & 10 Ma.	Closed
4	North Bound Fast	Light On	Light Off	Cut Off	Zero Bias	Definite Low Voltage	Accumulated Charge Leaking Off	Bias Returning To Normal	Approaching 10 Ma.	Maintains Position
5	North Bound Fast	Light On	Light On	Cut Off	Zero Bias	Definite Low Voltage	Normal Initial Charge	Normal Bias	10 Ma.	Maintains Position
1	South Bound Slow	Light On	Light On	Cut Off	Zero Bias	Definite Low Voltage	Normal Initial Charge	Normal Bias	10 Ma.	Maintains Position
2	South Bound Slow	Light On	Light Off	Cut Off	Zero Bias	Definite Low Voltage	Normal Initial Charge	Normal Bias	10 Ma.	Maintains Position
3	South Bound Slow	Light Off	Light Off	Cut Off	Zero Bias	Definite Low Voltage	Normal Initial Charge	Normal Bias	10 Ma.	Maintains Position
4	South Bound Slow	Light Off	Light On	Zero Bias	Cut Off	Approaches Supply Voltage	Charging	Drawing Current	50 Ma.	Closed
5	South Bound Slow	Light On	Light On	Cut Off	Zero Bias	Definite Low Voltage	Normal Charge + Accumulated Charge	Increased Negative Bias	Below 5 Ma.	Open
6	South Bound Slow	Light On	Light On	Cut Off	Zero Bias	Definite Low Voltage	Normal Initial Charge	Normal Bias	10 Ma.	Maintains Position
1	South Bound Fast	Light On	Light On	Cut Off	Zero Bias	Definite Low Voltage	Normal Initial Charge	Normal Bias	10 Ma.	Maintains Position
2	South Bound Fast	Light On	Light Off	Cut Off	Zero Bias	Definite Low Voltage	Normal Initial Charge	Normal Bias	10 Ma.	Maintains Position
3	South Bound Fast	Light Off	Light Off	Cut Off	Zero Bias	Definite Low Voltage	Normal Initial Charge	Normal Bias	10 Ma.	Maintains Position
4	South Bound Fast	Light Off	Light On	Zero Bias	Cut Off	Approaches Supply Voltage	Charging	Drawing Current	50 Ma.	Closed
5	South Bound Fast	Light On	Light On	Cut Off	Zero Bias	Definite Low Voltage	Normal Charge + Accumulated Charge	Increased Negative Bias	Between 5 & 10 Ma.	Closed
6	South Bound Fast	Light On	Light On	Cut Off	Zero Bias	Definite Low Voltage	Normal Initial Charge	Normal Bias	10 Ma.	Maintains Position

which in this case are considerably less than the external resistors, must also be considered. Because there is no convenient means of knowing exactly what these tube impedances are under the particular operating conditions, it is most convenient to determine the time constants or condenser values in these circuits experimentally. For instance, the curves shown by figures 9_b and 9_c can be made to cut-off or attenuate the gain of the amplifier at the same point. If these circuits are not perfectly matched, that is, if one starts to attenuate before the other, the resultant effect will be somewhat like that shown in figure 9_d. As the low-frequency cut-off of 9_d is not particularly sharp, it will be difficult to provide proper plate-supply filters to stop motorboating in an amplifier having the desired low-frequency gain.

The use of series-screen and self-bias resistors make this amplifier much less affected by possible variations in tube characteristics and by small variations of the circuit-resistor values than conventional circuits using fixed-supply voltages to all the tube elements.

HIGH-GAIN NON-MOTORBOATING AUDIO AMPLIFIER



MAXIMUM VOLTAGE GAIN > 1000000

FIG. 8

FREQUENCY CHARACTERISTICS OF AMPLIFIER
SHOWN IN FIG. 8

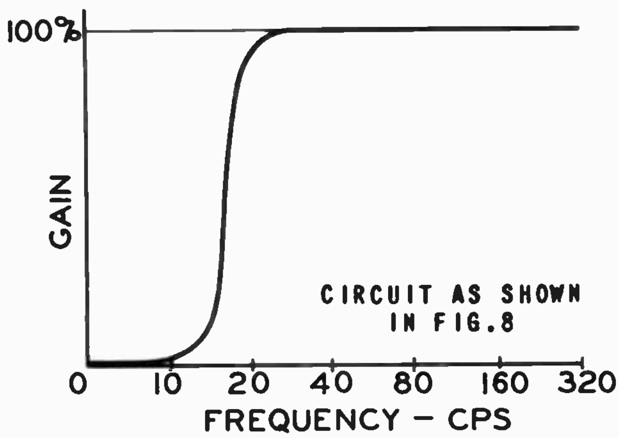


FIG. 9A

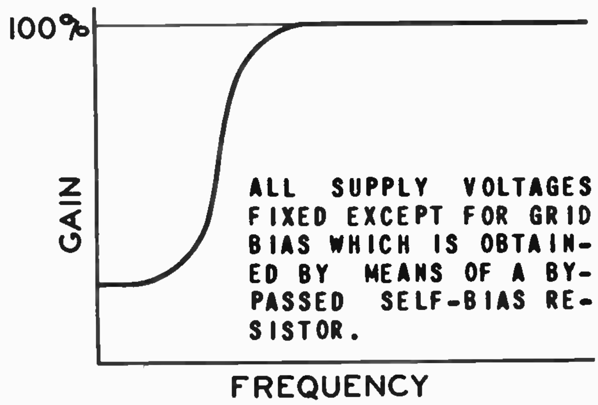


FIG. 9B

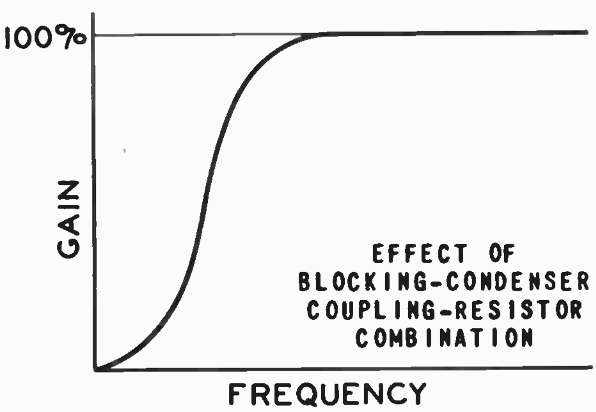


FIG. 9C

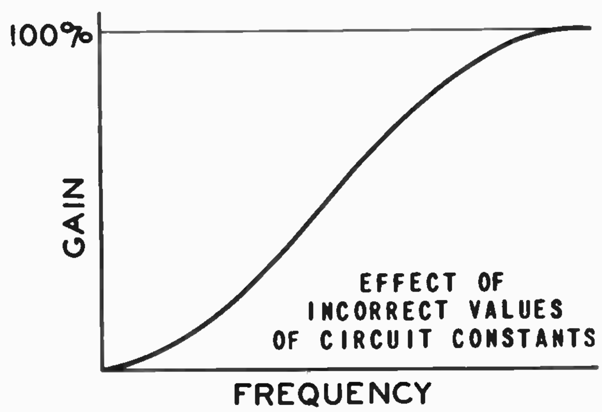


FIG. 9D

R. F. TRANSMISSION LINES

The December meeting, held on December 13th, was devoted to a symposium on radio-frequency transmission lines. Brief papers on several aspects of this extremely broad subject were read by Mr. C. W. Horn of the National Broadcasting Company, Mr. W. C. Tinus of the Bell Telephone Laboratories, Mr. Hugo Romander of the Kaltman-Romander Company and Mr. C. J. Franks of the Boonton Radio Laboratories.

Mr. Horn provided a brief paper on the historical aspects of the development of transmission lines in connection with broadcasting station design and operation; and his paper was, of course, replete with his usual and delightful reminiscences and philosophical digressions.

Mr. Tinus described in considerable detail the development and the practical form which is taken by lines and antennas for use in the ultra-high-frequency band, as typified by the newly installed police transmitter at Newark, New Jersey. Of especial interest was his brief story of the evolution of the antenna system there employed, and the samples of highly efficient, yet simple and economical, transmission lines for both low and high powers.

Mr. Romander discussed the matter of transmission lines from the standpoint of the amateur and supplied some exceedingly interesting estimates of the power distribution throughout a conventional system of antenna and line of the type commonly employed by the amateur. His review of the evolution and the reasons for the continued use of those types of antenna and line by the amateur, gave a most interesting picture not only of the state of the art in that field, but the fundamental economic and technical soundness of the methods used by the amateur.

Mr. C. J. Franks provided a short paper on the adaptation of the transmission line to the type of receiving system commonly employed in broadcast and all-wave receivers. His exposition of the details of the design of the several transmission lines--both R.F. and I.F. --in a remote-tuning type of receiver, for the design of which he was largely responsible some little time ago, was especially interesting.

The discussion brought out the fact that while the basic mathematical disclosure of the possibility of the propogation of waves along conductors was about at least as old as the practical disclosure of the means for the propogation of space waves, there appeared to be no simple explanation for the reason for the delay in applying the well-established fundamentals of the high-frequency transmission line theory to practical radio problems. It was suggested that, perhaps, the fact that much of the early development had been so largely in the hands of experimentalist might explain this condition in part, and the fact that both the instruction in and study of the propogation of waves on wires was so long in the hands of the pure physicists might explain that the now obvious simplifications both in the mathematics and physical concepts involved, resulting from the application of the rigorous analysis long available to the problem of high-frequency transmission lines, had been hidden from the engineer. It was, additionally, and more specifically, pointed out that for frequencies in and above the conventional broadcast range, the mathematical complications that may have retarded the adaptation of the transmission line to practical radio problems disappear when consideration is limited to transmission lines which are mechanically and otherwise suitable, and that once this simplification is thoroughly realized--as it is now only partially realized --their even more general adoption will doubtlessly result. Of especial interest was the point that with the simplifications of the mathematical relations, resulting from limiting considerations to high frequencies and to lines of conveniently large conductors, the engineer needs little more in the way of mathematics than simple trigonometry and the simplest of vector relationships to allow him to solve most of the problems of simple lines and antennas. It was urged that the membership investigate this possibility for themselves, and thus make themselves familiar not only with the simple mathematics of these useful electrical structures, but to note how completely the physics of their operations are revealed thereby.

It was further agreed that the difficulty of adjusting any system in which the frequency-determining circuit is directly connected to the transmission line may well have accounted for much of the delay in the adoption of transmission lines to practical radio problems and, in fact, have forced the adoption of the transmission line to wait upon the development and general use of the master-oscillator type of transmitter circuit.

Some objection was raised to the inference that the use of the transmission line in radio transmitters is of wholly recent origin, it being pointed out that within the experience of those present certain uses of lines had been made even before the advent of modern broadcasting. In the case of the early work with KDKA, as described by Mr. Horn, it was found, largely by accident, that the use of a relatively remote antenna connected to receiving equipment through a pair of wires, rather than through a single connecting wire, gave certain operational advantages; the reason for which was not clear at the time but appears now to have been a result of the limitation of true antenna effect to the remote structure and the provision of relatively efficient transmission in the connecting lines. In advance of these uses of transmission lines, however, it was noted that the location of transmitters on the lower levels of certain of our battle cruisers early in the late war, and the provision of antenna down-leads through large concentric shielding cylinders provided what was probably another partially, at least, inadvertent adaptation of the transmission principle. But earlier, by far, than these was the application described by the oldest member of the Club, Mr. Robert Marriot, in his work at Jersey City in 1910, in which the housing of the radio-frequency generator remote from much of the associated high-frequency apparatus forced the employment of a two-wire connection between the two in accordance with what was then known as the Shoemaker system which, in the light of subsequent developments, probably constituted the forerunner of a carefully adjusted, efficient, modern transmission line.

Whatever the cause of the slowness of the general adoption of transmission lines to radio transmitter problems, it was generally agreed that their introduction had freed the whole problem of transmitter operation from some of the most severe limitations. Without the use of transmission lines the elimination of the influence on transmitters of the intense fields directly under an antenna would, of course, have been difficult, if no economically impossible; and without its general use the devising and use of many of the extremely useful and effective directional arrays would have been made intensely difficult, if not downright impossible; and without its general use the highly desirable disassociation of the fragile, care-requiring, transmitting equipment from the rugged, weather-proof antenna structures would have been impossible, and the economic advantages of such a disassociation would doubtlessly have delayed the rapid development of all sorts of transmitting systems to a tremendous degree.

Much interest was shown in the description of the evolution of the UHF transmitting antennas as are used especially in police work. It was pointed out that one of the fundamental advantages of the ultra-high-frequency type of transmission is the fact that highly efficient antenna are economically possible; that is, that antennas having efficiencies of radiation quite without economic justification in the broadcasting service, because of the tremendous size and cost which are necessary, are readily possible in the UHF range because of the small size and cost of the radiating structure. Thus in the usual UHF transmission system, such as the police transmitter discussed, a half-wave radiator with its extremely high efficiency of radiation consists of about fifteen feet of copper, or brass, tubing of such dimensions as make it ruggedly self-supporting. There is, however, left a considerable problem in the devising of the line for the feeding of such an antenna since the very nature of the quasi-optical transmission in the UHF band requires that the antenna be located at the highest possible level in order that its horizon be as remote as possible. It was pointed out, however, that the half-wave radiator offers a pure resistance to a line which feeds it at any point along its length and thus provides the essential requirement for the use of a transmission line with minimum losses. Thus, such a half-wave radiator might be fed by a line at its center by the proper choice of line constants to provide for a matching of the characteristic impedance of the line and the resistance of the antennas as viewed from the point of attachment of the line. Such a structure includes a serious and difficult mechanical and electrical problem in the attachment of the line to the antenna at any point other than an end, and makes desirable the devising of means for providing for such an end attachment. Such an attachment, however, while solving the mechanical problems involved, in that it provides for the antenna being essentially a mechanical extension of the line proper, and thereby provides a simple and rugged structure, is complicated by the fact that where the size of the line and the antenna are approximately the same--as is mechanically desirable--the resultant lack of impedance match between the characteristic impedance of the line and the resonant end impedance of the antenna requires that some special expedient be resorted to to bring about proper matching. Such a matching can, it was shown, be brought about by taking advantage of the especial properties of half- and quarter-wave lines.

More specifically, it was indicated that a half-wave line of sufficiently low attenuation as is simply and practically possible, when terminated in any resistance, will provide at its unloaded terminals substantially the same value of impedance as constitutes its termination at its loaded terminals. While, in the case of the quarter-wave line, it was indicated that a suitably low-loss line terminating in a resistance, and fed by a transmission line will provide against reflection back into the line, and hence give greatest efficiency of transmission if the characteristic impedance of the quarter-wave line is the geometric mean between the loading resistance and the impedance of the transmission line proper.

This latter provides a workable solution to the UHF antenna problem. In practice, the antenna consists of a half-wave radiator which is structurally a part of one portion of the transmission line, while the other side of the line extends downward from the lower end of the antenna proper for a quarter-wave length at which point both sides of the quarter-wave line join the transmission line which makes connection with the transmitter. It is, of course, an essential element of this arrangement that the physical size of all the elements be so chosen as to give the required relationship between the several impedances.

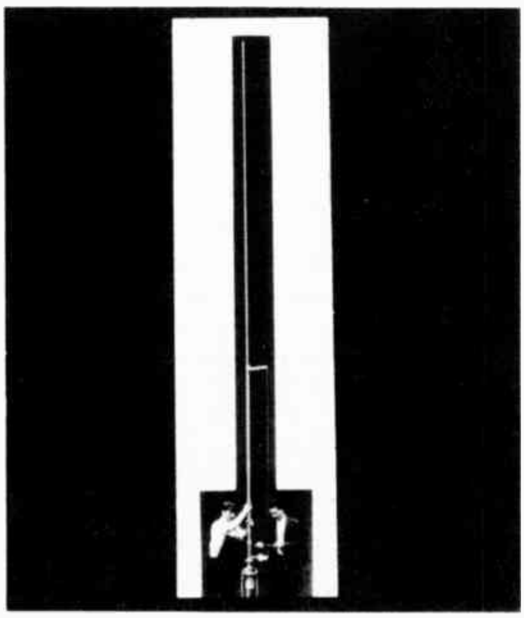


Fig. A

In Fig. A is shown a photograph of the actual arrangement used in Newark, New Jersey, where the 500-watt police transmitter operates at 30.1 megacycles. From this will be noted the physical sizes of the several parts of the system; it being, perhaps, necessary to point out, however, that the flag pole on which the antenna is located, provides the housing for the transmission line proper and above which is mounted the antenna itself.

Somewhat similar arrangements were reported to be in use in connection with the 60-megacycle transmitter atop the Empire State Building. In this case, further emphasis was placed on the point that in the UHF range the need for highest possible elevation and the unavoidable climatic hazards accompanying such location, bring with them the need for consideration of the mechanical details with care and completeness quite coordinate with the consideration required by the purely electrical aspects of the problem.

Samples of several standardized forms of transmission lines were shown. These consist essentially of copper tubing through which pass a concentrically located conductor of considerable size, contact between the two conducting members being provided against and the continued central location of the inner conductor being assured by the use of low-loss ceramic beads at frequent intervals along the length of the inner conductor, the beads being held in place along the inner conductor by slight deformation of the conductor proper, thus resulting in a combination conductor that can be bent and manipulated without damage to its operating constants. Where, however, something more closely approaching the flexibility of the conventional armored cable of the power-circuit practise is required, an alternate form of concentric conductor employing a braided covering of copper and including the ceramic beads is available. Samples of this were also shown.

Much of the discussion centered on the alternate possible forms which might be employed for the radiator and its association with the line--all of which further emphasized the need for the complete coordination of the mechanical and electrical details of the design of the antenna, line and supporting structure. An interesting form of antenna employing no matching elements was suggested for discussion.

It was pointed out that since the half-wave antenna offered a pure resistance at any point along its length--including its end--and since the characteristic impedance of a single-wire transmission line and the end impedance of the antenna were both simple functions of the diameter of structures constituting them, it should be possible to provide an antenna of so large a diameter that the end impedance of the antenna and the characteristic impedance of the line are substantially equal and so obtain a suitable match. Objection was raised to the usefulness of such a structure, in that it was felt that the large diameter of the antenna required to give an end impedance sufficiently low to match any otherwise useful line, would seriously interfere with the radio effectiveness of the antenna proper (or, as the speaker facetiously but tersely put it, "the standing waves on the antenna would not know which way to stand") although it was equally strongly felt that this was probably not of serious importance. The problem of the attachment of the line with a sufficiently sharp discontinuity in the physical structure to limit the standing waves to the antenna proper might, it was thought, offer some difficulties.

It was implied that the economic advantages of the short, cheap, highly efficient radiator of the UHF range were largely off-set by the quasi-optical characteristics of the radiation therefrom, which require an extremely great elevation of the antenna from the ground in order to secure a usefully distant horizon and an area of coverage of such magnitude as to justify the cost of the transmitting equipment. As against this, however, it was pointed out that it seems to be a characteristic of American cities that the population distribution with respect to the area of their tallest buildings is largely identical, and in addition, the relationship between the height of the tallest buildings and the population is fairly constant with the result that there is almost invariably already available in almost any of our larger cities a perfectly suitable antenna support in the form of the larger office buildings. Thus, much of the cost of the supporting structure required by the UHF antenna has already been absorbed in the cost of the available buildings. It was noted also, however, that without the availability of efficient transmission lines to provide for the location of the antenna at the topmost point of such buildings, while providing for the location of the remainder of the transmitting equipment at some other and more suitable point, much of the inherent advantage of these fortunate conditions would be lost.

In contradistinction to the niceties of design and construction which characterize the antennas and lines referred to in the preceding is the adaptations which the amateurs have made of the fundamental principles already discussed to their specific problems within the severe limitation, both technical and economic, under which, for the most part, they operate. For purely amateur uses it is obvious that any arrangement must inevitably be of low cost, easy of construction, adjustment and operation, and as efficient and reliable as is consistent with these requirements. For the most part the non-resonant transmission line finds little application in the amateur field because of the precision of construction and adjustment required for its successful use. And, in addition, the fact that the amateur almost invariably requires his transmitting system to operate on two or more of the amateur bands of frequencies, in addition to easy adjustment to the entire range within these bands, further points to the lack of suitability of the conventional types of matched lines. In addition to these completely rational reasons for the need for something other than the conventionally designed arrangements, it is probable that tradition which so largely determines the forms taken by amateur apparatus has had its influence in bringing into general use the so-called "Zep"- type of combination antenna and feed-line. It is something of a question as to whether the use of the approximately quarter-wave line, with one side extended beyond the other by an additional half-wave, was really introduced to the radio art in connection with radio transmission from lighter-than-air craft such as the Zepplins but, whatever the origin of the name, this type of antenna system, has found extremely wide-spread use and its description serves to indicate the form of structure commonly used by amateurs throughout the world.

In general, the amateur finds it quite impossible to provide himself with a mast or other supporting structure which will allow him to equip himself with a vertical antenna, notwithstanding the obvious advantages of this form of antenna as against the horizontal type he usually employs. This latter type requires only that it be attached to such supporting structures as are available to him--as for instance the roof-tops of his own home and that of his neighbors--and that it be fed in some manner that will deliver most of the output of his transmitter to the radiator. He usually chooses to make the length of the horizontal antenna a half-wave length for one of the lower frequencies of the amateur assignments using the same structure as a grounded antenna for the lowest wavelengths on which he operates and resorting to such expedients as are found necessary and workable to provide for operation at higher frequencies. The transmission line, or better, perhaps, the feed-line, is then chosen to be approximately a quarter-wavelength long and effort is made to supply it out of an impedance which matches the combination of half-wave antenna and quarter-wave line, with provision being made for the resonating of the reactance of such coupling as may be used and, incidentally, resonating the line to whatever degree may be necessary in view of the range of frequencies over which it is to be used within any one of the assigned bands.

Such a line-antenna combination results in approximately the same relationships attributed to the UHF antenna referred to previously, in that the characteristic impedance of the line is approximately the geometric mean of the terminating impedances. Or, more intelligibly, the combination of line and antenna results in a terminal impedance at its supply end determined by the ratio of the square of the characteristic impedance of the feed-line to the end impedance of the antenna. These simple relationships are, for several reasons, only approximately realized; but they serve to indicate the magnitude of the quantities involved. Thus, for the conventional 600-ohm line and an antenna having an end impedance of 10,000 to 15,000 ohms, a supply-end impedence of the order of 25 ohms results. This is an especially happy choice for the amateur, not only because the low value of the impedance makes the transfer of power to the line easily and conveniently possible, but also, which is equally important, it results in input currents of such relatively high value (in view of the powers commonly used) as to make measurement during adjustment easy with relatively simple, rugged and cheap instruments, and provide that stimulation to the operator which results only from the evidence of sizable and generally impressive values of current.

It is without point to neglect the significance of these purely psychological factors, since it is just such factors as these as have stirred the imagination and the ambition of the amateurs to strike out into new and previously little-explored fields, and thus to blaze the way for the more highly skilled, but perhaps less enthusiastic and persistent, professionals who fill the voids left in the work of the amateurs, and building upon this joint foundation, rear the precisely organized structure upon which commercial radio depends.

Nor is it to the point to be too critical of the amateur antenna system, even from the purely technical viewpoint; since, as was shown by the illustration given by Mr. Romander, it has little to apologize for even from the standpoint of the efficiency of operation.

More specifically, a set of data was supplied and is given in the following table, showing something of the power loss in the usual amateur antenna and in the com form of "tuned" line used therewith.

Frequency	Loss	Size	Efficiency
1.8 mc	2.35%	#10	(.935)N
1.8 mc	4.8 %	#16	(.875)N
14.2 mc	0.82%	#10	(.975)N
14.2 mc	1.66%	#16	(.953)N
58 mc	0.4 %	#10	(.988)N
58 mc	0.8 %	#16	(.976)N

$$(N = \frac{4L}{\lambda}.)$$

These sets of data are the result of careful computation, based on several simplifying assumptions. The second column shows the ohmic power loss in the antennas as compared to the power radiated, and is, in fact, the ratio of the computed high-frequency resistance of a half-wavelength wire properly weighted in accordance with the sinusoidal current distribution to the 75 ohms which is commonly accepted as the radiation resistance of such a wire. It is especially to be noted that, in all cases, the power lost in the conductor proper is only a few per cent of the total power lost in the antenna and, of even more importance, it is to be noted how little is lost through the use of what would probably be classed as a commercially unsuitably small wire size as compared with the larger wire sizes.

In the fourth column is given the results of a similar set of calculations involving, however, the transmission line and expressed as the efficiency of the lines. The figures given are such as to provide for the direct application to lines of more than a quarter-wave in length, since N is merely the ratio of four times the line length to the wavelength. Here again, it will be noted that the efficiency of transmissions shows itself to be strikingly high. And, in general, it is to be concluded that the apparently crude structures which are typified are, after all, quite respectable in their operating characteristics.

In the course of the discussion it was objected that no losses had been charged off against such insulating spacing members as might be included in the line structure and that, perhaps, even the radiation losses of the line might be an appreciable portion of the power loss. It was felt, however, that in view of the amateur's less rigorous requirements as to precise line spacing and the relatively few line spacers commonly employed, that the di-electric losses in the line structure might without serious error be neglected. And that, in general, the radiation losses from the lines were not of such serious magnitude at even the highest frequencies as to require inclusion in these admittedly unprecise but, withal, interesting and significant figures.

It is not, however, to be assumed, that only the "Zep"-type of antenna is employed by the amateurs since, as was pointed out, the center-fed, open doublet does find considerable use. In this case the dimensions which characterize the previously discussed structure provide two quarter-wave antennas feeding from opposite sides of the line which itself is usually made the equivalent of half-wave in length, and thus provides at its input terminals approximately the 75 ohms provided by the antenna which loads the other end of the line. This value of impedance is sufficiently low to meet the technical as well as the purely psychological needs of the amateur, and is found to serve quite well.

MEETING NOTES

January Meeting:

The meeting on January 10 was devoted to a paper by Mr. S. Y. White on the subject of signal-seeking circuits. Such circuits are characterized by the fact that when the system including them is tuned by the usual manipulation to be in only approximate tune with the signal that it is desired to receive, the automatic action of the circuits bring the system into more precise tune with no further manual adjustment. Many possibly useful schemes, both mechanical and electrical were described by Mr. White, amongst them a purely electrical one which he had developed and demonstrated, showing the operating characteristics of that type of automatic action.

The extended discussion that followed the presentation of the paper indicated the keen interest of the attendants in this type of circuit arrangement and, in addition, brought out the fact that work on arrangements of similar function had been in progress for some time both here and abroad.

February Meeting:

The meeting of February 14 was devoted to a paper by Mr. Glen F. Gillette on the subject of "Theoretical and Practical Aspects of Station Coverage as an Engineer Sees Them." It will be remembered that Mr. Gillette had devoted the last several years to the measurement of the transmission characteristics of a large number of American broadcasting stations, and on the basis of data so gathered was able to offer a most interesting summary of this data as well as certain conclusions as to the means that might be employed in the predetermination of these characteristics as well as for their general improvement.

While unusual preparation has been made for the formulation of Mr. Gillette's paper for publication, it has not yet been possible to complete it for that purpose. It is expected, however, that it will be available for the September issue of the Proceedings and will there be published.

March Meeting:

The meeting of March 14 was devoted to a paper by Mr. John F. Rider on the subject of "Engineering Developments in the Service Field for the Future". Mr. Rider pointed out at considerable length the need for providing for the servicing of radio receivers such apparatus as will make analysis of the causes for receiver failure---both partial and complete---more easily and expeditiously possible. For this purpose he proposed the use of two cathode-ray tubes in a particular circuit relation which he described and discussed. The discussion of Mr. Rider's paper directed itself as much to the educational and economic problems of the servicing of radio receivers as to the purely technical problems involved.

April Meeting:

The meeting of April 10 was held in the special auditorium of the General Electric Building in New York City and was devoted to the series of electrical and optical demonstrations that are commonly there given under the intriguing title of the "House of Magic". While the several demonstrations were unusually well prepared and generally interesting, it was most unfortunate that the formal lecture accompanying them was obviously designed for the non-technician and the willing-to-be-mystified, and of little, if any, interest to the attendants. Happily, however, the latter portion of the meeting was enlivened by a discussion of the then newly-announced all-metal tubes, of which a series of samples were available for examination.

Under these conditions the equation for the primary current is,

$$I_1 = \frac{E}{R+j\left(\omega L-\frac{1}{\omega C}\right)+\frac{\omega^2 M^2}{R+j\left(\omega L-\frac{1}{\omega C}\right)}} \quad ------(13)$$

in which E, R, L, C and ω have the same significance as in equation (1), R, L and C being the same for both primary and secondary, and M is the mutual inductance between the primary and secondary tuning coils.

Equation (13) may be simplified by applying the same approximations that were used to reduce equation (1) to equation (9), which is equivalent to substituting $R(1+jv)$ for $R+j\left(\omega L-\frac{1}{\omega C}\right)$

Making this substitution in (13), we have,

$$I_1 = \frac{E}{R}\cdot\frac{1}{(1+jv)+\frac{\omega^2 M^2}{R^2}\cdot\frac{1}{(1+jv)}} \quad -----(14)$$

We may assume that, for a given value of M, $\omega M/R$ is very nearly constant over the frequency range of interest. Let $\omega M/R = m$. Then (14) becomes,

$$I_1 = \frac{E}{R}\quad\frac{1}{(1+jv)+\frac{m^2}{(1+jv)}} \quad ------(15)$$

Rationalizing denominators and collecting terms, we have,

$$I_1 = \frac{E}{R}\cdot\frac{1}{\left(1+\frac{m^2}{1+v^2}\right)+jv\left(1-\frac{m^2}{1+v^2}\right)}$$

$$= \frac{E}{R}\cdot\frac{\left(1+\frac{m^2}{1+v^2}\right)-jv\left(1-\frac{m^2}{1+v^2}\right)}{\left(1+\frac{m^2}{1+v^2}\right)^2+v^2\left(1-\frac{m^2}{1+v^2}\right)^2} \quad ----(16)$$

Equation (16) is in the form of two factors, the first of which is the amplitude of the current at resonance of the uncoupled tuned circuit, the second being a function not only of v, as in the case of the single tuned circuit, but also of m which represents the degree of coupling. When m = 0, equation (16) reduces to (9) as it should, because the primary then becomes an uncoupled tuned circuit, and the tuning coil of the secondary is fixed in a position where none of the magnetic flux from the primary coil links with the turns of the secondary coil. When m is increased from zero, which is equivalent to altering the position of the secondary coil so that it links with more and more of the flux from the primary coil, marked, but gradual, changes occur in the phase, amplitude and frequency characteristics of the primary current. In order to show these changes, we have elected to halt at several arbitrary values of m and plot characteristics corresponding to those already given for the single tuned circuit. Figs. 2a, 2b, 2c and 2d show such curves, each figure containing a set of curves for m = 0, 1/2, 1 and 2. That is, m is substituted in equation (16) and a series of values given to v for each value of m. Fig. 2a shows the real term, in phase with the impressed e.m.f., fig. 2b, the imaginary term, in quadrature with the impressed e.m.f., and fig. 2d, the absolute magnitude, all versus v, which has exactly the same significance as explained in connection with the single tuned circuit. Fig. 2c is plotted from equation (16) in the same manner as fig. 1c, so that a straight line connecting the origin with any point on the curves, corresponding to some value of v, shows by its length the amplitude of the current, and, by the angle it makes with the x-axis, the phase with respect to the impressed e.m.f.

The values of m selected, incidentally, also provide a basis for comparing the curves as to critical coupling (m = 1), less than critical (m = 1/2) and greater than critical (m = 2). The significance of critical coupling is treated in detail in most text books dealing with coupled tuned circuits, and we need only mention at this point that it represents a borderline case in the shape of the frequency characteristic of the current in the secondary circuit and will, therefore, be discussed more in detail in a later paragraph.

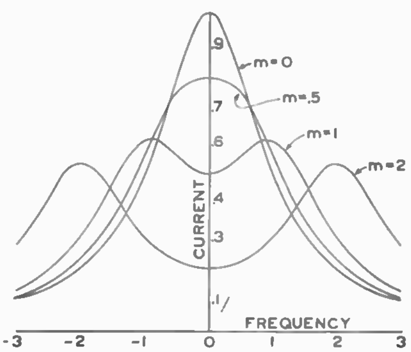


FIG. 2a

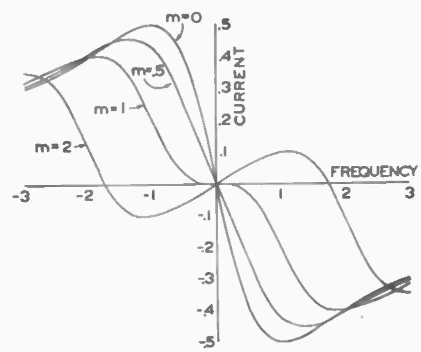


FIG. 2b

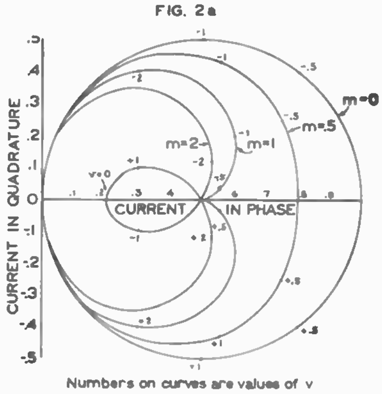


FIG. 2c

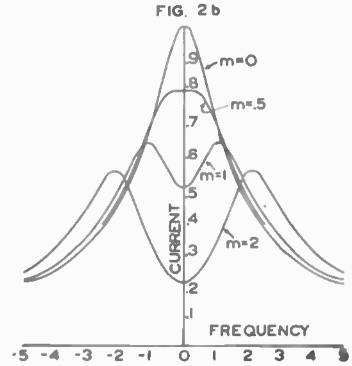


FIG. 2d

* In all figures the Frequency Scale is in terms of v and the Current Scale must be multiplied by E/R.

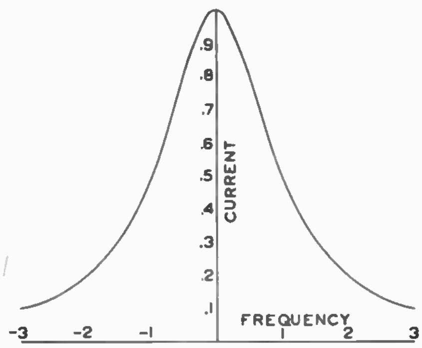


FIG. 1a

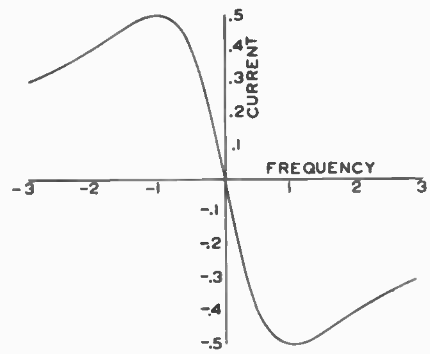


FIG. 1b

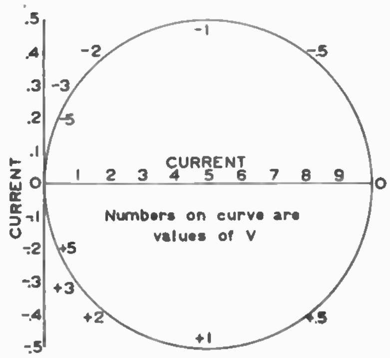


FIG. 1c

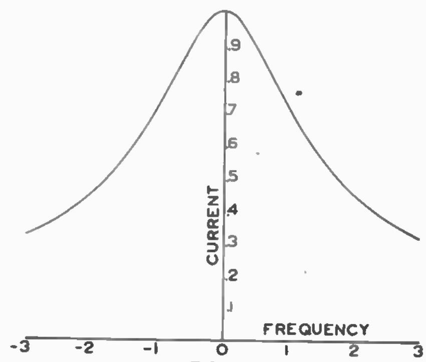


FIG. 1d

was made to simplify equation (8), throwing it into the form of equation (9), and, therefore, appears arbitrary and without any practical significance; let us see if this is so. We have already hinted that the v scale in figs. 1a, 1b, 1c and 1d may be translated into a frequency scale, and we proceed to show how this may be done. The factor $\frac{\omega L}{R}$, or the Q of the tuning coil, is known to be nearly constant over its useful frequency range as a tuning inductance, and is certainly constant enough over the small range of frequency involved in resonance to be considered constant in the definition of v. It goes without saying that the resonant frequency f_o is constant in any particular case, as it is the frequency at which the inductance and capacitance of the circuit resonate. Therefore, the only variable in the definition of v is Δf_o, so that we may say v is proportional to the number of cycles per second off resonance. In view of this, we can convert the v scale into a frequency scale which shows the frequency departure from resonance merely by substituting the value of $\frac{\omega L}{R}$ and f_o for the individual case.

For example, let $\frac{\omega L}{R} = 200$ and f_o = 1000 KC.

Substituting these values, we find that Δf_o = 2.5v KC, so that for v = 1 on the v scale we put 2.5 KC, for v = 2, 5.0 KC, etc., the minus values of v giving the frequency departure below resonance.

Before leaving the subject, we call attention to the relative unimportance of the larger values of v in describing resonance. What corresponds to our v = -1 and v = +1 have been called quadrantal values by Prof. A. E. Kennelly, the frequency values between which the phenomenon of resonance may be said to occur. These are the frequencies above and below resonance at which the reactance is numerically equal to the resistance, the phase angle of the current being minus or plus 45°, respectively. Referring to the circle diagram of fig. 1c, it will be seen that in going from v = -1 to v = +1 one half the circumference is traversed. In covering the range of v from -5 to +5 nearly 7/8 of the circumference is passed over, so that very little room is left for larger values of v. Reference to the other figures will show that the current amplitudes at these values of v are very small compared with the magnitude of the current at resonance, or v = 0. This is our justification for making the approximations employed in deriving equation (8) from (3), along with the assumption that $\frac{\omega L}{R}$ is reasonably large.

We could have avoided the approximations noted by letting v equal $\frac{\omega L}{R}\left(1 - \frac{f_o^2}{f^2}\right)$ in equation (3), but that would have necessitated a more complicated procedure in converting the v scale to a frequency scale in a particular case. We prefer not to do that because it merely obscures the problem at hand, which is to obtain a simple mental picture of the manner in which the characteristics of the simple tuned circuit are modified as a second tuned circuit is coupled to it.

THE COUPLED TUNED CIRCUIT PRIMARY:

We now propose to investigate the characteristics of a series tuned circuit when a second tuned circuit is coupled to it, that is, we shall direct our attention to what happens to the primary current as a secondary tuned circuit is coupled to it at various strengths of coupling. We shall assume, for definiteness, that the coupling is magnetic, and, to avoid too much complication, that the primary and secondary circuit constants are exactly alike. The latter assumption applies, of course, to cases where the primary and secondary constants are not alike providing the power factor and resonant frequency of the primary and secondary are equal, because referring the primary to the impedance level of the secondary, or vice versa, reduces the problem to the case where the constants are exactly alike.

* In all figures the Frequency Scale is in terms of v and the Current Scale must be multiplied by E/R.

practical purposes. This is especially true at radio frequencies where the ratio of reactance to resistance, or the Q of the coil as it is sometimes called, of a tuning coil may reasonably be 200, for example. Assuming the value of 200 it may be shown that the amplitude of the current drops to 1/5 of the amplitude at resonance when the frequency of the impressed e.m.f. deviates from the resonant frequency by slightly over one percent. Consequently, the major portion of the selectivity curve does not require consideration of frequencies which deviate more than a few percent from the resonant frequency. Confining our attention to the frequency range comprised in deviations of only a few percent from the resonant frequency is the same as saying that $\Delta f_o/f_o$ will be always very small as compared with unity. Therefore, in the right hand expression of equation (5) we may neglect $\Delta f_o^2/f_o^2$ which is much smaller still.

Then if we carry out the operation of division indicated in the expression $\dfrac{1}{1+\frac{2\Delta f_o}{f_o}}$ there results,

$$1-\frac{2\Delta f_o}{f_o}+\left(\frac{2\Delta f_o}{f_o}\right)^2-\text{etc} \text{-------}(6)$$

where $\left(\dfrac{2\Delta f_o}{f_o}\right)^2$ and all the remaining terms are, of course, also negligible. The right hand expression of equation (5) becomes by the substitution of (6),

$$\left(1-\frac{f_o^2}{f^2}\right)=\left(1-\frac{1}{1+\frac{2\Delta f_o}{f_o}}\right)=1-\left(1-\frac{2\Delta f_o}{f_o}\right)=\frac{2\Delta f_o}{f_o} \quad (7)$$

Making use of these approximations, equation (3) may be written,

$$I=\frac{E}{R}\cdot\frac{1}{1+j\frac{\omega L}{R}\left(\frac{2\Delta f_o}{f_o}\right)} \text{-------}(8)$$

Finally, if we place $\dfrac{\omega L}{R}\left(\dfrac{2\Delta f_o}{f_o}\right)$ equal to v, we obtain,

$$I=\frac{E}{R}\cdot\frac{1}{1+jv} \text{--------}(9) \; *$$

which is equation (1) greatly simplified by the foregoing approximations.

Equation (9) expresses the current as two factors of which E/R is the amplitude of the current at the resonant frequency, and $\dfrac{1}{1+jv}$ is a function of v, v being defined in the preceding paragraph. Since E/R has a fixed value in any particular case, the whole story of resonance is contained in the second factor $\dfrac{1}{1+jv}$ which does not explicitly involve the particular values of R, L and C which compose the tuned circuit. Therefore, assigning values to v and plotting various curves for the second factor provides a set of curves which is applicable to any series tuned circuit. To interpret the curves in the light of a particular tuned circuit involves merely the translation of the v scale into a frequency scale, as will be explained in a later paragraph.

It will be instructive to rationalize the denominator of $\dfrac{1}{1+jv}$ by multiplying both numerator and denominator by 1 - jv, which gives,

$$\frac{1}{1+jv}=\frac{1}{1+v^2}-j\frac{v}{1+v^2} \text{--------}(10)$$

Of this result, $\dfrac{1}{1+v^2}$ is the term in phase with, and $-j\dfrac{v}{1+v^2}$ the term in quadrature with, the impressed e.m.f. The two terms are shown plotted in figs. 1a and 1b, respectively. The real term, in phase with the impressed e.m.f., is symmetrical about v = 0, i.e., has equal values for equal positive and negative values assigned to v, and is always positive in sign. The imaginary term, in quadrature with the impressed e.m.f., has values of opposite sign but equal magnitude for equal positive and negative values assigned to v. On comparing the curves, it will be seen that for v = -1 both terms have the same magnitude and sign and for v = +1 they have the same magnitude but opposite signs, the magnitude in each case being one half the value of the real term for v = 0. The phase angle of the resultant current with respect to the impressed e.m.f. is, of course, the angle whose tangent is the imaginary term divided by the real term. For negative values of v the tangent is positive, indicating a leading current, or capacitive effect, while for positive values of v the tangent is negative, indicating a lagging current, or inductive effect. This accords, as it should, with the customary reactance diagram for a series tuned circuit, the net reactance being capacitive below resonance and inductive above.

We may plot equation (10) in still a different way. We may consider the real and imaginary terms to be the co-ordinates of a point corresponding to a value of v in which the real term is the distance along the x-axis and the imaginary term along the y-axis in locating the point. Such a plot is shown in fig. 1c and will be recognized as the familiar circle diagram. We needn't have gone to the trouble of plotting a series of points, as just stated, except, perhaps, to label the points so found with the corresponding value of v. We may determine the construction of the circle by considering equation (10) written in a slightly different form,

$$\frac{1}{1+jv}=\frac{\sqrt{1+v^2}}{1+v^2}\,\angle\tan^{-1}-v=\frac{1}{\sqrt{1+v^2}}\,\angle\tan^{-1}-v \quad (11)$$

It is easy to see that the cosine of the angle whose tangent is minus v is equal to $\dfrac{1}{\sqrt{1+v^2}}$, so that,

$$\frac{1}{1+jv}=\cos\left(\tan^{-1}v\right)\angle\tan^{-1}-v \text{------}(12)$$

This is the equation of a circle with a radius of 1/2 and center on the x-axis at a distance of 1/2 to the right of the origin, the angle designation indicating that positively increasing values of v succeed one another in a clockwise direction.

Finally, we have plotted in fig. 1d the absolute magnitudes of equation (10), i.e., $\dfrac{1}{\sqrt{1+v^2}}$, against v. This is the current response curve of a series resonant circuit, and corresponds to the characteristic which would be obtained by measurement of the current flowing in the circuit as the e.m.f. of constant amplitude is varied in frequency according to v.

We turn now to the practical question of how to interpret the variable v, which has been defined as equal to $\dfrac{\omega L}{R}\left(\dfrac{2\Delta f_o}{f_o}\right)$. This definition

* An alternative derivation leading to the same result as equation (9) may be found in Dr. A. E. Guillemin, 'Communication Networks', Vol. I, page 117 et seq.

PROCEEDINGS of the RADIO CLUB OF AMERICA

Volume 12 September, 1935 No. 2

AN ANALYSIS OF COUPLED TUNED CIRCUITS AT RADIO FREQUENCIES

BY

L. A. KELLEY*

Delivered before the Radio Club of America
September 18, 1935

An analysis of coupled tuned circuits begins properly with a detailed study of the tuned circuit uncoupled followed by consideration of the changes brought about as the second tuned circuit is coupled with it first weakly and then with gradually increasing strength. We are especially interested in the steady state characteristics involving current amplitude, phase and frequency as they are affected by the different degrees of coupling. We shall see that there are no abrupt changes in these characteristics during the process and by emphasizing this viewpoint we obtain the more satisfying mental picture of the characteristics exhibited by coupled tuned circuits which relates them closely to those shown by uncoupled tuned circuits instead of keeping them in separate watertight mental compartments.

Our plan will be to simplify the problem by making certain assumptions and approximations. We shall, for example, take the usual expressions for the current in terms of the impressed e.m.f., resistance, inductance and capacitance and convert them into expressions which do not involve the particular values of resistance, inductance and capacitance. We shall also assume in the case of coupled circuits that the power factor of the primary circuit is equal to the power factor of the secondary circuit. From these derived expressions we shall plot a series of curves, inspection of which will show clearly how the characteristics of coupled tuned circuits may be evolved from those of the uncoupled circuits. Finally, we shall give a graphical method of obtaining the same characteristics.

THE UNCOUPLED TUNED CIRCUIT:

The magnitude of the alternating current which flows in a series tuned circuit when an alternating e.m.f. is impressed is given by the expression,

$$I = \frac{E}{R + j\left(\omega L - \frac{1}{\omega C}\right)} \quad \text{--------(1)}$$

where,
E = the impressed alternating e.m.f.
R = the resistance
L = the inductance
C = the capacitance
and ω = 2π times the frequency of the impressed e.m.f.

The ωL in the reactance term in the denominator of equation (1) may be factored out and the reactance term rewritten as follows,

$$\left(\omega L - \frac{1}{\omega C}\right) = \omega L\left(1 - \frac{1}{\omega^2 LC}\right)$$

Substituting for 1/LC its value in the expression for resonance, namely, $\omega_0^2 = 1/LC$, the reactance term may again be rewritten,

$$\left(\omega L - \frac{1}{\omega C}\right) = \omega L\left(1 - \frac{\omega_0^2}{\omega^2}\right) = \omega L\left(1 - \frac{f_0^2}{f^2}\right) \quad \text{-----(2)}$$

In the last expression, f_0 is the frequency of resonance and f is any frequency which the impressed e.m.f. may have. Substituting (2) into (1) and dividing both numerator and denominator by R, equation (1) may be transformed to,

$$I = \frac{E}{R} \cdot \frac{1}{1 + j\frac{\omega L}{R}\left(1 - \frac{f_0^2}{f^2}\right)} \quad \text{--------(3)}$$

The frequency f may be expressed in terms of the resonant frequency f_0 by,

$$f = f_0 + \Delta f_0 \quad \text{-----------(4)}$$

where Δf_0 is the difference between f and f_0. Thus, if the resonant frequency is 1000 KC and the frequency in which we are interested is 1001 KC we may express it as $f_0 + 1$ KC. Substituting equation (4) in the expression $(1 - f_0^2/f^2)$ there results,

$$\left(1 - \frac{f_0^2}{f_0^2 + 2\Delta f_0 f_0 + \Delta f_0^2}\right) = \left(1 - \frac{1}{1 + \frac{2\Delta f_0}{f_0} + \frac{\Delta f_0^2}{f_0^2}}\right) \quad (5)$$

When the reactance of the tuning coil $\omega_0 L$ at the resonant frequency is very large in comparison with the effective resistance R we find that we need consider only a relatively small range of frequencies to obtain a selectivity curve which is sufficiently complete for

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Referring to fig. 2a, it will be seen that the most rapid changes accompanying an increase in m occur at and close to v = 0, the frequency of resonance of the uncoupled tuned circuit. As m goes from zero to 1/2, the peak of the m = 0 curve is flattened out while scarcely any changes occur in the legs of the curve. In going to m = 1, the rapid decrease in the amplitude at v = 0, with the legs of the curve changing but little, causes two peaks to form, one above v = 0 and one an equal amount below. Further increase in the coupling, to m = 2, causes a deeper depression at v = 0 and two peaks with wider separation than in the case of m = 1.

We can throw more light on these peaks by resorting to an explanation which is somewhat academic, but, nevertheless, affords a simple way of comprehending them. Let us suppose that we increase m to indefinitely large values and see what happens. A word of caution should be inserted here to the effect that there is an upper limit to the value of m in any practical case because the coefficient of coupling cannot be greater than 100%. However, we are dealing with radio frequency tuned circuits where the coefficient of coupling is not more than a few percent, so there is ample room for increasing m to values large enough to approximate the conditions assumed in this explanation. Now let us go back to the first expression for I_1 given in equation (16). If we place the reactance term equal to zero we find that $v^2 = m^2 - 1$ and, since we are assuming m to be large, $v^2 \doteq m^2$ or $v = \pm m$. In other words, the primary current resonates at v = +m and -m when m is large. With this hint, let us replace v by w where v = m+w, that is, so that w = 0 when v = m and w varies about m in the same manner that v varies about zero. Now let us make this change of variable in the expression $\frac{m^2}{1+v^2}$. In the first place, assuming w to have only small values as we did in the case of v,

$$v^2 = (m+w)^2 = m^2 + 2mw \quad \text{(approximately)} \qquad (17)$$

So that, $$\frac{m^2}{1+v^2} = 1 - \frac{2w}{m} \quad \text{(approximately)} \quad (18)$$

Therefore, $$1 + \frac{m^2}{1+v^2} = 2 \quad \text{(approximately)} \quad (19)$$

and,

$$v\left(1 - \frac{m^2}{1+v^2}\right) = (m+w)\left(1 - 1 + \frac{2w}{m}\right) = 2w \text{ (approx.)} \quad (20)$$

Substituting (19) and (20) into the second expression of equation (16),

$$I_1 = \frac{E}{R} \cdot \frac{2 - j2w}{4 + 4w^2} = \frac{E}{2R} \cdot \frac{1 - jw}{1 + w^2} \text{ ------ } (21)$$

We note that the second factor in (21) is exactly the same as (10) except that it is in terms of w instead of v. Upon comparing it with (9), the second factor of which is rationalized in (10), we observe that the current at resonance factor is E/R in (9) and E/2R in (21), that is, the current at resonance in (21) is 1/2 the amplitude of the current at resonance in the uncoupled tuned circuit. If we had placed v = -m + w, so that w = 0 when v = -m and w varies about -m in the same manner that v varies about zero, we should arrive at the same equation (21), and in this case it would apply to the resonance conditions about v = -m. We may conclude, therefore, that for large values of m, the primary current resonates at two frequencies, v = +m and -m, the expansion of the frequency scale about these resonant frequencies being in terms of a new variable w.

How shall we go about expanding the frequency scale in terms of the new variable w? Taking the resonant frequency v = m as an example, we see, first of all, that we ought not assume the approximations which hold only when v is small because now we assume m to be large. We go back, therefore, to equation (3) and let v = $\frac{\omega L}{R}\left(1 - \frac{f_o^2}{f^2}\right)$. Since $v = \frac{\omega L}{R}\left(1 - \frac{f_o^2}{f^2}\right) = m + w = \frac{\omega M}{R} + w$

we solve for w $$w = \frac{\omega(L-M)}{R} - \frac{\omega L}{R} \cdot \frac{f_o^2}{f^2} \text{ ------ } (22)$$

Next we replace the resonant frequency f_o with the new resonant frequency f_m which, of course, is equal to f when w = 0, or, $0 = \frac{\omega(L-M)}{R} - \frac{\omega L}{R} \cdot \frac{f_o^2}{f_m^2}$, from which, $\frac{f_o^2}{f_m^2} = \frac{L-M}{L}$. Since $\frac{f_o^2}{f^2} = \frac{f_o^2}{f_m^2} \cdot \frac{f_m^2}{f^2}$ we may substitute it in (22), which gives,

$$w = \frac{\omega(L-M)}{R}\left(1 - \frac{f_m^2}{f^2}\right) \text{ ---------- } (23)$$

Now we may apply the same method of approximation as we did with v and obtain, $\frac{\omega(L-M)}{R}\left(\frac{2\Delta f_m}{f_m}\right)$

It is clear that we are in a position to interpret w in exactly the same manner as explained for v merely by substituting f_m for f_o. The expression (L-M) in place of L simply means that the decreased inductance brings about resonance at a higher frequency.

We are now better able to interpret the trend of the curves plotted by substituting numerical values in equation (16) for successive increasing values of m. Returning to fig. 2a, for example, it is easy to see what is happening as m increases. The curves intermediate m = 0 and m equals a large magnitude represent a smooth transition from a single tuned circuit of resistance R to what is the equivalent of two separate single tuned circuits of resistance 2R, the resonant frequencies of the latter being at v = +m and -m. The m = 2 curve in fig. 2a already shows a close approach to the large m condition. The peaks occur at v = ±2 = ±m, the maximum amplitudes are slightly greater than 1/2 the maximum amplitude for m = 0, and the shape of the two resonance curves is very nearly the same as the m = 0 curve with the ordinates all reduced in proportion. Turning next to fig. 2b, which shows the curves for the second, or imaginary, term of equation (16), we observe that the curve for m = 0 crosses the horizontal axis only once, going from positive to negative at v = 0. As m increases, it continues to cross the axis, but at a less steep angle, and is accompanied by a flattening effect near v = 0. Curve m = 1 represents a boundary condition (critical coupling) where the curve flattens out sufficiently to follow along the axis for an appreciable distance on either side of v = 0. Further increase in m brings about three crossings of the horizontal axis, positive to negative at a negative value of v, negative to positive at v = 0, and finally positive to negative at a positive value of v. The important thing is that it is crossing the axis from positive to negative twice, once above and once below v = 0 and if m is made large, according to our analysis, the crossing points will then be at v = +m and -m, the curve in the neighborhood of the crossing points being exact duplicates of the m = 0 curve with all of the ordinates at half value. This tendency is clearly borne out by the m = 2 curve where the crossing points are at ±1.75, or slightly less than ±2. In view of what has been said already, the absolute magnitude curves of fig. 2d are self-explanatory.

The polar curves of fig. 2c reveal the trend from a single tuned circuit of resistance R to two separate tuned circuits of resistance 2R in a novel and striking manner. The m = 0 curve is, of course, a duplication of the circle diagram of the uncoupled tuned circuit. As m is made to increase, the circle flattens slightly in the right hand portion where the small values of v occur. The flattened part takes on a decided depression at v = 0 as m is further increased until at m = 1 the curve has assumed a distinct cardioidal shape. This is the case of critical coupling and again we note that it is a borderline case because a greater value of m is accompanied by the formation of a loop in the place of the depression. According to our analysis, this loop should grow in size along with a contraction in the remaining part of the curve until, when m is large, it becomes one of two coincident circles, the remaining portion of the curve forming the other circle, the diameter of which is exactly 1/2 the diameter of the m = 0 circle. Examination of the m = 2 curve indicates that this result is well on the way.

THE COUPLED TUNED CIRCUIT SECONDARY:

It is highly important that we should also investigate what goes on in the secondary of a coupled tuned circuit, because the load, or energy receiver, is located in the secondary circuit, and the primary purpose in using a coupled tuned circuit is to utilize the peculiar characteristics associated with the transfer of energy from a source of oscillations connected to the primary circuit to a load connected to the secondary circuit. That brings us to the main difference between the primary and secondary circuit problems, namely, that in the primary circuit, the impressed e.m.f. and the resulting current are in the same circuit, while in the secondary circuit, the impressed e.m.f. and the resulting current are in different circuits. The current flowing in the secondary circuit with an impressed e.m.f. of value E in the primary circuit is given by;

$$I_2 = \frac{j\omega ME}{\left[R+j\left(\omega L-\frac{1}{\omega C}\right)\right]^2+\omega^2M^2} \quad \text{------}(24)$$

where R, L, C, M and ω have the same significance as previously. Upon comparing this with equation (13) for the primary current we observe the following relation,

$$I_2 = \frac{j\omega M I_1}{R+j\left(\omega L-\frac{1}{\omega C}\right)} \quad \text{----------}(25)$$

Equation (25) leads to the conclusion that we may regard the secondary circuit as a single tuned circuit with an impressed voltage equal to $j\omega MI_1$. We see that the secondary is fed by the voltage developed by the primary current flowing through the mutual reactance, a potentiometer effect. Let us follow up this idea, but first let us apply to equation (25) the simplifications already developed, i.e., $R(1+jv) = R+j\left(\omega L-\frac{1}{\omega C}\right)$ and $m=\frac{\omega M}{R}$, so that,

$$I_2 = \frac{jm}{1+jv}\times I_1 = \frac{jm(1-jv)}{1+v^2}\times I_1 \quad \text{-----}(26)$$

We now follow out the multiplication indicated in (26), using the value of I_1 in equation (16), and obtain,

$$I_2 = \frac{E}{R}\cdot\frac{m\left[2v+j(1+m^2-v^2)\right]}{4v^2+(1+m^2-v^2)^2} \quad \text{---}(27)$$

If the simplifications are applied directly to equation (24), the result may be put in the same form as (27).

The curves shown in figs. 3a, 3b, 3c, and 3d are plotted from equation (27) for the same values of m and the same range of v in each case as in figs. 2a, 2b, 2c and 2d, which were plotted from equation (16) for the primary circuit. We have employed a uniform system of designating the figures in which the numbers 1 refers to a single tuned circuit, 2 to the primary of a coupled tuned circuit, and now 3 to the secondary of a coupled tuned circuit, and of the letters following the numbers, a refers to the real part of the current, b to the imaginary part, c to the polar diagram in which the magnitude versus angle is plotted, and d to the absolute magnitude, or the square root of the sum of the squares of the real and imaginary terms, plotted against v.

Before proceding with a detailed examination of the last set of curves, let us make a general analysis of the effect of giving extreme values to m. In order to make use of our previous similar discussion in connection with the primary current, let us keep in mind equation (26) which expresses the secondary current in the form of two factors, one of which is the primary current. If m = 0, the secondary current is, obviously, zero for all values of v. If, however, m is greater than zero, but still quite small, we have already seen that the primary behaves like a simple tuned circuit and (26) shows that, under this condition, the net result is the same as two simple tuned circuits in tandem, without any complicated reactions being perceptible. Referring again to the discussion of the primary case, we determined that when m is large there are two distinct frequencies of resonance at v = + m and -m, respectively, and now, on account of the relation in (26), we conclude that the same is true for the secondary current. Let us replace the variable v by w such that v = m + w, where m is large and w small, and substitute into equation (27). Neglecting quantities which are relatively small, we obtain,

$$I_2 = \frac{E}{2R}\cdot\frac{1-jw}{1+w^2} \quad \text{------}(28)$$

which is exactly the same as equation (21) for the primary current. This is the equation showing resonance about the frequency v = + m in terms of the frequency variable w. On the other hand, if we define w such that v = -m + w and substitute into (27), we find,

$$I_2 = -\frac{E}{2R}\cdot\frac{1-jw}{1+w^2} \quad \text{---------}(29)$$

which is the same as (21) and (28) with the sign reversed. It will be remembered that the primary current for this case, i.e., resonance about the frequency v = -m, is also expressed by (21) without a change in sign. This signifies that for resonance about v = + m the primary and secondary currents are equal and in phase, while for resonance about the frequency v = -m they are equal but in opposition. The conversion of w into a frequency scale about the resonant frequencies is, of course, the same as explained for the primary circuit. With these general remarks we now pass to a consideration of the plots for the secondary current.

Fig. 3a shows the real component of the secondary current as given by (27) for values of m equal to 1/2, 1 and 2. The first thing that strikes us is the resemblence between these curves and those of fig. 2b for the imaginary component of the primary current, the main difference being that where fig. 2b is positive fig. 3a is negative and vice versa. The multiplier jm in (26) accounts for this because, when the multiplication is carried out the imaginary term of the primary current is made real and the sign reversed by the product of the two j's. The resemblance applies only to the general outlines of the curves, the numerical values being substantially different. A horizontal line coinciding with the x-axis would represent the $m = 0$ curve. As m is made to increase, the characteristic shape of the curves appears, a positive peak above $v = 0$ and a negative peak below, with the crossing point at $v = 0$, the peaks increasing in magnitude and spreading away from $v = 0$ at equal and opposite values of v. Our analysis tells us that at large values of m we should approach resonance about the frequencies $v = +m$ where the primary and secondary currents are equal and in phase and $v = -m$ where the currents are equal but in opposition. This tendency is quite apparent in the trend of the curves as m takes on the values of 1/2, 1 and 2 progressively. How well advanced it is by the time $m = 2$ may be seen by comparing the $m = 2$ curves in fig. 3a and fig. 2a, where the peaks occur at $v = \pm m = \pm 2$. As predicted, they are in phase at $v = +2$ and in opposition at $v = -2$, the peak magnitudes in fig. 2a being slightly above, and in fig. 3a slightly below, the limiting value of 1/2 the value at resonance of the single tuned circuit. Notice, too, in fig. 3a the curvature developing in the line crossing at $v = 0$ as m increases, accomodating its shape to that which it must assume at large values of m, the same as fig. 1a at half amplitude.

The curves of fig. 3b, showing the imaginary component of the secondary current as given by equation (27), have a suggestion of the appearance of those of fig. 2a, the reason being clear from (26). Here, as before, a horizontal line coinciding with the x-axis would correspond to the $m = 0$ curve. With increasing m a positive peak rises up at $v = 0$, see $m = 1/2$ curve, reaching a maximum value of 1/2 for $m = 1$ (critical coupling), then receding back towards zero again for m larger than unity, the legs spreading out all the while and crossing to negative values at values of v approaching $+m$ and $-m$. Eventually, as m is still further increased, the righthand portion of the curve becomes the same as fig. 1b for the single tuned circuit at half amplitude, the crossing being at $v = +m$, while the lefthand portion becomes the negative of fig. 1b at half amplitude, the crossing being at $v = -m$.

The absolute magnitude curves plotted from (27) in fig. 3d show the tendency to be expected from the consideration given the curves of figs. 3a and 3b. They also show the $m = 1$ curve in relation to curves in which m is less than or greater than unity, namely, $m = 1/2$ or 2. The $m = 1$ (critical coupling) curve is of great importance in the practical use of coupled tuned circuits because the current amplitude is nearly constant for an appreciable range of frequencies above and below $v = 0$. It is the borderline case between a curve having a single maximum, as in the $m = 1/2$ curve, and a curve having two maxima, as in the $m = 2$ curve. Once the two maxima appear, the peak values remain constant at 1/2. This is to be compared with the double peaks which develop in fig. 2d as m increases, where the peak values are greater than 1/2 when the peaks first appear, decreasing asmptotically to 1/2, and, eventually, coinciding with the curves of fig. 3d at large values of m.

Now let us examine the polar curves in fig. 3c. A point at the intersection of the coordinate axes is sufficient to represent the case of $m = 0$. We notice, first of all, that the axis of symmetry of these curves is the vertical axis, while it was the horizontal axis for fig. 2c. As m is assigned increasing values, the curves take on a cardioidal shape which is reminiscent of the $m = 1$ curve of fig. 2c. Indeed, the $m = 1$ curves of figs. 2c and 3c have exactly the same dimensions, so that if the $m = 1$ curve of fig. 3c were rotated in the plane of the paper 90° counterclockwise with the center at the origin of axes, then slid bodily to the right along the horizontal axis a distance of 1/2, and finally rotated 180° about the horizontal axis, it would then coincide in every respect with the $m = 1$ curve of fig. 2c. After the cardioidal curve expands to a maximum vertical distance of 1/2 for $m = 1$, a depression

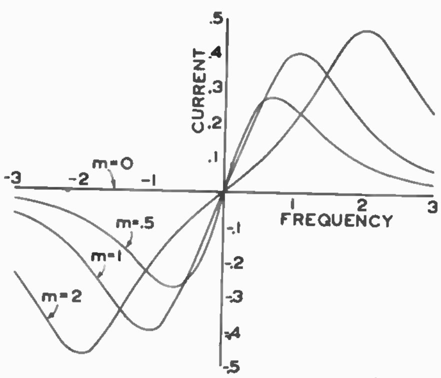


FIG. 3a

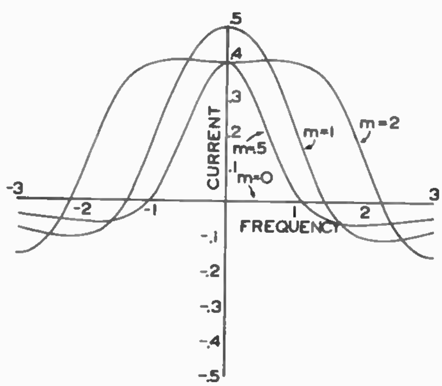
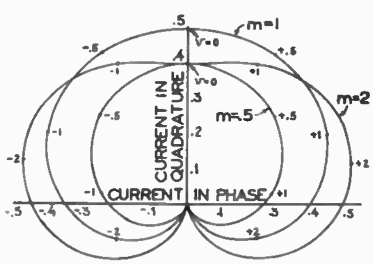


FIG. 3b



Numbers on curves are values of v

FIG. 3c

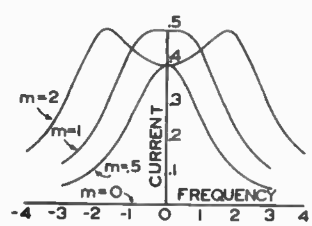


FIG. 3d *

* In all figures the Frequency Scale is in terms of v and the Current Scale must be multiplied by E/R.

appears about the vertical axis where v = 0 which deepens with further increase in m until, finally, at large values of m the curve evolves into two tangent circles of equal diameter, the vertical axis forming the tangent to them at the origin of axes, the horizontal axis from 0 to +1/2 and from 0 to -1/2 being their diameters. The righthand circle coincides with the coincident circles of fig. 2c for large m. One of the primary circles is for resonance about v = +m and should be paired with the righthand secondary circle which is also for resonance about v = +m, and with which it coincides exactly, the other primary circle is for resonance about v = -m and should be paired with the lefthand secondary circle with which it is in opposition.

GRAPHICAL METHOD FOR CONSTRUCTING THE POLAR DIAGRAMS:

As we have seen, the polar diagrams, figs. 1c, 2c and 3c, contain all the information provided by the remaining figures, providing the values of v are marked along the polar curves. By drawing straight lines connecting the origin with successive values of v along the polar curves, the horizontal projections of these lines with the corresponding values of v are all the data we need to plot the (a) curves, the vertical projections and the lengths of the lines against v for the (b) and (d) curves respectively. Therefore, let us see if we cannot construct the polar curves graphically.

The circle diagram for the single resonance circuit is so well known that we shall proceed directly to the problem of the coupled tuned circuit. For this purpose, it will be convenient to alter the form of equation (15) by multiplying the numerator and denominator by $1+jv$, giving the result,

$$I_1 = \frac{E}{R}\cdot\frac{1+jv}{(1+jv)^2+m^2} \quad \text{-(30)}$$

Noting that $m^2 = (-jm)^2$ we can factor the denominator of the second factor.

$$I_1 = \frac{E}{R}\cdot\frac{1+jv}{[1+j(v-m)][1+j(v+m)]} \quad \text{----(31)}$$

Now expand the second factor into partial fractions,

$$\frac{1+jv}{[1+j(v-m)][1+j(v+m)]} = \frac{A}{1+j(v-m)} + \frac{B}{1+j(v+m)} \quad (32)$$

in which A and B are constants as yet undetermined. In order to determine them, clear (32) of fractions.

$$1+jv = A[1+j(v+m)] + B[1+j(v-m)] \quad (33)$$

Next equate the coefficients of the j terms,

$$v = A(v+m) + B(v-m) \quad \text{--------(34)}$$

If we let m = v in (34), we find that A = 1/2, and, similarly, if we let m = -v, B = 1/2. Consequently,

$$I_1 = \frac{E}{2R}\cdot\left[\frac{1}{1+j(v-m)} + \frac{1}{1+j(v+m)}\right] \quad \text{-(35)}$$

We shall now derive a similar expression for I_2. Using the first expression in (26), and the value of I_1 given in (31), we obtain the following expression for I_2,

$$I_2 = \frac{E}{R}\cdot\frac{jm}{[1+j(v-m)][1+j(v+m)]} \quad \text{------(36)}$$

Expanding the second factor into partial fractions,

$$\frac{jm}{[1+j(v-m)][1+j(v+m)]} = \frac{C}{1+j(v-m)} + \frac{D}{1+j(v+m)} \quad (37)$$

Where C and D are constants to be determined. Clear (37) of fractions and equate the j terms,

$$m = C(v+m) + D(v-m) \quad \text{--------(38)}$$

When v = m, C is found to be 1/2, and when v = -m, D is found to be -1/2. Therefore,

$$I_2 = \frac{E}{2R}\cdot\left[\frac{1}{1+j(v-m)} - \frac{1}{1+j(v+m)}\right] \quad \text{------(39)}$$

We may consolidate (35) and (39) into one expression,

$$I_1, I_2 = \frac{E}{2R}\cdot\left[\frac{1}{1+j(v-m)} \pm \frac{1}{1+j(v+m)}\right] \quad (40)$$

in which the + sign is used for I_1 and the - sign for I_2.

Equation (40) suggests the graphical method because each of the terms in the brackets may be represented by a circle diagram, the circles of which coincide, the scales being different. Referring to fig. 4, the construction consists

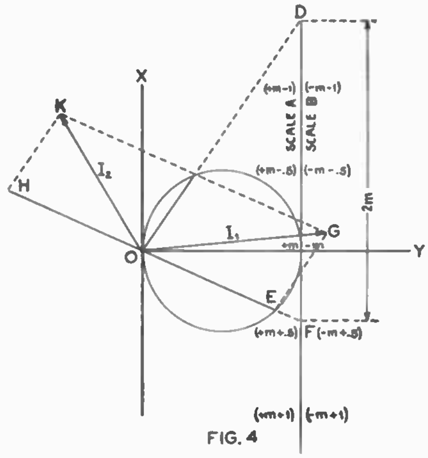


FIG. 4

first in laying out a circle of radius 1/2 with the center on the x-axis at a distance of 1/2 to the right of the origin. The line for the scales of v is next drawn perpendicular to the x-axis at a distance of 1 to the right of the origin, i.e., tangent to the circle. For convenience, let us call the first term in the brackets A and the second term B. As a starting point in marking the A scale, we observe that the phase angle of A is zero when v = +m, so we mark the scale where it touches the extremity of the diameter with the value of +m. Then we mark off the A scale in both directions from the starting point, the units divisions of which upward from the starting point are (m-1), (m-2), etc., and downward from the starting point, (m+1), (m+2), etc., each unit being the length of the diameter of the circle. We find, similarly, that the starting point for the B scale should be labeled with v equal to -m, the units divisions upwards from this starting point being marked with (-m-1), (-m-2), etc., and downward with (-m+1), (-m+2), etc., the length of each unit being, as before, the length of the diameter. The point on the A scale corresponding to a value of v = 0 is (m-m), or m units above the starting point, while on the B scale v = 0 corresponds to (-m+m), or m units below the starting point on the B scale. In going from v = 0 to v = +1, we move down one unit on each scale, and so forth, so that the distance between points on the A and B scales for corresponding values of v remains constant for a given value of m and equal to 2m, the point on the A scale always being above that on the B scale.

To find I_1, assuming the scales to be laid out for a given value of m, locate points D and F, fig. 4, corresponding to the chosen value of v on the A and B scales, respectively. Connect O and D by a straight line intersecting the circumference of the circle at C. The line OC is the vector representing the first term in the brackets of equation (40), its magnitude being the length of the line in units of the diameter, and its phase angle being the angle between OC and OX. Connect O and F by a straight line intersecting the circumference at E. The line OE is the vector representing the second term in the brackets of (40). Finally, add the vectors OC and OE by completing the parallelogram of which they are adjacent sides, and the diagonal OG is the vector of I_1. If, now, we wish to find the corresponding vector for I_2, we merely substract the vector OE from the vector OC by extending OE back through O to H such that OH is equal in magnitude to OE, and then add the vectors OH and OD getting vector OK for I_2. Repeating this for a whole range of values of v we obtain a series of vectors for I_1 and I_2. Drawing a curve through the extremities of the primary current vectors results in a curve such as those in fig. 2c, and a similar procedure for the secondary current vectors results in a curve such as those in fig. 3c.

It is instructive to consider our reasoning about extreme values of m in the light of this graphical construction. When m = 0, for example, points D and F coincide, and, likewise, vectors OC and OE, so that the curve traced out by the extremities of I_1 is another circle of twice the diameter of the circle OCE, or a single tuned circuit with the current at resonance equal to E/R. Since the vectors to be subtracted to obtain I_2 are always equal under the condition m = 0, I_2 is equal to zero for all values of v. This is also apparent, of course, by substituting m = 0 into equation (40). On the other hand, when we come to look at the case where m is large we see that points D and F are widely separated for corresponding values of v. Imagine D and F sliding down from large negative values of v, keeping a constant but large distance apart. Line OF will be the first to cut the circle to any great extent, and it will have cut nearly all the way around the circumference before line OD will begin to cut into the circle appreciably. Thus, we have the effect of two single resonant circuits resonating at the frequencies v = -m and +m, the amplitude of the current at resonance being E/2R. Note that OF is the first to traverse the circle, of which the chord OE is the second vector in the brackets of (40). Now, this is positive for I_1 and negative for I_2, so that the primary and secondary currents are in opposition when they resonate at v = -m. On the other hand, when OD traverses the circle after OF has practically finished, the chord OC is the vector representing the first term in the brackets of (40) and this is seen to be positive for both I_1 and I_2. Therefore, the primary currents are in phase when they resonate at v = +m. All of the foregoing checks in an interesting way our previous speculations on extreme values given to m.

CONCLUSION:

We have chosen to emphasize the behavior of coupled tuned circuits in terms of the less complex phenomena of simple series tuned circuits. In this view, the coupling is regarded as increasing gradually from zero to large values accompanied, as we have described, in the primary, by a smooth transition from simple resonance at one frequency to simple resonance at two widely separated frequencies, the current amplitude at resonance for each being half the amplitude of the former, and, in the secondary, from zero current at all frequencies to simple resonance at the same two widely separated frequencies, the current amplitude at resonance for each being the same as for the primary, in phase at one resonant frequency and opposed at the other. The curves in our figures are snapshots on the way. Most practical importance is attached to a limited range in the midst of the transition as, for instance, at and near critical coupling, where certain features of the current versus frequency curve are found desirable, but it is interesting to think of this as an intermediate stage in the merging of two cases of simple resonance.

MEETING NOTES

The meeting held on June 13, 1935 at Columbia University was devoted to papers prepared by Messrs. D.E. Harnett, M.P. Case, and Walter Lyons of the Hazeltine Service Corporation, and demonstrations of two new developments in broadcast radio receiver design, by these gentlemen. The first of these to be shown was their automatic bass compensation system whereby an auxiliary audio-frequency amplifier effective only in the lower register, is added to the conventional radio receiver, with such automatic control as makes it effective only at the lower output levels, and, in fact, automatically makes it increasingly effective as the output level is lowered.

Operational demonstrations of any kind are invariably difficult in a large meeting room such as is used by the Club; and in this case the fact that the arrangements were effective only at low receiver output levels made demonstrations unusually difficult; notwithstanding all of which, the demonstration of the equipment was thoroughly indicative of its interesting characteristics. It was especially interesting to note how the tonal picture of the music remained little unchanged as the general level was lowered or raised, notwithstanding the human ear's lack of sensitivity both to the lower and extreme upper registers.

The second paper concerned itself with the variable selectivity superheterodyne which was demonstrated. This system comprises an intermediate frequency-amplifying system in which the frequency band width is controllable by means of variation in the inter-circuit couplings. This demonstration also was highly successful, in that a great change in fidelity possible through the variation of I F selectivity, as well as possibilities in the direction of avoiding adjacent channel interference, were readily evident.

Radio Club of America Celebrates 25th Anniversary

Dedicated to the "spirit of good fellowship and the free interchange of ideas among all radio enthusiasts" the Radio Club of America, Inc., has issued a special Year Book in commemoration of its 25th anniversary. Among the names appearing in the roster of present and past leaders of this pioneer club (said to be the oldest radio club in the world) are to be found many long since prominent in the world of radio communication as it is known today, and several now appearing on the roster of active members of the A.I.E.E.

Quoting from opening paragraphs of the Year Book:

"The story of the Radio Club of America begins over a quarter of a century ago, during the really dark ages of the radio art, about 1907. . . . Here we find a group of small boys who, according to the true American spirit, were so interested in flying that they formed the Junior Aero Club of U.S.

"In conjunction with their experiments in aviation, these youngsters had, for some time, also been interested in what was then known as wireless. In fact, the new idea of sending messages without wires had proved itself so fascinating, that they found themselves actually devoting most of their spare time to tinkering with wireless apparatus. There were at this time a small number of so-called amateur wireless experimenters in and about New York City, so the boys decided to form a new club with wireless as an object."

"Accordingly, . . . a special meeting of the Aero Club, for the purpose of forming a new club, with wireless telegraphy and telephony as its main interest. . . was held at the Hotel Ansonia in New York City on January 2, 1909. . . . Thus, the Junior Wireless Club Limited was founded," and bore that name until October 21, 1911, when it was changed to the Radio Club of America in recognition of its expanding membership and interests.

The Year Book presents a comprehensive outline of the history of the club, lists the major contributions of its members to communication literature, and includes a roster of members and of past and present officers. Copies of the Year Book are said to be available at the club's executive headquarters, 11 West 42nd Street, New York, N. Y.

ELECTRICAL ENGINEERING

25 Years of Radio

JANUARY 2, 1909 the Junior Aero Club of U. S. met at the Hotel Ansonia at the instance of W. E. D. Stokes, Jr. to consider a new hobby, radio. The members of this group, now the Radio Club of America, were of gentle age—Mr. Stokes was then 12. Today these boys are grown but their interest in wireless survives. The history of their club is the history of radio in America.

In the Year Book will be found photos of Louis Pacent and Harry Sadenwater listening for signals from Europe, Armstrong's regenerative apparatus, radio stations of E. V. Amy and others, Harry Houck's home-made loose coupler, station 1BCG which pumped 200 meter signals to Europe and caused M. I. Pupin and David Sarnoff to go to Greenwich, Connecticut, to see what "the boys are doing."

The entire book brings back memories of the old and glamorous days. Evidently the committee, under George Burghard, is still amateur at heart. In the Book is a history of each of the several hundred members.

ELECTRONICS

Twenty-fifth Anniversary Year Book, Radio Club of America, Incorporated. The oldest radio club in the world was 25 years old in 1934. In commemoration of the event, the club has issued the Anniversary Year Book. The greater part of this book is devoted to a history of the club and its members, richly illustrated with reproductions of letters, newspaper clippings, pictures of apparatus and their constructors.

The first meeting was held on January 2, 1909 in the Hotel Ansonia of New York City. At that meeting there were five members and the Club was then called the "Junior Wireless Club, Limited." The president, Mr. W. E. D. Stokes, Jr., was 14 years old and had made his own wireless set. Apparently, the entrance requirements were that you had to make your own set.

By 1911, the members had become sufficiently numerous to warrant issuing a typewritten list. At this time the name of the Club was changed to "Radio Club of America." The Club fought for the rights of amateurs and succeeded in preventing an unfavorable bill from being passed in Washington.

Many members of the Club have become important figures in the radio industry, contributing many new ideas. A list of papers read before the Club and a list of members are included in the book. The membership is now 320.

RADIO NEWS

"RADIO CLUB" SILVER ANNIVERSARY YEAR BOOK

IT GIVES US a great deal of pleasure to be able to review the Twenty-Fifth Anniversary Year Book of the Radio Club of America, Inc. Dedicated to "The Spirit of Good Fellowship and the Free Interchange of Ideas Among All Radio Enthusiasts," this book contains a wealth of information and inspiration in its many pages. In fact, a comprehensive review would, we feel, require nearly as many pages as this year book contains; and so this review shall consist of only a very brief summary of a few of the many highlights incorporated in its 85 pages.

The preface, which outlines the spirit that has contributed so much to the growth and prestige of the Radio Club of America from the eight charter members of the Junior Wireless Club Limited to its present 320, was written by George Eltz, Jr., who it happens was one of those charter members.

"A History of the Radio Club of America, Inc.," by George E. Burghard, traces in a very interesting and vivid manner the story of the Radio Club from its beginning in 1909. Mr. Burghard has condensed his material into some 38 well-illustrated pages.

Immediately following is a foreword by Lawrence C. F. Horle. This foreword precedes the complete listing of the Proceedings of the Radio Club. To appreciate the value of this listing of the Proceedings, even if only for reference purposes, one needs but glance at names of the engineers who have presented papers.

An enrollment of past officers, condensed history of each member, and the Constitution of the organization completes this year book.

The members of the Twenty-Fifth Anniversary Year Book Committee, namely, George E. Burghard, Ernest V. Amy, Edwin H. Armstrong, George J. Eltz, Jr., John F. Grinan, Lawrence C. F. Horle, Frank King, Robert H. Marriott, Fred Muller, Joseph J. Stantley, and W. E. D. Stokes, Jr., deserve a vote of thanks from the Radio Club.

RADIO ENGINEERING

A few copies still available

25TH ANNIVERSARY YEAR BOOK

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Essential to all Libraries.

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11 West 42nd Street,
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Proceedings

of the

Radio Club of America

Incorporated

R C A

October, 1935

Volume 12, No. 3

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11 West 42nd Street + + New York City

The Radio Club of America, Inc.

11 West 42nd Street - New York City

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PROCEEDINGS
of the
RADIO CLUB OF AMERICA

Volume 12 October, 1935 No. 3

THE CATHODE RAY TUBE IN TELEVISION RECEPTION

BY
I. G. MALOFF*

Delivered before the Rodio Club of Americo
September 18, 1935

The discoveries of the electron and of electron emission and the development of means of controlling the electron emission, opened a wide way which led to the present day radio broadcasting, sound recording and communication systems in general. In exactly the same way the gradual development of means for concentrating electron beams and especially the development of means for controlling these beams in intensity and direction, are directly responsible for the present day television systems of high definition.

An electron beam is a narrow pencil of negatively charged particles moving with great velocities of the order of 30,000 miles per second. While electron beams were discovered and used as early as the end of the last century, it is only in the last decade that their properties were understood, and means for their generation and control were developed.

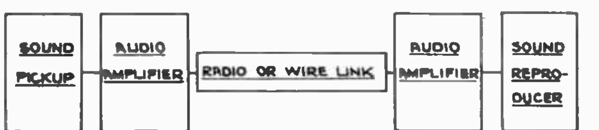
A television system of high definition, just as any television system, must have several component parts.

Fig. 1a shows a block diagram of a practical television system.

It has been shown time after time that in order to transmit a picture over wires or radio it is necessary to split the picture having two dimensions,

FIG.1.

-A-



-B-

*Research Division, R. C. A. Mfg. Co., Inc., Camden, N. J.

namely, height and width, into one dimension of length. In other words, we have to scan a picture in order to transmit it, and moreover we have to scan synchronously the picture at the receiver and to vary the intensity of the spot at the receiver in synchronism with the variation of brightness under the scanning spot at the transmitter.

The only difference between television systems of high and of low definition is in the degree of detail transmitted and received. The old mechanical systems are satisfactory for televizing pictures having definitions up to 100 lines and reach their limit of definition somewhere between 100 and 200 lines. High definition systems employing electron beams have hardly a limit to their capability of definition. The limitations on these are imposed chiefly by the associated communication apparatus.

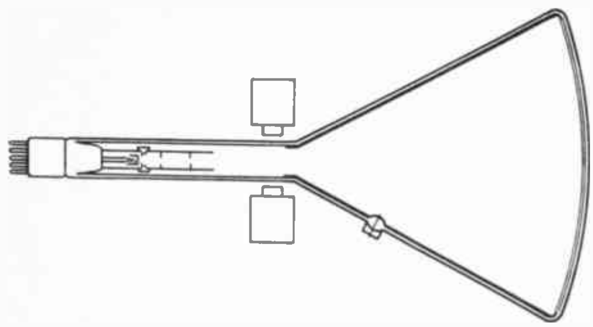
On Fig. 1b a block diagram of a complete sound communication system is shown. The only essential difference between the two systems is the fact that a television system requires synchronizing arrangements while a sound system does not. Outside of that, the similarity is striking.

True, the amplifiers have to pass frequency bands of differing widths, but in all essentials the systems are alike. Both systems pick up forces affecting our senses and translate these forces into electrical intensity vs. time variations; both systems pick them up at low levels and have to use vacuum tube amplifiers in order to bring them to levels suitable for transmission over distance; both systems may or may not use carriers to transmit the intelligence over the distance by means of either the air or cable. On the receiving end, both systems have to demodulate and amplify the received signals to bring them to the required power levels and then, by means of suitable transducers, reproduce the forces similar to those which were originally picked up.

The microphone in the sound system corresponds to the "Iconoscope" in television, and the power output stage and the loud speaker similarly correspond to the cathode ray tube with its electron gun and fluorescent screen.

While both the "Iconoscope" and the cathode ray receiving tube, or the "Kinescope" employ electron beams, the structure, operation and characteristics of the two are widely different. In this discussion we will deal with the "Kinescope" or the

FIG. 2



cathode ray television receiving tube, with its construction, characteristics and operation.

THE "KINESCOPE"

A typical "Kinescope" is shown on Fig. 2.

It consists essentially of five component parts: first, a glass envelope, sealed for maintenance of high vacuum; second, a cathode from which the cathode rays or, utilizing the modern term, the electrons are emitted; third, a device for concentrating, controlling and focusing of the electron beam; fourth, an arrangement (either internal or external) for deflecting the beam; and fifth, a fluorescent screen on which the received image is reproduced.

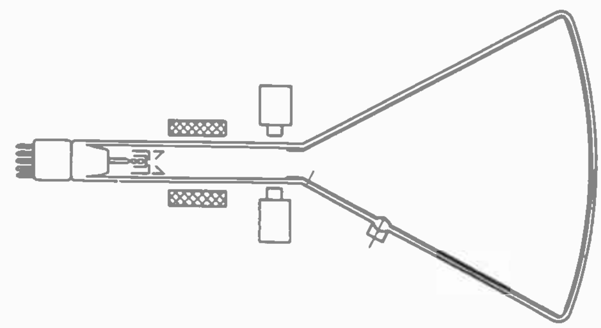
The envelope is usually made of glass, strong electrically and mechanically, and is designed to withstand atmospheric pressure with a large margin of safety. The tube during its processing is painstakingly baked and outgassed to maintain a very high vacuum of the order of 10^{-7} cm. of mercury. The early tubes were partially gas filled and utilized for beam concentration the space charge resulting from collisions of electrons with gas molecules. This so-called gas focusing is rather uncertain, since the gas pressure in any tube varies with life. If in a tube the focusing of the beam is critically dependent on pressure, the useful life of such a tube is short. Most of the modern cathode ray receiving tubes are of the high vacuum type where the concentration of the electron beam is accomplished by means of electric or magnetic fields produced between electrodes and poles. The vacuum treatment of the modern tubes is such that even at the end of the tube's life, that is, when the cathode emission has fallen off, the vacuum is still sufficiently high so that collisions between electrons and molecules of the residual gas seldom occur.

The cathode is of a tubular form with a flat emitting surface covered with a preparation of barium oxide. Only the flat end, facing the fluorescent screen, is covered with the electron emitting material. A tungsten heater, non-inductively wound and insulated with a heat resisting material, is located inside of the tubular cathode.

The electron gun, or the device for concentrating, controlling and focusing of the electron beam, consists of a grid sleeve, and a first anode. Sometimes it includes another electrode, usually called screen grid, not essential for operation and not shown on the figure. The grid sleeve is of a tubular form with a disc parallel to the flat emitting surface of the cathode. A circular hole in the center of the disc is coaxial with the cathode sleeve. The first anode cylinder coaxial with the rest of the system is usually mounted by means of insulators on the grid sleeve. It carries diaphragms or aperture discs on the inside for stopping or limiting the beam angle and for limiting the penetration of electrostatic fields. The glass envelope of the "Kinescope" carries a black conductive coating on its inner side and has a sealed-in conductor leading to this coating. The conductive coating forms the last or the second anode. The final electron accelerating potential is applied between the cathode and the second anode.

The purpose of the first anode is to stop the beam similarly to an optical stop in a lens and to create an axially symmetric electrostatic field which would start the initially divergent electrons of the beams towards the axis. By adjusting the voltage on the first anode the fluorescent spot on the screen can be brought to a minimum diameter. The voltage on the first anode for best focus is usually about 1/4 or 1/5 of that on the second anode.

FIG. 3.



An electromagnetic system of focusing is shown on Fig. 3. A short first anode and the second anode are connected together to the source of the final accelerating potential and the concentrating field is produced by a multilayer solenoid coaxial with the tube and the gun.

The fluorescent screen is placed on the inner side of the front face of the tube. This face is usually as flat as mechanical strength permits. The material is usually either a sulphide or a silicate of zinc and is deposited in a very thin translucent layer.

OPERATION OF THE "KINESCOPE"

The operation of the tube is as follows: The electrostatic field created by the potential applied to the first anode penetrates the grid opening and draws the emitted electrons into a well defined beam. The grid is usually at a somewhat lower potential than the cathode and in this way limits the beam intensity. By applying a more negative potential to the grid the beam can be completely cut off. After entering the first anode, the beam passes through a masking diaphragm which cuts off some of the irregular peripheral portion of the beam. Then the beam enters the region of the field produced by the difference of potentials between the first and second anodes. In this field a strong focusing action takes place, which gives the electrons a radial component of velocity directed toward the axis of symmetry of the beam. The radial momentum acquired by the electrons is sufficient to bring them, after a flight through the equipotential space of the main body of the tube, to a focus at the screen.

Some of the electrons originally drawn from the cathode are cut off by the masking aperture. They return to the emf source through the first anode lead. The rest of the electrons strike the fluorescent screen. They excite the screen and dissipate most of their kinetic energy there. This kinetic energy has been acquired by the electrons through acceleration from the very small velocities of emission to that corresponding to the second anode voltage.

Some of the energy of the beam is transformed into light, some goes into heat raising the temperature of the glass, while the rest is spent in knocking out secondary electrons from the screen material. These low velocity secondaries flow in a steady stream to the conductive coating of the second anode. An equilibrium condition is quickly established, the conductive coating acquiring the highest possible potential with respect to the cathode while the fluorescent screen slides a few score of volts below the potential of the coating. The difference of potentials between the fluorescent screen and the conductive coating is such that it draws the secondaries to the coating at exactly the same rate as the primaries are arriving to the screen.

Just as soon as the beam leaves the first anode it is subjected to the action of either magnetic or electric fields for the purpose of deflecting the beam in a predetermined manner. This deflection is scanning in television.

There are three ways of scanning, namely: first, by means of two electrostatic fields at right angles to each other; second, with two electromagnetic fields also at right angles to each other; and third, with an electrostatic and an electromagnetic field parallel to each other. It may be remembered at this point that the electrostatic field deflects electrons along the lines of force and that an electromagnetic field deflects them perpendicular to the lines of force.

The electrostatic deflecting plates are usually placed inside of the glass envelope while the electromagnetic deflecting coils are invariably outside the envelope.

SCANNING REQUIREMENTS

Since most of the tube characteristics can be observed and measured only while actually scanning, we will take up the scanning means first. The purpose and principle of scanning have been treated at great length by many writers, so we will limit ourselves to the recollection of scanning requirements in a high definition television system.

In our organization we have made a thorough study of this subject and the following are some of the conclusions reached.*

If we qualify and limit the ability to tell a desired story to specific conditions, the experience we have had with television allows us to make some interesting approximate generalizations. If we take as a standard the information and entertainment capabilities of sixteen-millimeter home movie film and equipment, we may estimate the television images in comparison.

60 scanning lines	entirely inadequate
120 " "	hardly passable
180 " "	minimum acceptable
240 " "	satisfactory
360 " "	excellent
480 " "	equivalent for practical conditions.

This comparison assumes advanced stages of development for each of the line structures.

We may say therefore that a number of scanning lines in the immediate vicinity of 360, say 340, will give a very good performance comparable with 16 mm home movie film.

In motion pictures the taking, or the camera frame frequency determines how well the system will reproduce objects in motion. This has been standardized at 24 frames per second. In television it is assumed that we shall use a frame frequency of 24 per second or greater. Since this is satisfactory for motion pictures, it is also satisfactory for television and this characteristic of frame frequency will, therefore, not be considered further.

In the reproduced image there is another effect of frame frequency. This is the effect of frame frequency on flicker. Motion picture projectors commonly used are of the intermittent type. The usual cycle of such a projector is that, at the end of each projection period, the projection light is cut off by a "light cutter", the film is then moved and stopped so that the succeeding frame registers with the picture aperture; the light cutter then opens, starting the next projection period. This is repeated for each frame - 24 per second. Since projection at 24 light stoppages per second with illumination levels used in motion pictures causes too great a flicker effect, the light is also cut off at the middle of the projection period for each frame for a time equivalent to the period that it is cut off while the film is moved from one frame to the next. This results in projection at 24 frames per second with 48 equal and equally spaced light impulses. Such an arrangement provides a satisfactory condition as regards flicker. In television we also may have a reproduced image at 24 frames per second, but because of the manner in which the image is reconstructed, a continuous scanning process, it is not practicable further to break up the light impulses by means of a light chopper in a manner similar to that used in the projection of motion pictures. We, therefore, have for the usual systems of television a flicker frequency which corresponds with the actual frame frequency (24 per second, for example). This is satisfactory at very low levels of illumination but becomes increasingly objectionable as the illumination is increased.

It has been concluded that, in a television system, satisfactory flicker conditions exist if each frame consists of two groups of alternate lines and that there should be 24 or above of the complete double frames. This so-called interlaced scanning is equivalent to 48 or above frames per second as far as flicker is concerned.

At 48 equivalent frames and with 60 cycles power system, the effects of hum or ripple travel across the reproduced image. The choice of 60 equivalent frames or 30 interlaced frames per second provides a complete solution to the visual requirements, i.e., motion, flicker and ripple.

ACCESSORY CIRCUITS

If we have to lay out a circuit to receive the television picture we have to provide: first, a suitable "Kinescope" with suitable power supply; second, a deflecting yoke for deflecting the beam vertically and horizontally at, say, the just-mentioned frame and line frequencies; third, for driving this yoke synchronously with the incoming signal, and fourth, an electric circuit to drive the grid of the "Kinescope" to provide the gradations of brightness on the screen.

The last item, of course, includes circuits for demodulating, amplifying, etc. of the incoming signal.

So far we have been discussing elementary generalities. Let us now consider somewhat in detail the question of the accessory circuits and of the arrangements required by the output device of a television receiver.

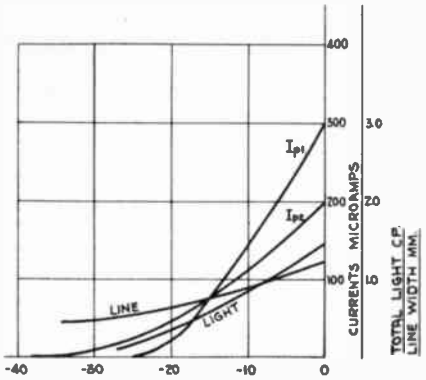
The first item is: a suitable "Kinescope" with a suitable power supply.

A typical set of characteristic curves of a "Kinescope" are given on Figure 4. It is very similar to that of a four-electrode tube. The second anode current I_{p2} is shown as a function of grid bias; the current on the first or the auxiliary anode is also shown. But there are two quantities on the characteristic which do not appear on that of an ordinary vacuum tube. They are line width vs. grid voltage and total light vs. grid voltage.

The line width is measured by means of a microscope with a scale focused on an isolated scanning line, when a pattern is scanned at normal scanning rates. The vertical deflection is increased until the adjacent lines separate by a centimeter or so. The total light is measured by means of a suitable illuminometer. The brightness units are rather con-

*See papers by E. W. Engstrom, IRE Proc. Vol. 22, page 1241, Nov. 1934, and IRE Proc. Vol. 23, page 295, April, 1935.

FIG.4.



fusing, just as all the light units. It is not hard to remember, however, that a perfectly diffusing area of 1 square foot and having brightness of one foot candle emits 1 lumen of luminous flux. One lumen of luminous flux from a flat surface is generated by a source whose intensity is $1/\pi$ candle power.

The total light in most of the modern tubes is approximately proportional to the beam wattage or beam current at a constant beam voltage, the coefficient of proportionality being between 1 and 2 c.p. per watt. The tube, the characteristics of which are shown on Fig. 4, has a screen efficiency of 1.5 c.p. per watt.

Now let us see what kind of picture we can produce on this "Kinescope". Suppose it has a screen large enough to accommodate a 6" x 8" picture. For 340 lines it should have a line width of 6/340 = .018", or roughly .5 mm.

Experience, however, has taught us that the lines can be somewhat larger than this theoretical value because the intensity of the luminous spot falls rather rapidly as we go away from the center of it. A reasonable correction factor is about 1.6 times the theoretical value. This means that we may tolerate a line width up to .8 mm. From the characteristic curve we take the corresponding value of beam current: 86 microamperes at -14 volts on the grid. The cutoff we find to be -40 volts. This means that a grid swing of 26 volts peak to peak will drive this tube from black to maximum permissible high light.

From the same set of curves we find that the total light for this high light will be about .75 c.p. Now if the picture is 6" x 8" and we have an entirely white picture, we will get .75 c.p. of total light from 48 square inches. This amounts to 2.36 lumens from 1/3 of a square foot, equivalent to 7.1 foot candles. So the maximum brightness of the high lights in the received picture will be 7.1 foot candles. Now, remembering that the brightness of a picture in home movies is of the order of 10 to 20 foot candles, we may conclude that we will have a picture of a brightness comparable with that of home motion pictures.

The useful information which we obtained from the characteristic curve can be summarized as follows:

The "Kinescope" under consideration will produce a picture of detail corresponding to 340 scanning lines and 30 interlaced frames with approximately 7 foot candles in high light and a grid swing of 26 volts peak to peak. The power supply will have to provide 6000 and 1200 volts and have a sufficient regulation for 80 microamperes. The adjustable or automatic bias supply should go down to -40 volts.

DEFLECTING SYSTEM REQUIREMENTS

Four factors are important when considering a particular arrangement for deflecting or scanning. First, the system must require not more than a reasonable amount of power for a full size pattern; second, the luminous spot must maintain its size and shape when deflected to the edges of the pattern; third, the pattern must not deviate from its normal rectangular shape; and fourth, the system must be capable of giving a high enough ratio of the picture to return sweep. The properties corresponding to these requirements are:

Deflection sensitivity.
Freedom from defocusing of the luminous spot.
Freedom from distortion of the pattern.
High enough overall frequency response.

The above requirements apply to any system of deflection, but the mechanics of magnetic and electrostatic deflection differ greatly. Let us consider magnetic deflection.

The magnetic field for deflecting electron beams is produced by combinations of coils and poles which are often called the magnetic deflecting circuit. Since to supply power to such a combination an electrical network or circuit is required, the latter also has been called the magnetic deflecting circuit. It sounds reasonable to call the whole combination "magnetic deflecting system and its component parts; magnetic driving circuit and magnetic deflecting yoke".

While the magnetic field in which we are interested is formed in air or vacuum rather, the magnetic deflecting yoke often contains iron for the purpose of confining the field and reducing reluctances of return paths, thereby reducing the total energy stored in the field for a given deflecting effect.

A magnetic field gives an electron an acceleration at right angles to the direction in which it travels. Since it is always at right angles the electron cannot change its speed and can change only the direction in which it travels. The kinetic energy of an electron moving in a magnetic field is a constant quantity and therefore the radius of curvature "R" of the orbit can be calculated from the law of conservation of energy. It comes out as

$$R = \frac{mv}{eH}$$

where e, m, v and H are charge, mass and velocity of the electron and H is the intensity of the magnetic field. Reduced to practical units this expression becomes:

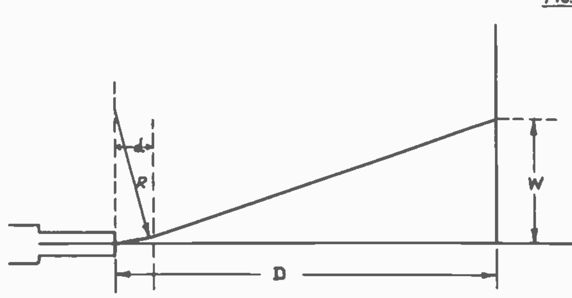
$$R = 3.36\,\frac{\sqrt{V}}{H}$$

where R is in cm., V is in practical volts, and H is in Gauss or in Gilberts per cm. Referring to Fig. 5, let D be the gun screen distance, d be the length of the magnetic field. The magnitude of the deflection W comes out as follows:

$$W = R + \frac{Dd - R^2}{\sqrt{R^2 - d^2}}$$

If d is small compared to D we get: $W \doteq Dd/R$

FIG.5



For a magnetic yoke of increasing length (d), with the inductance kept at a constant value by corresponding reduction in the number of turns, the current required for a given deflection is proportional to the square root of the reciprocal of the length of the magnetic yoke. The power required for a given deflection and also the energy stored in the magnetic field comes out inversely proportional to the length of the deflecting yoke. This means that if we can deflect a given beam by means of two power tubes, doubling the length of the deflecting yoke will require the use of only one tube to accomplish the same result.

A measure of the sensitivity of a particular magnetic deflecting system is the amount of total energy Σ stored in the magnetic field for a given full deflection, from one edge of the tube to the opposite edge.

$$\Sigma = \frac{L_o I^2}{2}$$

Here Σ is in joules, L_o is in henries and I is in amperes. If the picture is repeated n times per second and after each picture sweep this energy is dissipated, then the output tube should be capable of delivering $n\Sigma$ watts to the yoke. This value, however, is not as important as the product in voltamperes of the voltage across the yoke during picture sweep by the maximum current amplitude thru the yoke.

The ability of the power tube to supply a deflecting yoke has been treated in detail in one of the earlier papers (1932) by the author, and will not be repeated here.

DEFECTS OF THE SCANNING PATTERN

There are two main forms of defects of the scanning pattern on the screen of cathode ray tubes. The first is defocusing of the luminous spot, and the second is the distortion of the scanning pattern. By defocusing of the luminous spot is meant the change of the size of the spot when deflected. By distortion of the scanning pattern is meant the deviation of the pattern from its normal rectangular shape.

The degree to which the above defects may be present in a particular deflecting system is determined primarily by the shapes and types of the deflecting fields. There are two more common defects caused more or less by the deflecting circuit as a whole. They are: non-uniform distribution of the scanning pattern or non-linearity of the sweep, and the crosstalk between the vertical and horizontal circuits. For the first of these, the wave shape of the magnetic driving circuit and the frequency response of the yoke are responsible. For the second, either the coupling between corresponding driving circuits or the coupling between the fields of the yoke may be the cause.

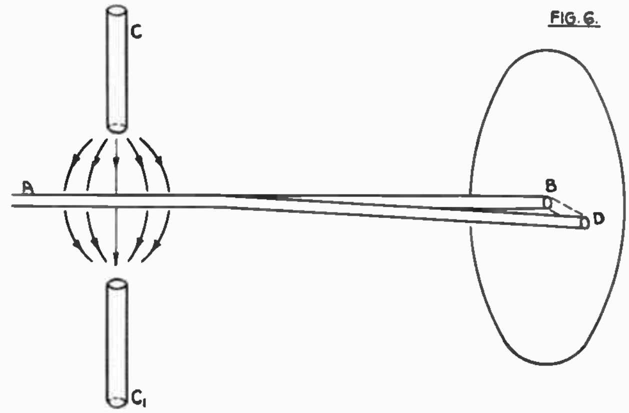
Both the static and the magnetic deflecting systems are subject to the defects enumerated above, and the work on improving both types has been in progress for several years. The early high definition systems in this country employed magnetic deflection both ways, early foreign systems showed preference for the electrostatic both ways. At present most of the systems used in this country utilize either a combination of static magnetic deflection or the all-magnetic systems.

The combined system provided only a partial solution, however. The main source of trouble in such a combination is the defocusing of the spot by the electrostatic field. A certain small amount of similar defocusing shows itself even in the best modern magnetic deflecting systems. The old magnetic systems had an exceedingly large amount of defocusing. All-magnetic systems seem best from the viewpoint of defocusing difficulties. Most of what follows refers to all-magnetic deflecting system.

DEFOCUSING OF THE LUMINOUS SPOT

Magnetic defocusing is caused by two factors: first, for a given non-uniform magnetic field it is a function of the diameter of the beam while it is under the action of the field, and second, for a given cathode ray tube it is a function of the non-uniformity of the field in the direction of deflection. The mechanism of defocusing will be better understood by considering Fig. 6. Take an

FIG. 6.



electron beam of a circular cross-section with electrons moving parallel to each other. Such a beam before it is deflected will produce a luminous spot B on the screen. This spot will be of a circular shape. Now let us deflect the beam to one side of the screen by means of a magnetic field produced by electromagnets: C and C_1. Following the right-handed screw rule the beam will be so deflected that the spot will shift to D. The magnetic field produced by the two coaxial bar magnets will be of a barrel shape form and will be the densest in the middle. The cylindrical electron beam had initial direction toward the center of this field, but when deflected it will miss the axis. The side of the beam which is closest to the axis will be deflected more. The side directly opposite will be deflected less. The spot will be compressed along the direction of deflection. It can be shown mathematically that any non-uniform magnetic field possesses a certain curvature, which is a function of the non-uniformity.

FIG. 7

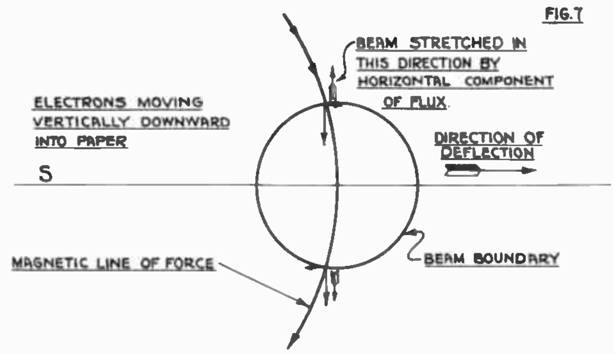


Fig. 7 shows a beam of cylindrical shape being deflected away from the center of a barrel shape field. Away from the plane of symmetry of the field, the curvature of the field results in a component of the field parallel to the plane of symmetry. These components, however, have opposite directions on the opposite sides of the plane of symmetry. In the case shown the upper and the lower parts of the beam will be stretched away from the plane of symmetry in opposite directions. This will change the shape of the spot from a circle to that of an ellipse with a major axis perpendicular to the direction of deflection. Therefore we may conclude that the non-uniformity of the field and the curvature of it both act to change the luminous spot into an ellipse with its major axis perpendicular to the direction of deflection. But this will hold only if the direction of deflection is away from the region of the field where it is most concentrated.

When a cylindrical beam is pulled into a field towards the region where it is more concentrated, the beam is stretched into an ellipse with its major axis parallel to the direction of deflection. We may look at the effects of non-uniform fields from another angle. A non-uniform field affects a cylindrical beam as a divergent cylindrical lens.

For deflection towards weaker regions of the field, the axis of this lens is parallel to the plane in which the direction of the deflection lies. For deflection toward stronger regions of the field, the axis is perpendicular to this plane, and the larger the beam diameter the larger the effect of a given field.

So far we considered only the cylindrical beams. In practice we always have converging beams, which are either focused, or underfocused, or overfocused. It can be shown by reasoning similar to that just given that if a field stretches an overfocused beam in a particular direction, a readjusting of the focusing field to give an underfocused condition will stretch the spot in a direction perpendicular to the former.

DISTORTION OF THE SCANNING PATTERN

By distortion of the scanning pattern is meant the deviation of the pattern from its normal rectangular shape. When all the four corners are pulled away farther than they should be, we get a pincushioned pattern and when these corners are not pulled far enough we get a barrel pattern.

Distortion as well as defocusing is caused by the non-uniformity and the curvature of deflecting fields. A combination of two magnetic deflecting fields, each of which is of barrel shape distribution, causes a pincushion pattern. A combination of two pincushion fields produces a barrel shape pattern. The reason for these effects can be better understood by considering Fig. 8.

FIG 8.

Fig. 8A shows how the components of two pincushion fields add together and give a comparatively small resultant for corner deflection and a barrel shape pattern. Similarly the components of two barrel shape fields add together as shown on Fig. 8B and give a comparatively large resultant and a pincushion pattern.

OVERALL FREQUENCY RESPONSE

To reproduce a saw tooth wave shape the magnetic deflecting yoke should be capable of responding to many harmonics of the saw tooth frequency. Other ways of obtaining the same result have been suggested but so far have not proven sufficiently advantageous to warrant a treatment here. For an infinite ratio of picture to return sweep the coefficients of successive harmonics are inversely proportional to the order of the harmonic. If the amplitude of the fundamental is 1, the second harmonic comes out as a half, and the third harmonic as a third, and the tenth as a tenth. Meaning that the tenth harmonic is of an amplitude equal to ten per cent of the fundamental. Now this is sort of high. Let us figure it out: 340 lines and thirty frames - this makes 10,200 lines or sweeps or cycles of the fundamental per second. This means that the tenth harmonic has a frequency of 102 kilocycles and contributes ten per cent to the wave. Fortunately we synchronize the picture every frame and every line. For positive synchronizing we have to take about 10 per cent of the time. This permits us to have, say, a ten to one ratio. Now for a nine to one ratio (which is easier to compute than the 10:1 case) of the saw tooth wave, if the amplitude of the fundamental is 1, the amplitude of the second harmonic comes out as .495, the third .300, the fourth .187, the fifth .131, and the tenth comes out negligible. So we may add to the requirements of a deflecting system that it must be capable of responding to a frequency band extending from the fundamental of the saw tooth frequency to its tenth harmonic.

CROSS TALK

Frequently in a deflecting system, a serious cross talk takes place between the horizontal and vertical circuits. Usually it is the horizontal impulse which finds its way into the vertical deflecting circuit and produces wavy zigzag scanning lines instead of straight lines. It may be caused by coupling of some sort between the driving circuits. This kind of cross talk is usually eliminated by electrically isolating and shielding the respective circuits. Often, however, it takes place because of either electrostatic or magnetic coupling between the coils of the deflecting yoke. The type and degree of coupling is usually definitely connected with electric, magnetic and physical arrangements peculiar to this particular type. It cannot be treated, therefore, in general, and has to be studied individually with every particular type of deflecting system.

As a rule, however, the cross talk can be eliminated by so arranging the coils on the yoke that the undesired induced voltages and currents buck each other out. Sometimes it calls for connecting horizontal coils in parallel and vertical in series. In other cases, both should be connected in parallel, while in some, no cross talk is produced under any conditions.

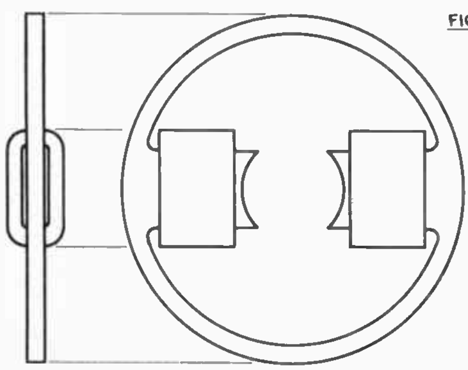
IRREGULAR DEFECTS

In our discussion of defects of the scanning pattern, we considered so far only the perfectly symmetrical yoke and a centrally located electron beam. If, however, for any reason either the beam is not centrally located with respect to the yoke, or the magnetic return legs of the yoke are not symmetrical, or the coils are not symmetrically located, the irregular defects of the scanning pattern result. If the deflecting field is sufficiently uniform, the position of the beam with respect to the yoke is not as critical as in the case of a non-uniform field.

Any non-symmetry in the yoke, however, ruins the uniformity of the field and immediately shows itself by producing defocusing in a part of the picture, stretching a corner or a side of the pattern and usually producing serious cross talk. The symptoms of the irregular defects are such that they are easily located and eliminated by tracing defective coils and by checking the geometry of the yoke and the cathode ray tube.

In conclusion let us consider a deflecting yoke of the type shown in Fig. 9. Two such yokes sufficiently spaced give a very good pattern for a 340-line 30 interlaced frame picture. It is balanced to give a very uniform field along the directions of deflection. Along the beam it naturally gives a wall of flux, so to speak, and a wall of uniform height.

FIG. 9.



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PROBLEMS OF ALL-WAVE NOISE-REDUCING-ANTENNA DESIGN

BY

JULIUS G. ACEVES*

Delivered before the Radio Club of America
November 14, 1935

The radio antenna is the gateway through which come the signals to the radio receiver. When radio was in its infancy, problems of antenna design were given detailed consideration; early text books were replete with the mathematics of the theory of antennas; much experimentation was done in the determination of their characteristics. Then came, in rapid succession, a series of inventions of apparatus and circuits culminating in radio receivers of almost unbelievable sensitivity, while at the same time, however, the antenna itself both figuratively and literally was relegated to the attic. Meanwhile, other electrical apparatus performing their usual function, continue also in their production of high frequency energy radiated through space or conducted along power lines.

This has resulted in a veritable bedlam of noise whenever stations other than the most powerful locals were to be received and interference reduction has become a necessity. As most radio receivers possess more amplification than can be used without undue noise, it follows that noise reduction, even at the expense of a little signal strength, is essential to the operation of the receiver at full capacity.

Antenna design underwent another radical change with the advent of popular short wave broadcasting, especially from foreign countries. Whereas, a simple Marconi Type antenna was sufficient for the 500-1500 K.C. band, it is inadequate for the 6-20 M.C. band particularly from the interference standpoint, and the Hertz type dipole or doublet largely superseded the simple Marconi antenna for short waves.

TYPES OF INTERFERENCE:

The purpose of this paper is to present some facts concerning the nature of interference, the remedies available, and some suggestions concerning the antenna and transmission line for best performance, not from purely theoretical consideration but also from the practical viewpoint.

It is a common experience with radio engineers and service men that what is good in one place for interference elimination is bad in another. Many engineers have found certain schemes that promised unqualified success in the laboratory, but as soon as they are put in practice they fail miserably. In fact, most of us have to confess failure at some time or other in the development of anti-noise circuits and apparatus. This has been due mainly to the fact that interference has been assumed to be just one sort of an animal to be hunted down and annihilated while, in reality, there is whole species to be exterminated.

The most deceiving circumstance is the multiplicity of ways through which interference enters the radio receiver. Accordingly, it may come in - (1) By the antenna; (2) by the downlead; (3) by the power line; (4) by the ground connection or by a combination of all these.

(1) The only remedy is to place the antenna as far above the roof as possible and some times at right angles from the source of the strongest interference. Obviously, the most suitable type of antenna for the wave lengths to be received should be selected, and experience has shown that a horizontal wire about 60 or 70 feet long with a gap in the middle so that it may constitute a doublet, as well as T antenna in conjunction with a transmission line downlead, is the most practical form of all-wave antenna from the standpoint of simplicity and facility of erection. Improvements upon this may be secured by the use of "X" or "V" doublets, multiple doublets, and other various forms of dipoles well known to the radio art.

(2) The downlead is subject to electromagnetic induction just the same as an antenna. There are two ways of preventing its effects - by shielding it and by substituting for the downlead a balanced transmission line. For practical reasons the second expedient is to be preferred for short wave reception.

The electrostatic coupling is by far the strong-

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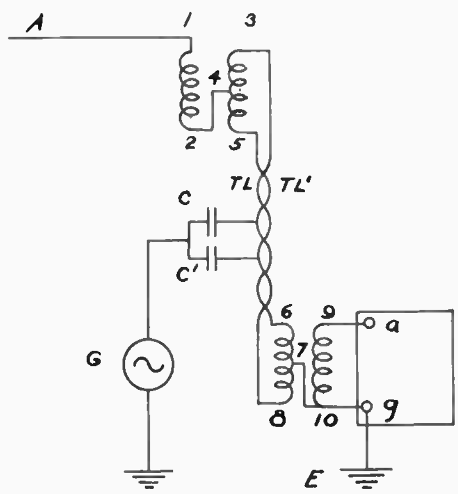

Fig. 1

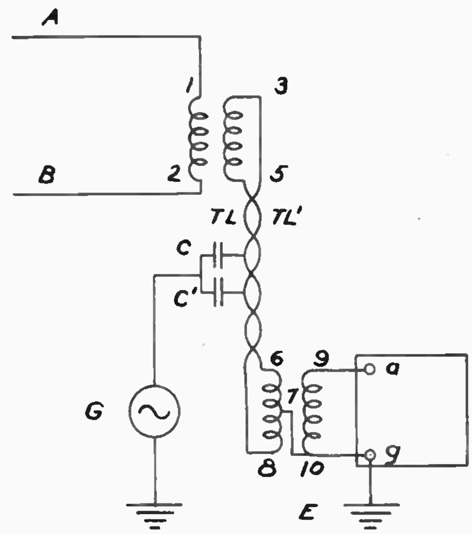

Fig. 2

est of the two components. Schematically, the circuits for signal and interference may be represented as in Fig. 1, in which we substituted the open downlead by a balanced transmission line, with two transformers at each end. We may represent the circuit of the interference by two equal capacities from the source G to the two wires of a twisted pair transmission line. Following the circuit from G through C, C' to the lines TL, TL', and entering at opposite ends of the center tapped coil 6, 7, 8, the current will go to earth from the center 7, to E and return to the generator G.

If we make coil 6, 7, 8 so that there is unity coupling between the two halves 6-7 and 7-8, and that their mutual inductance to the secondary coil 9, 10 absolutely equal, no E.M.F. will be induced therein by the passage of currents through the circuit G,C and C', L and L', 6-7 and 8-7 to E, while the signaling current from the antenna A through primary 1-2 will induce a secondary E.M.F. across the transmission line TL-TL' which will produce a current flowing down one wire and up in the other at the same time, and which we shall call "circulating current". It will enter the primary at 6 and leave at 8, inducing an E.M.F. across 9-10 which feeds the radio set terminals a-g.

If nothing else happened, there would be a complete elimination of the interference from capacity coupling to the downlead. Unfortunately, current from the generator G through capacities C-C' will not only flow to earth but will also flow upwards towards the antenna. In so doing it will traverse the primary 2-1 which will thereby induce an E.M.F. across the transmission line L-L' and force circulating currents just in the same manner as the signal in going to earth through the primary 1-2. This explains why the interference is not totally eliminated, unless the antenna impedance was infinite. It is fortunate, however, that the impedance of the branch from C-C' to earth is much smaller than the reflected impedance of the antenna, and this explains the success of the devices in the market based upon this principle of operation.

For a more complete reduction of interference, the complete severance of the metallic connections between the transmission line and the antenna coupling transformer is necessary. In Fig. 2, the signal voltage is derived from the difference of potential between two wires of a different effective height, that is, between an antenna A and a counterpoise B. The latter may be substituted in some cases by a ground connection, if such is available, or from a large metallic surface so that it will act as an effective ground rather than the surface of the earth. It must be free from other currents traveling downwards towards earth.

In the event that the wave length of the signal to be received is not much longer than twice the length of the horizontal wire, the upper coupling unit may be dispensed with, and the two wires of the transmission line connected to the center of the wire where a gap is opened, thereby converting it into a Hertz doublet. In this case all the considerations concerning the paths of signal and interference currents just discussed for long waves will still hold true, except that currents traveling upwards from the capacity coupled interference source cannot bring back circulating currents. A method of combining in one horizontal wire (with a gap) the Marconi or "T" antenna and the Hertz or doublet and a transmission line will be described later.

(3) Interference introduced by the power supply line. Next to the lead-in pickup, the power line brings in the greatest amount of interference. To understand the modus operandi, let us consider the circuit of Fig. 3 which shows schematically a source of radio frequency voltage G' across the impedance Z' of the line to earth. Measurements show that this impedance is extremely low at 60 cycles but at broadcast frequencies it is one or more hundreds of ohms depending upon frequency, distance to earth, etc. The impedance of the ground return of the radio receiver is not negligibly small either; let it be represented by Z in Fig. 3. It is obvious that the electromotive force of the source G' will send a current through the radio receiver (directly like in A.C.-D.C. sets, or indirectly by capacity coupling in other sets) and return to earth through the impedance of the ground connection Z which is common to the antenna pickup circuit made of the antenna proper, A. the effective capacity of the aerial to

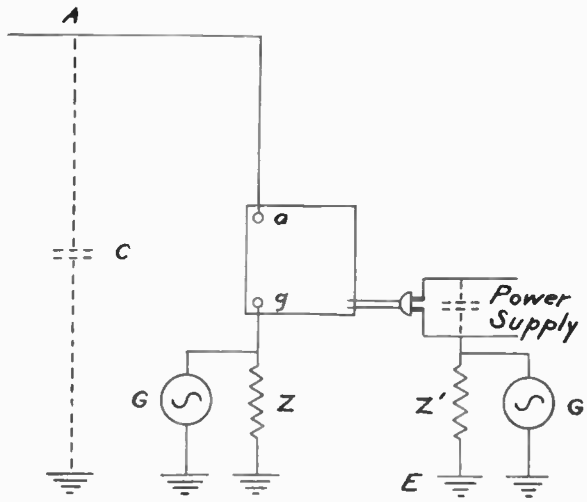

Fig. 3

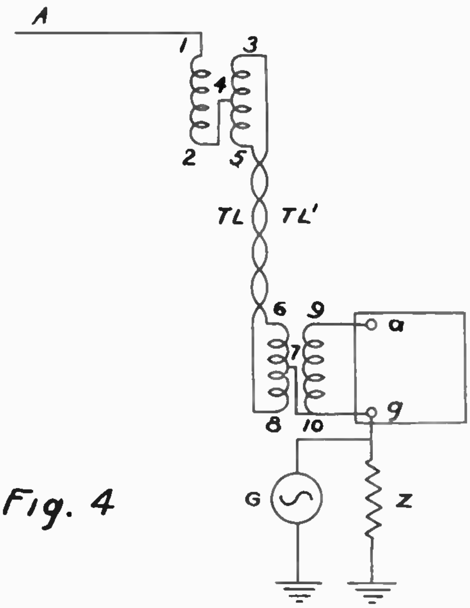

Fig. 4

ground, C, and earth, E. This constitutes two circuits with a common impedance, Z. If the distance from the radio set to earth is not very small, return currents to earth from some other sources of radio frequency disturbance such as G will also introduce an interference in the antenna pickup circuit. The existence of voltages in the ground return circuits, such as steam or water pipes, fire escapes and other conductors can be demonstrated by the common practice in many apartment houses to connect the antenna post of the radio receiver to such conductors leaving the ground post free, and obtaining sufficient radio frequency voltage to receive local as well as moderately distant stations.

The interference elimination circuit of Fig. 1 will be found ineffective in totally reducing this kind of interference. In Fig. 4 we can see how the interference currents will follow an upward path to the antenna. It will be apparent at once that the source G will force a current upwards which eventually will pass through coil 1-2 and into the antenna. Circulating currents will be immediately sent up and down the transmission line and the interference will appear as voltage across a-g. Hence, this system will not eliminate altogether interference of this type. We say altogether because the impedance matching of the source of interference is not as good as that of the antenna to the system, and hence there will be discrimination in favor of the signal. There is, however, a very simple way of stopping the passage of interference currents upwards into the antenna, and it is the insertion of an "isolation" transformer, as shown by windings 11, 12, 13, 14, in Fig. 5. It is of 1/1 ratio designed for about 100 ohms input and output impedances, or for some other value of the line impedance. The windings should have as nearly 100% coupling as is possible so as to avoid appreciable insertion losses, while the capacity between windings should be kept very low to prevent substantial flow of current from the source G up the transmission line and into the antenna.

Another way of producing the same results would be to break the ground connection between the primary center tap and the chassis, that is, between 7 and 10 in Figs. 1, 2 and 4, and eliminate the capacity coupling by means of a Faraday screen or other form of shielding. Unfortunately, shielded transformers are not very efficient because the presence of the shield introduces magnetic leakage and distributed capacity. Another serious trouble arises from unequal distributed capacity between the shield and the two ends of the primary winding which would result in an unbalanced flux for parallel currents. This effect is best illustrated by the schematic diagram of Fig. 6, where the distributed capacity of the primary 6-8 to the shield S is represented by condensers C, C'. Assuming that C and C' are unequal, it will be noted that more current flows in the first turns of the winding 6-8 to ground through C than through C' (if C is larger than C'), and hence there will be more ampere-turns in one direction than in the opposite, and therefore, there will be a resultant field inducing voltage in the secondary 9-10.

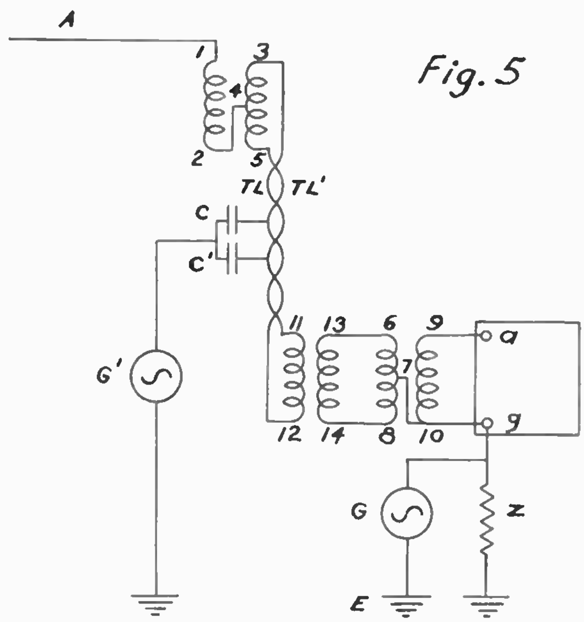

Fig. 5

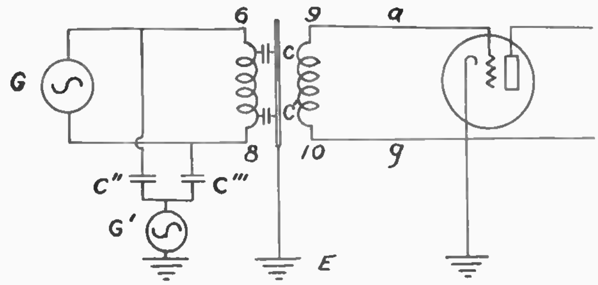

Fig. 6

The recent progress in the development of ferromagnetic substances suitable for cores of R.F. transformers is going to simplify R.F. transformer design by providing a higher mutual inductance with a minimum of turns and distributed capacity, and it is possible that a good isolation transformer may be built with a shield between windings that will introduce a small attenuation for signalling currents.

Another expedient to prevent the passage of currents from the ground connections upwards consist in isolating the set from ground as far as possible, and to effectively do so it is necessary to introduce R.F. chokes in the power line leads and not to connect the chassis to earth. After all, the only voltage that eventually produces an appreciable sound in the loud speaker is only the difference of potential between antenna and ground terminals of the radio receiver, and the chassis of the set may be considered as ground as far as reception is concerned. In practice, it is very hard to predict which is the best way to operate a radio receiver; grounded or not grounded, and if so, to what metallic bodies. It has been found that connection of the chassis to the "B.X." that encloses the power lines in one instance was the only remedy to stop severe interference from a neon sign in the same building, while in some other cases complete isolation was best. When an antenna is not in a completely "dead" field for interference, it has been possible to balance out induced interference in the antenna by allowing the downlead to pick up interference but in opposite polarity. At any given frequency it has been possible in the Laboratory to balance out the interference as completely as it is done in making measurements in a Wheatstone Bridge. In Fig. 7 is shown an arrangement where-

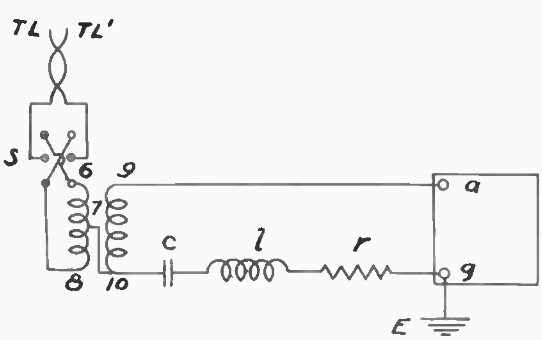

Fig. 7

by the E.M.F. across the a-g posts of the set is made of the vectorial sum of the induced secondary voltage across 9-10 and the drop across the impedance l-r-c made of three variable units so that the magnitude as well as the phase is under control. In order to make it possible to balance out any E.M.F. from the transmission line wires TL, TL', a reversal of these lines may be necessary so that the secondary voltage always may be opposed by the drop across l-r-c. In practice, a very satisfactory partial balance may be secured by the use of resistance only in lieu of the l-r-c combination, and there are coupling units in the market made with this arrangement.

PERFORMANCE ON SHORT WAVES:

When the antenna has such length that it is possible to use the horizontal wire as a dipole, the problem is considerably simplified. The signal voltage is obtained by virtue of a difference in phase of the electromagnetic wave at the various points in the wire which tends to send a circulating current into the transmission line without the necessity of introducing transformers such as those shown in Figs. 1 to 5. Interference traveling upwards along the transmission line cannot bring back circulating cur-

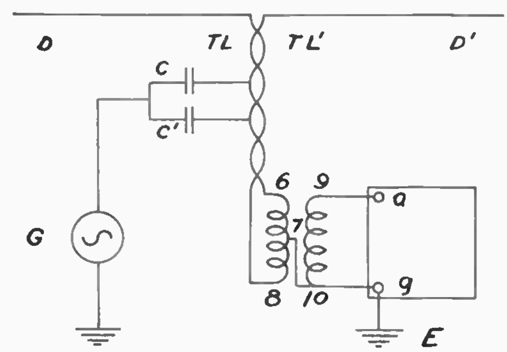

Fig. 8

rents, and a simple transformer such as shown in Fig. 8 will be found quite satisfactory. Connection between 7 and 10 is necessary to provide neutralization of capacitative coupling between the primary winding and the "live" end of the secondary. The connection from 7 to either 10 or the g post of the radio set must be very short, as at high frequencies the inductance of the wire will introduce a considerable reactance between these points, the drop across which will produce a voltage in series with the induced secondary voltage across the secondary 9-10 and interference will thusly be reintroduced into the radio set. If the connection 7-10 is broken, a shield should be interposed between windings as in Fig. 6, but with some extra attenuation to signal currents for the reasons discussed above in connection with broadcast frequency reception.

COUPLING UNITS FOR ALL-WAVE RECEPTION:

If it was possible to design a transformer that would cover the whole frequency spectrum, the lower coupling unit would be nothing else than a single center-tapped transformer like in Fig. 1.

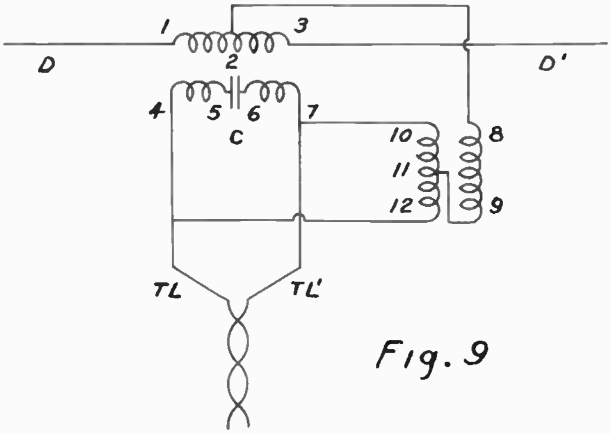

Fig. 9

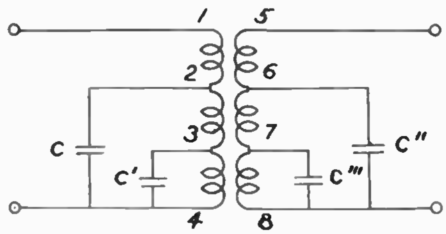

Fig. 11

The antenna coupler, however, has to be more complicated because it is necessary to produce circulating currents at low frequencies and at the same time transfer the energy from the dipole at high frequencies without interference from each other's operation.

In Fig. 9 it is shown a magnetically coupled R.F. unit consisting of the primary 1-2-3 between the two branches of the dipole, and a split secondary 4-5, 6-7 across the transmission line TL, TL'. It includes a condenser C that has low reactance for high frequencies but high for low ones. The low frequency transformer is similar to the one shown in Fig. 3 and the primary winding starts from the center of the primary winding of the H.F. unit and terminates in the mid-tap of the secondary of the low frequency unit, this latter winding being connected across the transmission line.

In Fig. 10 is shown capacitative coupling between the dipole and the transmission line. The frequency selection is accomplished by suitable choice of the electrical constants of the circuits.

At frequencies bordering the low and high frequency bands both transformers will act to some extent and the transition is very gradual, and care must be taken so that they never may act in opposition to each other by using the correct polarities. Anti-resonant circuits may be inserted in series with the two transformers tuned to frequencies in the border line, so as to make the change-over more definite and avoid losses due to phase opposition when both transformers are operating at low efficiency on some frequency near the limit.

The set coupler may be made of a number of transformers with condensers in series of multiple, as for example in the diagram of Fig. 11. In practice, however, only two units are found to be required for ordinary all-wave reception. Fig. 12 shows a unit consisting of a low frequency transformer with its primary center tapped at 7 and grounded to the chassis of the radio set g, the secondary grounded at 10. The high frequency transformer has condensers, C and C' inserted in series with the primary winding 11-12-13. The secondary 14-15 is connected in series with a small condenser and in parallel with the other secondary; both of them feed the input circuit a-g of the set.

INTERFERENCE ELIMINATION TESTS:

There are many coupling units in the market intended to be noise reducing outfits. What criterion and by what means can we ascertain their merits?

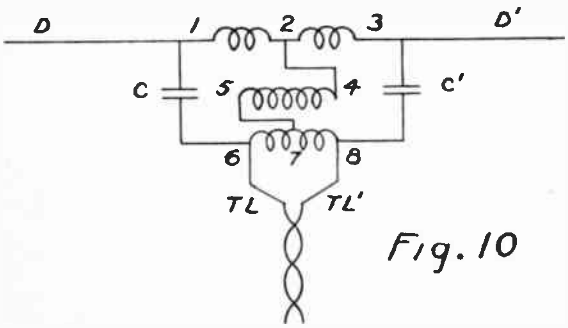

Fig. 10

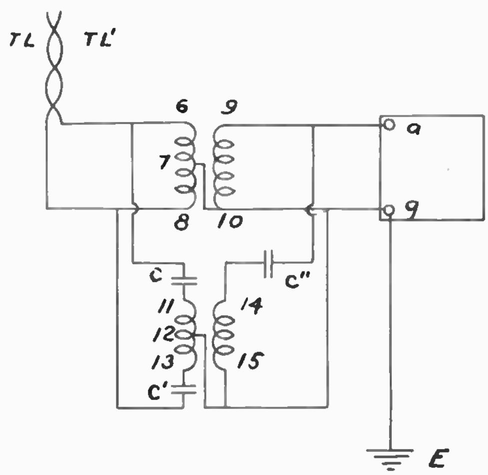

Fig. 12

At present there is great diversity of requirements and methods of measurement of interference in the various countries, as was well pointed out at the I.R.E. convention last year in Rochester, N.Y. Various committees are working with the object of submitting a set of standards of performance as well as methods of testing noise reducing devices and also radio receivers themselves. Until we have something to go by, it seems that the following method can give fairly satisfactory guidance in measuring the comparative ability to pick up interference of two systems, one of which usually is an open downlead and the other is a combination of coupling units and their transmission line, connecting the same length of horizontal antenna wire in both instances, to a calibrated radio receiver.

The method which has been used in the development of noise reducing antenna systems will be described below. It involves the use of a sinusoidal, single frequency interference, instead of some artificial source of damped oscillations produced by sparking apparatus. The assumption is made that any pulsating current may be decomposed into an infinite number of frequencies very close together, according to Fourrier's integral. It is also assumed that the receiver is substantially unaffected by disturbances the frequencies of which are outside the narrow band intended to be passed through the various tuned circuits. Cases of shock excitation may be said to form an exception, but, even then, the frequency that causes any response in the speaker is within the band admitted by the tuners. Consequently, if we can eliminate an interfering sinusoidal oscillation (modulated at some convenient audio frequency so as to make it audible), it may be safely assumed that all other forms of interference within reasonable magnitude will also be eliminated to the same extent.

A microvolter is used as source of interference and it is connected to the system in the various forms shown in Figures 1 to 8. In order to establish a direct comparison between an open downlead without coupling units and a system consisting of the same length of horizontal antenna conductor with a switching arrangement that will permit the same total length of wire to act in both as an ordinary "T" antenna and open downlead and as a noise reducing system. Fig. 13 shows the circuit schematically. When S and S' are thrown up, the antenna and transmission line act as a "T" antenna directly connected to the antenna post, a, of the set, and with S and S' down, the units U and U' are thrown into the circuit in the manner called for by the system.

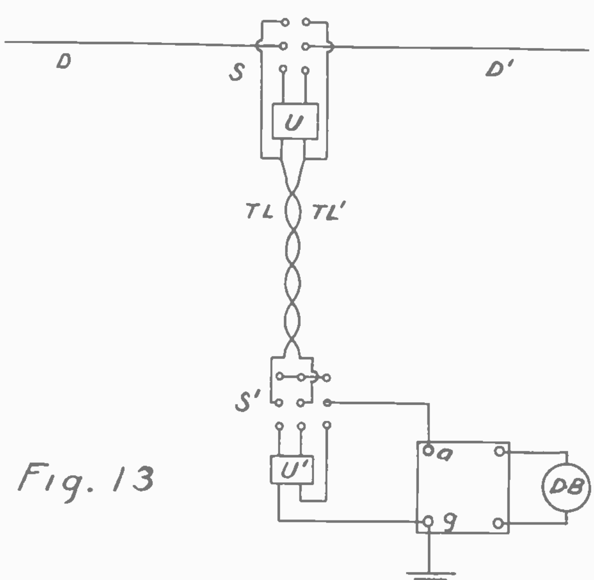

Fig. 13

When making measurements with actual signals, the output level should be observed first with the system on, and immediately with the "T" antenna, using the same station, and noting the signal strength difference, preferably in D.B. A graph should be plotted with frequency as abscisae and D.B. differences as ordinates.

Then the source of interference is connected in the manner to be selected; by means of two small and equal capacities from the microvolter to the lines of the twisted transmission line as in Fig. 2, or in series with the ground lead across a low resistance as in Fig. 4, and readings of the output meter and attenuator taken to determine the D.B. difference in output and a second graph plotted as previously, this will show how much interference reduction is there and then the difference between the two graphs will show the effective interference elimination properties of the system in D.B.

The details of the receiver are interesting. The actual model used in the Laboratory is a superhetrodyne with band pass filter input circuit of the inductively coupled antenna type. In Fig. 14 only the essential parts of the circuit are shown. The terminals a-g constitute the input. Across it there is a resistance R_1 which represents the input impedance of the receiver and can be made of any value desired. A potentiometer of much higher total resistance serves as attenuator. In lieu of this potentiometer a better form of calibrated attenuator should be used when very accurate measurements are required. Resistance R_2 is large compared to the effective impedance of the primary of the first R.F. transformer T_1, so that the potentiometer setting may be almost independent of its load and also that the band pass filter of the set (not shown in Fig. 14) may not become inoperative. V_2 is an R.F. or I.F. amplifier, and there are a number of such tubes in the actual receiver. The I.F. stage immediately preceding the detector tube is provided with means for the modulation of the carrier of the station being received by means of 60 cycles. A low tension transformer such as bell ringing transformer with its secondary in series with the cathode of the last amplifier tube V_2 will modulate the R.F. or I.F. which, after it is amplified, will be fed to the detector V_3 of the linear diode type so that the output shall be proportional to input voltage as far as possible. In the output stage V_4, an output transformer tuned to 60 cycles feeds a copper oxide rectifier DR and a current indicator, M.A. The advantage of modulating the I.F. is very great, as the output current in the mili-ammeter will be independent of the modulations in the program of the station and will be proportional to the carrier which is modulated in the tube V_2.

The tuned output transformer reinforces the 60 cycle modulation and reduces the modulations from the signal. In this manner the signal strength is easily measured by means of a commercial type radio receiver equipped with little extra apparatus and controls. It must be thoroughly shielded and isolated from the power line connections by means of chokes. The A.V.C.

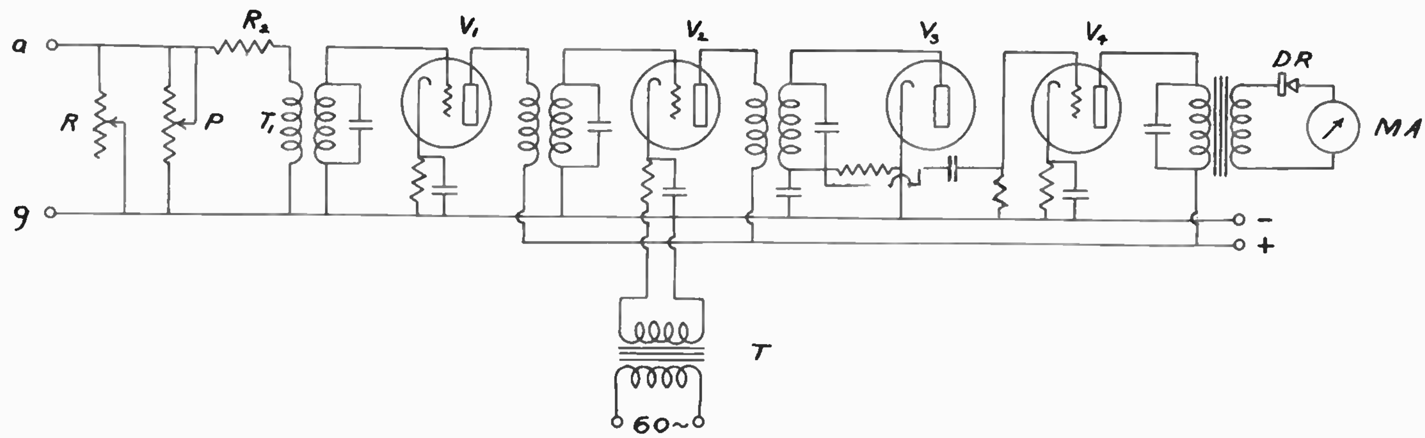

Fig. 14

must be rendered inoperative by using fixed biases or otherwise.

By means of this method, the antenna and transmission line characteristics may be studied experimentally.

ANTENNA DESIGN:

We shall not attempt to cover this broad subject from theoretical or statistical considerations; this is obviously impossible.

From a commercial standpoint, an antenna should be -

a. Easy to erect
b. Inconspicuous
c. Capable of picking up signals throughout the all-wave band
d. Particularly efficient at the regular broadcast band, 15, 19, 25, 31 and 49 meter bands.

Starting with the last requirement, a dipole of some form is necessary which will have no anti-resonant frequencies within the specified short wave band. Straight doublets are preferred for the fulfilment of (a) and (b) but, if these requirements are not very stringent, multiple doublets, "X" doublets or cage doublets are preferable. The "X" doublet of Fig. 15 has the advantage over the straight doublet of the same length, in that its impedance is more uniform and can be used more efficiently at the longer wave band of 49 meters. Characteristic resistance, impedance and power factor curves for this type of antenna are also shown in Fig. 15 (by courtesy of the Hazeltine Corporation).

Double doublets act much the same as double transformers tuned to two frequencies near the ends of the band to be covered. The well known phase reversal of a tuned circuit when the frequency passes from below to above the resonance value, makes it necessary to reverse the doublets with respect to each other to avoid anti-

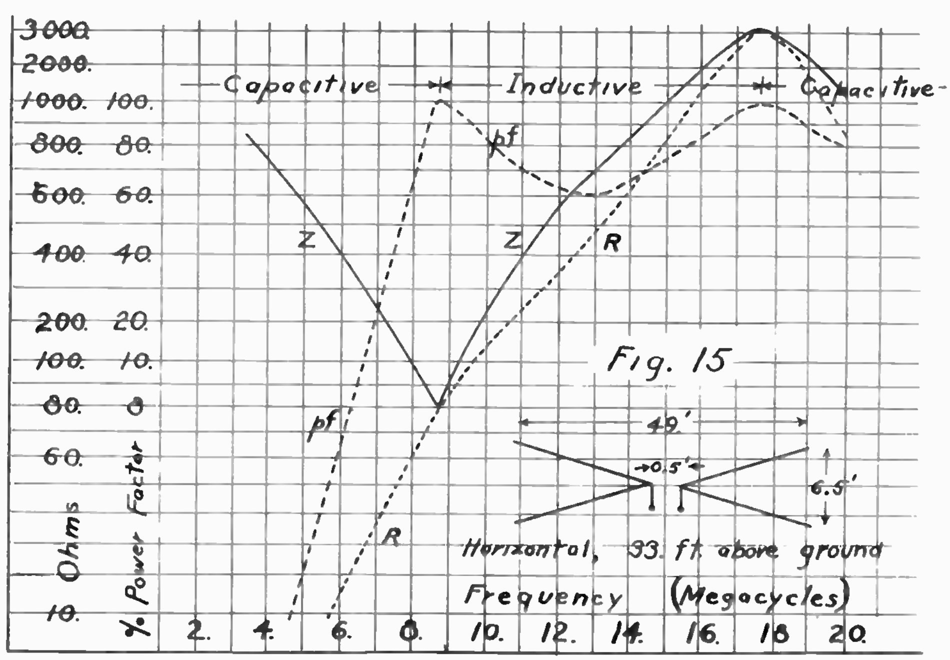

Fig. 15

nodes at some frequency between the two natural periods of the doublets.

If we now use only one of the branches of each of the two doublets, we have an assymmetric doublet. It is used to some extent on account of its simplicity and diversity factor, but as a noise reducer it cannot be as good as a perfectly symmetrical doublet because when interference currents travel upwards on the transmission line, the currents in the two wires are unequal and their difference may be considered as a circulating current which will enter the set coupler just the same as a signal current, as we have already shown in our discussion of interference elimination.

Any of the preceding doublets will act as a flat-top antenna for the broadcast band by means of any of the couplers previously discussed. A pointer in design of these couplers: the leakage reactance of the low frequency transformer, such as 8-9-10-12 of Fig. 9, when the secondary is attached to the line, should tune the capacity reaction of the flat-top antenna at about the middle of the broadcast band, or at a point where desirable signals come in weak.

TRANSMISSION LINES:

The most common forms of transmission line are: a twisted pair cable and a parallel transposed line. The latter has much lower insertion losses, especially in damp weather, but it is much less practical and more difficult to install. With modern receivers of enormous sensitivities, a loss of 4 or 5 db., such as may occur in the cable type line is not serious. The greatest of the losses is ordinarily the dielectric loss between conductors, but, at the very low potentials used, is always small.

The surge impedance of a transmission line is

$$Z = \sqrt{\frac{j\omega L + R}{j\omega C + G}}$$

where L, R, C and G are respectively the inductance, resistance, capacity and conductance per unit length. Lines used as downleads attenuate so little that R and G are very small compared to ωL and ωC and therefore

$$Z = \sqrt{\frac{L}{C}}$$

and Z has unit power factor. This property enables us to measure Z very easily by means of a Wheatstone bridge, as shown in Fig. 16, consisting of a shielded transformer with an accurately center tapped bifilarly wound secondary, a non-reactive resistance R, the value of which may be measured by an ohm meter after the bridge is balanced, and the transmission line terminated by a resistance R', also of the same type as R. When the bridge is balanced, R and R' will be equal and R will be equal to $\sqrt{\frac{L}{C}}$.

When LG = CR, Z is likewise a pure resistance, but in other cases Z will not be a pure resistance; then an approximate balance instead of a dead zero will be obtained, but it is close enough for the determination of Z within a few percent.

Once Z is known, the attenuation of the line may be measured by means of a linear radio receiver (one without AVC and linear detection) and an output DB meter connected as in Fig. 17.

It is interesting to know that transmission lines of the twisted pair type shows a rather low attenuation even at 15 M.C.

The following table shows insertion losses in D.B. per hundred feet of twisted pairs having seven strands of #32 copper wire with 1/32" rubber insulation and cotton braid serving over both conductors.

Frequency in M.C.:	0.5	1.5	5	15
D.B. per 100 ft.:	0.5	0.5	0.5	5.5

Moisture seems to have very little effect. With some of the ordinary twisted pairs there is an increase of 3 D.B. only at about 10 M.C. and up.

The use of the transmission line is not limited to operation of a single receiver. In apartment houses where as many as twenty radio sets must be fed from one antenna, a system of multiple operation has been developed to couple radio sets to the line and known as the DOUBLET MULTI-COUPLER SYSTEM. Fig. 18 illustrates a typical arrangement. The only difference in design is two-fold: The ratio of transformation of the couplers must be such that the reflected impedance of the twenty units may not be too low that will nearly short circuit the line, or introduce abrupt changes in the continuity of the line constants, thereby originating nodes or loops by the formation of quasi-stationary waves along the line. The other feature is the insertion of a decoupling resistance in each coup-

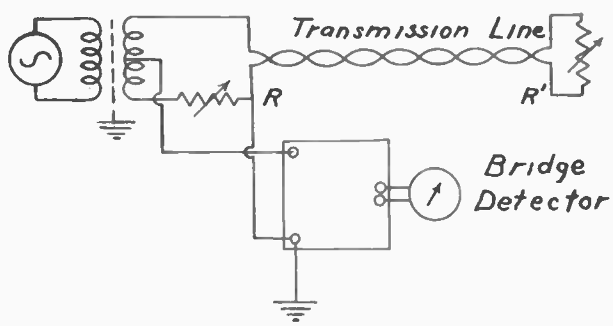

Fig. 16

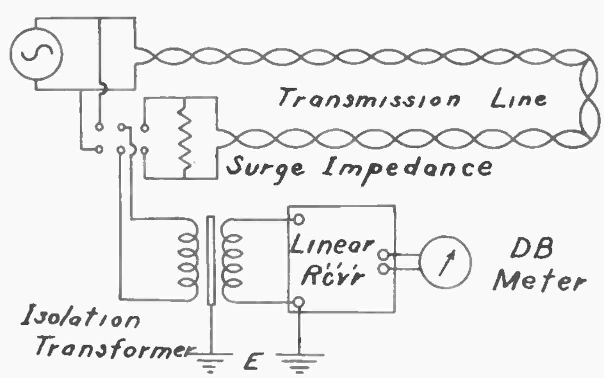

Fig. 17

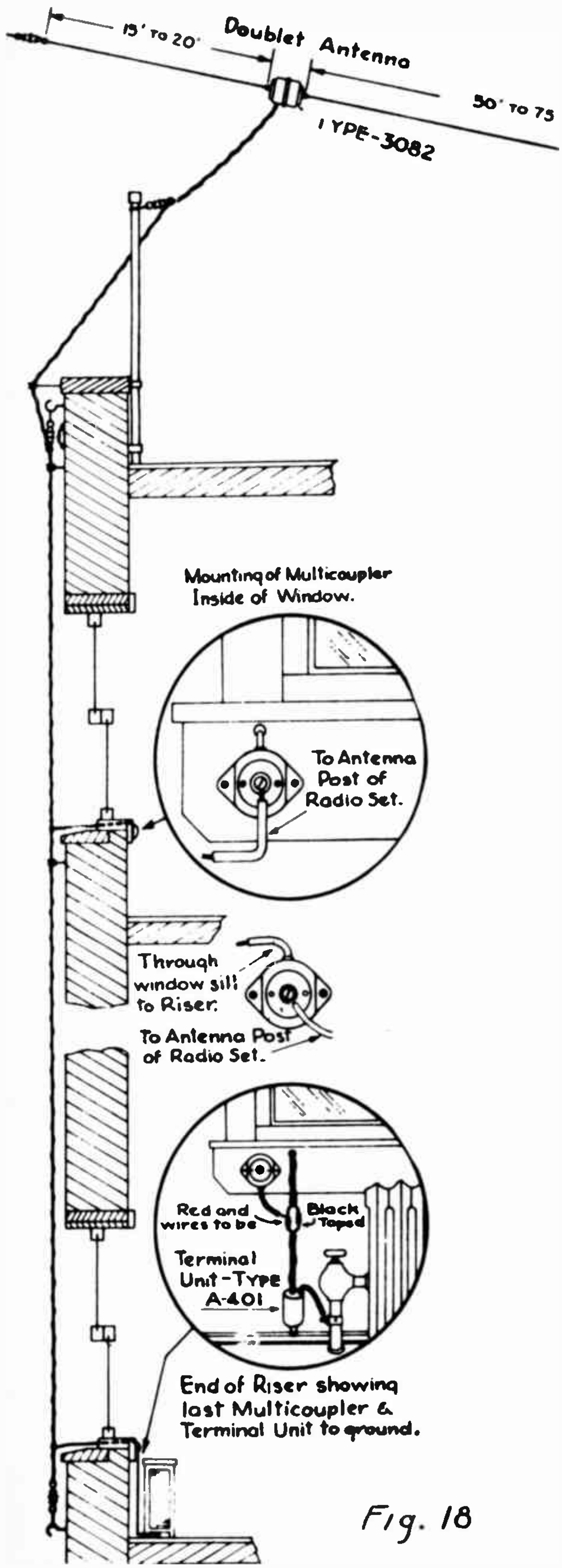

Fig. 18

ler to limit impedance variations due to the tuning of the receivers to various frequencies. A decoupling reactance is not as satisfactory because there is always the danger that some radio set may be adjusted so as to give an equal and opposite input reactance and a short circuit will occur at a certain frequency across the transmission line.

CONCLUSION:

The problem of radio interference is very complex. A segregation of the various components is necessary for its complete analysis. Practical conditions limit the extent to which the existing and proposed remedies may be applied. The design of commercial units has to take all these elements into consideration. The conditions are so variable that experiments, even though very elementary, should be performed in each case to ascertain the choice of grounds. The choice of antenna, its place of erection, and the apparatus should be governed by the relative importance of the various requirements.

In conclusion, the writer wishes to acknowledge his sincere appreciation for valuable advice and cooperation to Mr. E. V. Amy, and for experimental work and assistance in the preparation of this paper to our assistant, Mr. Edward Sieminski.

TWENTY-FIFTH ANNIVERSARY YEAR BOOK OF THE RADIO CLUB OF AMERICA.

A group of schoolboys drawn together by their common interest in the then new art of radio formed the Junior Wireless Club Limited on January 2, 1909. Out of that first meeting has grown the Radio Club of America comprising in its membership every radio experimenter, technician and engineer of importance in the industry today. In celebration of the first quarter century of its activities the club has published this year book which contains photographs of early apparatus, members who were instrumental in promoting the club and a complete history of the development of the organization. An examination of the proceedings of the club from its inception discloses that the majority of the radio inventions which form the bulwark of the modern radio receiver were first divulged to the members. The Anniversary Year Book contains a complete Who's Who of the roster which now includes 320 members.

N.Y.Sun 3/2/35

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PROCEEDINGS of the RADIO CLUB OF AMERICA

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SOUNDPROOFING APARTMENT HOUSES FOR RADIO

BY

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Delivered before the Radio Club of America
February 13, 1936

INTRODUCTION

One man's music may be another man's noise. A delightful radio program in my apartment is transformed into a nuisance by simply leaking into the next apartment. It is apparent that the psychological factors are also very important in dealing with sound control problems.

SOUND PROOFING DEFINED

In this short discussion we shall confine our remarks very largely to the objective and physical aspects. The term "sound proofing" has been used for a great many years. The exact meaning is somewhat indefinite and in some cases confusing. Consequently, let us use the two terms "sound insulation" and "sound absorption". Both terms indicate the reduction of the transmission of sound from one point to another, but the respective means for accomplishing this are different. Heavy walls, ceiling, and floor act as good insulators of a sound source in a room in reducing the transmission of sound to a point outside. Sound absorption applied to the interior surfaces cause a lowering of the acoustic intensity and hence adds to the reduction in the transmission of sound. It is apparent, then, that "insulation" and "absorption" are both means for sound proofing.

SCOPE

We shall further limit this paper to radio loud speakers as a sound source. Among the other sources of sound in an apartment house are the following:

Musical instruments; piano; violin; drums, etc.
Singing and speaking voices; Shouting voices, crying of children, barking of dogs, etc.
Footfalls; jumping; dancing, etc.
Rumbling of toys, bicycles, wagons, etc.

Sewing machines - Vacuum cleaners - Typewriters
Electric fans - Refrigerators - Kitchen aids-
Alarm clocks

Doors, windows, awnings. Elevators, dumb waiters, incinerators. Steam radiators and steam pipes. Water pipes and plumbing equipment.

BUILDING DEVELOPMENT

In this list of sources of sound there are many which have been added within the last generation. The type of building has gradually changed during this period. In the place of massive brick or stone walls we now have thin, flexible walls. In the place of relatively soft thick lime plaster on wood lath we have hard gypsum plaster on metal lath. For wood floors we often find concrete and tile. For the wood doors in wood bucks we now have metal clad doors in metal bucks.

Practically all modern developments are improvements in heating, sanitation and fireproofing. It is evident, however, that these developments have in each case contributed to the magnification and transmission of noise. The public is beginning to recognize this deplorable condition.

OUR NEXT DEVELOPMENT

We, as engineers, tenants, or owners are confronted with the necessity of quieting the modern apartment house, without giving up the improvements in sanitation, ventilation, heating, and fireproofing. It is also desirable to impose less restrictions on the activity of tenants without annoyance to others.

PURPOSE OF THIS PAPER

It is no secret that very little has been done in sound proofing apartment houses. One of the purposes of this paper is to stimulate interest in noise abatement and insulation in apartment houses. Furthermore, as radio engineers, we are anxious that the radio shall not become a nuisance due to improper operation and lack of acoustic control. Hence, a second purpose is to indicate some of the factors which determine when the radio is truly a nuisance due to loudness. A third purpose is to indicate certain improvements which can be made in apartments to provide more soundproofing.

AUDITORY CHART

In the last analysis a noise becomes a nuisance only when it can be heard. When the intensity and frequency of a sound falls within the hearing range it can be indicated on the auditory chart. (Fig 1-D) This chart is now generally accepted as the standard for average hearing. The pitch or frequency extends from 25 cycles per second on the left to 17,000 on the right. The intensity and loudness extend vertically. The horizontal straight lines correspond to physical strength or power of the sound in air. The family of curved lines in-

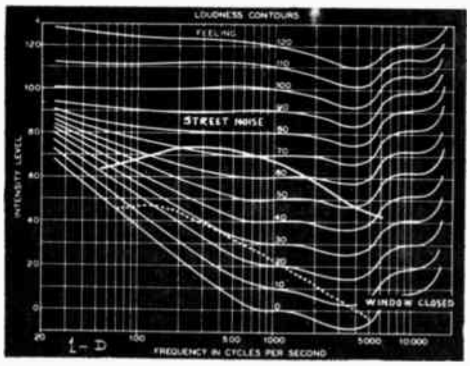


Figure 1-D

dicate contours of equal loudness as sensed by the average ear. For example, a 200 cycle tone whose intensity is 60 will be sensed as only 50 in loudness. The zero loudness corresponds to minimum audibility for each frequency. Hence the 200 cycle tone would have to be reduced only 37 decibels in order that it be placed at zero loudness which is the threshold of audibility. If on the other hand the tone has a pitch of 1000 cycles with an intensity level of 60 it would be sensed as 60 in loudness. This is 60 decibels above minimum audibility It will be noted that at only two points do loudness and intensity level have the same numerical value, namely, 1000 cycles and 5000 to 6000 except for the contours marked 110 and 120 where the equivalence occurs at 6500 and 7500 respectively.

STREET NOISE

One of the factors which influence the acoustic level at which the radio is adjusted is the noise level in the listening room. Among other sources the most universal is street noise. An average curve for the intensity in the street is indicated by the solid line, on the auditory chart. If the listener comes into the room next to the street with the window closed the intensity drops to the dotted line due to the attenuation of the outside wall and closed window. It is characteristic that the high frequencies are reduced much more than the low frequencies. It is quite usual in city apartments to have a noise level of 30 to 40 decibels under these conditions. It is clear that the area below the dotted line is subtracted from the total area which represents the full hearing range.

RADIO CHARACTERISTICS

In "1-A" is shown representative characteristics for the midget and high fidelity radio sets complete. It will be noted that the range of the high fidelity is almost an octave lower and extends almost an octave higher than that of the midget. These curves are placed at a level between 70 and 80 decibels which is a very comfortable loudness assuming that no abnormal noise is present. Due to the limited frequency range it is quite usual for a radio listener to adjust the loudness of the midget set for any individual

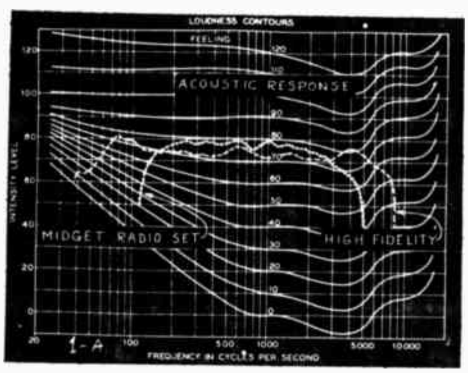


Figure 1-A

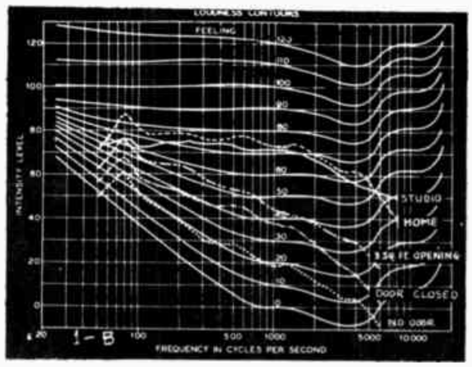


Figure 1-B

middle range tone to a higher level than that for the high fidelity set.

TRANSMISSION THROUGH WALLS AND DOORS

In Fig. 1-B is shown a solid curve which represents an average spectrum of sound of a studio program.

After reproduction the curve becomes the dashed line marked "HOME". The absolute acoustic levels are, of course, assumed. The "STUDIO" curve might actually be much higher as in the case of an orchestra or it might be lower as in the case of a singing voice up close to the microphone. The radio program may be adjusted in level at the will of the listener as best suits his particular hearing conditions.

The "dash-dot-dot" curve corresponds to the resulting curve of the sound spectrum after transmission to the air in the adjacent room when the door (or window) is opened about 6 inches to give a net area of 3 square feet. These curves apply to average apartments which are furnished, providing from 50 to 100 sabines of absorption. When the door (or window) is closed the transmitted curve drops to a lower level as indicated. If the door is removed entirely leaving a continuous wall the curve drops to the dotted curve marked "NO DOOR".

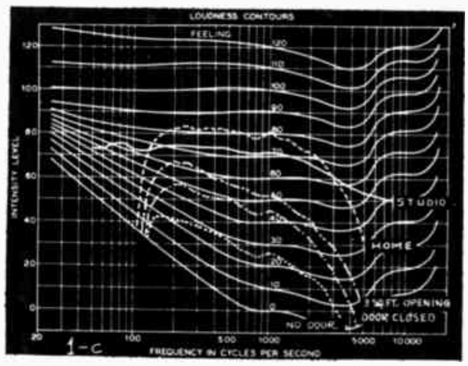


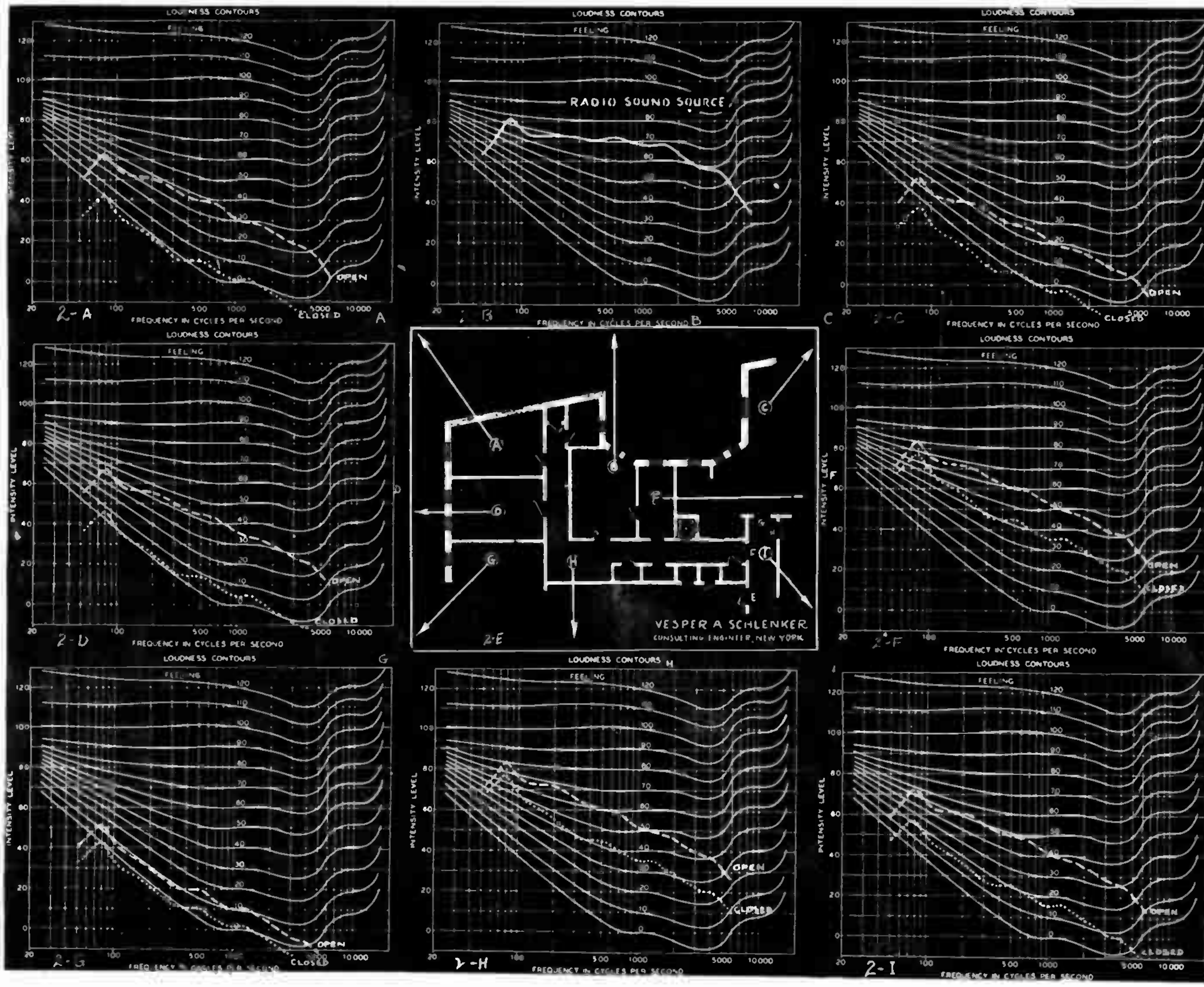
Figure 1-C

Figure 1-C include curves for a midget radio which correspond to the curves shown in the preceding figure. In the presence of 30 decibels of noise it will be noted that the frequency range of the midget is effectively reduced thereby impairing its quality much more than a corresponding condition for the high quality radio.

TRANSMISSION CHARACTERISTICS FOR AN AVERAGE APARTMENT

An average size apartment is shown in plan in Figure 2-E. The radio is located at position "B" near the double windows which look out over an open court. The acoustic spectrum of instrumental music is given in Figure 2-B as reproduced by a high fidelity radio set.

The curve of this spectrum drops to various levels at positions "A", "D", "H", and "F"., of the other



rooms of the apartment under two conditions--"all doors open" and "all doors closed" as noted on the corresponding numbered figures. The curves for "H" and "F" are approximately equal since there is only one door between them and the source. The rooms "A" and "D" have a much greater attenuation because there are two doors and a hall between. Taking 500 cycles per second for example, when the loudness in room "B" is 70 it is 60 at "H", 42 at "D", and 38 at "A", when all the doors are open. If, however, all the doors are closed the loudness drops to 43 at "H", 9 at "D", and 5 at "A". In the presence of a noise of 20 to 30 db. it is evident that the radio will be heard when the doors are open but not when the doors are closed. The adjacent room "F" will have a loudness of 60 with the door open and 42 with the door closed. Therefore, room "F" will not be a suitable bedroom while rooms "A" and "D" will be free from annoyance of the radio when the doors are closed. This illustrates the importance of arranging a room or hall between the living room and bedrooms.

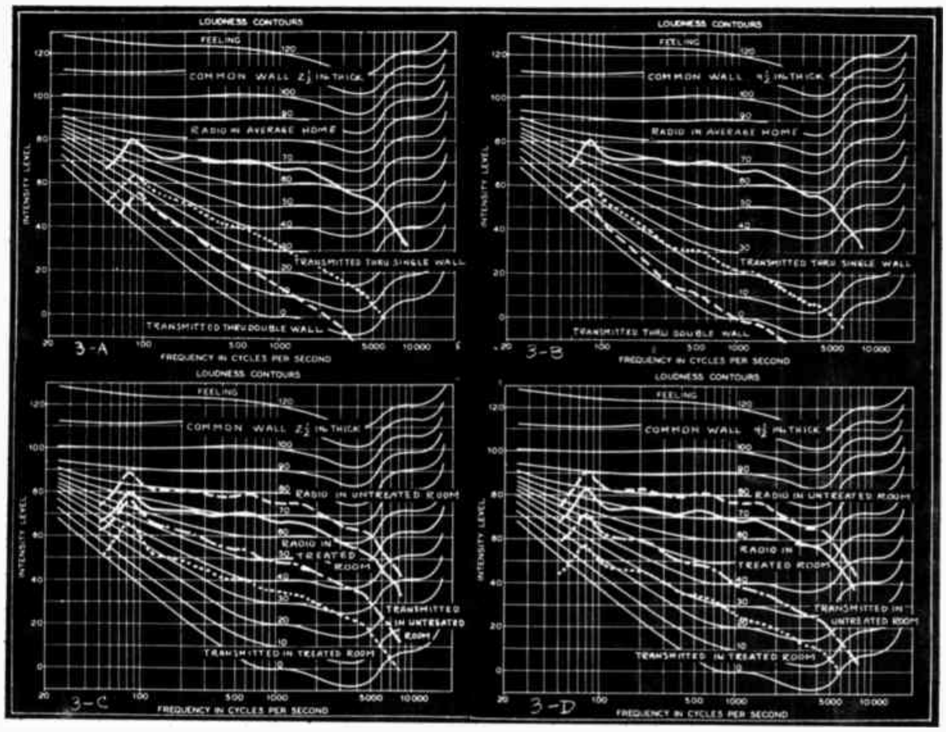
In like manner the loudness can be determined from the curves for the public hall "I", the east apartment "C" and the south apartment "G". In the case of "C", the sound travels through two windows and the open court. This attenuation is approximately equivalent to that for the path by way of the hall "I" and the entrance doors. In the case of apartment "G" the sound must travel through continuous walls. It will be noted, then, that in the presence of 20 to 30 db. of noise that neither apartment will be annoyed by the radio at "B" whether the windows and doors are opened or closed. This would not be true in the case of "C" if the intervening court were closed on all sides. In that event, the curves for "C" would be similar to those for "D". Open windows would then result in considerable annoyance from the radio.

DOUBLE WALLS AND FLOORS

The effect of different wall and floor construction is illustrated in Figure 3. The solid curve again represents the source. The dotted curves in 3-A indicate the spectra after transmission through single continuous walls about 2½ inches thick, constructed of gypsum tile and plaster. If now a second wall similar to the first is built providing an air space of 2 inches the attenuation is improved as is shown by the dashed curve. Figure 3-B gives the corresponding spectra for the 4½ inch wall, single and double. It is evident that in the presence of a noise loudness of 20 that the 4½ inch single wall will most probably be entirely satisfactory whereas the 2½ inch wall will permit too much transmission. It is also interesting to note that the double wall 2½ inches thick is only 6 db. lower in transmission than the single wall 4½ inch thick. It can be concluded that the double 4½ inch wall will be satisfactory even in the absence of masking noise for the given condition of 70 db. loudness of the source.

SOUND ABSORPTION TREATMENT

The value of absorptive treatment for sound insulation is frequently assumed to be much greater than it actually is in practice. Figures 3-C and 3-D indicate the reduction in transmission for the 2½ inch and the 4½ inch walls previously considered. The solid curve indicates the average spectrum of sound in the room if it were heavily treated to prevent appreciable reverberation. The dashed curve represents the condition in the same room



due to noticeable reverberation when the interior has very little absorption, which results in a sound level of about 7 decibels higher or 77 db. The attenuation of the wall lowers this curve 30 db. but the reverberation of the receiving room raises it another 7 db. resulting in a level of 54.

On the other hand, if both rooms are heavily treated with absorption the resulting level will be approximately 40. Therefore, if both rooms are treated the insulation is 14 db. better than if both rooms are reverberant. Similar conditions are true for other types of walls.

FLOORS AND CEILINGS

The previous considerations of walls are equally true of floor and ceiling construction which corresponds in stiffness and mass. For example, a 4" concrete slab with a plastered ceiling underneath would have approximately the same characteristics as that shown for 3-B. If a metal lath and plaster ceiling were suspended below the slab the attenuation would, of course, be greater.

SOUND FROM IMPACTS

Since this discussion is limited to radio as a source of sound the serious disturbances resulting from percussion on the floor are not included in the preceding considerations. In other words, we are assuming that the transmission is from air to air.

VARIATIONS IN SPEECH INTENSITIES

In all the preceding characteristics the curves represent average values in intensity. In applying these charts to actual conditions it must be remembered that the variations from the average are usually plus or minus 10 db. and may be as much as plus or minus 20 db. This fact is effectively illustrated by a reproduction of an actual oscillographic photograph of speech picked up by the microphone about 18 inches from the speaker. The speech is separated into three traces recorded simultaneously. The upper trace are made up of frequencies below 500 cycles which include the fundamentals of the vowels. The center trace consist of frequencies between 500 and 2000 cycles which contain the harmonics and overtones of the vowels. The bottom trace record the essential frequencies of the consonants, which lie above 2000 cycles. This division of the frequency spectrum has been made in order to analyze the behavior of speech transmission in auditorium surveys where the effect of special acoustical treatments must be determined. In this record the deflections are proportional to the sound pressure impressed on the microphone. Hence a half deflection in pressure is 6 decibels below full amplitude. In like manner one fourth corresponds to 12 decibels below since the energy of the sound is proportional to the square of the pressure. For average English

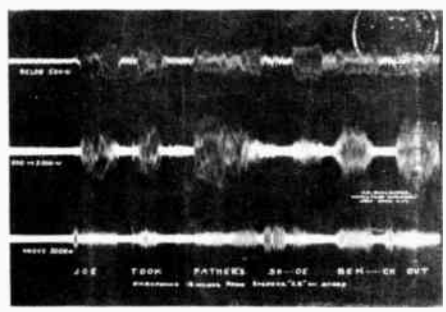


Figure 4

speech the consonants are 14 decibels below the level of the vowels. Of course, this relationship is widened when speech is transmitted through walls and openings as is illustrated in the preceding charts. This accounts in part for the very low articulation of speaking voices when heard through a wall or door. The other important factor is the parasitic reverberation and transients introduced by multiple reflections and vibration of the wall itself.

WINDOW VENTILATORS

It is quite evident from the preceding discussion that windows and doors should be kept closed. The outside noise must be effectively attenuated so that the noise level in the listening room is low as possible, certainly far below 40 db. This is important for comfortable listening and makes it possible for the listener to adjust the acoustic level of the radio output to a reasonably low level, not higher than 70 db. or 80 db. at most. If the radio is adjusted to higher levels the transmission to adjoining rooms and apartments will be raised

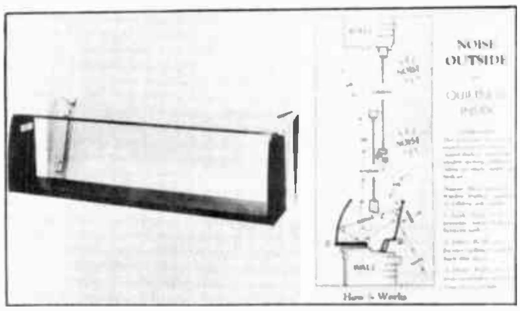


Figure 5

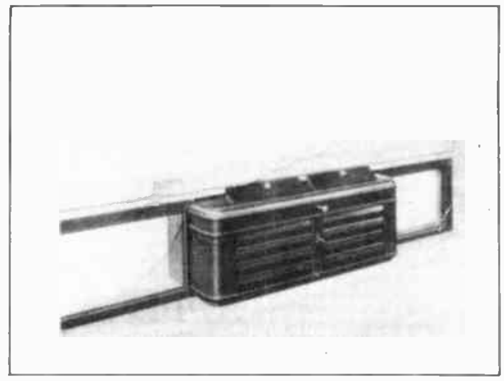


Figure 7

in a corresponding amount, thereby causing an annoyance to neighbors in spite of substantially built walls.

One of the first commercial window ventilators which was placed on the market for the expressed purpose of excluding the noise but not the air is shown in Figure 5. The explanation of its action is also given to show how the manufacturer relied on reflection to exclude the noise. It is fairly well appreciated now that only the higher frequencies have appreciable directional properties and hence this ventilator is effective where the noise has a substantial amount of its energy in the middle and high frequencies.

A much more efficient (and costly) window unit has been placed on the market in the last few years. Figure 6 shows a phantom view which illustrates how the air is brought in through a dust filter and sound absorbing path by an adjustible speed fan and electric motor. The louvers are also adjustible in direction and area of opening. The reduction of outside noise through this unit is approximately equal to that of the closed window, which usually falls between 15 and 25 decibels of reduction.

Later models offered by two leading manufacturers are illustrated in Figure 7 and 8. These units do not extend past the inside edge of the window sill and hence do not interfere with curtains or take up floor space. These units can easily be lifted out so that the windows may be cleaned in the usual manner.

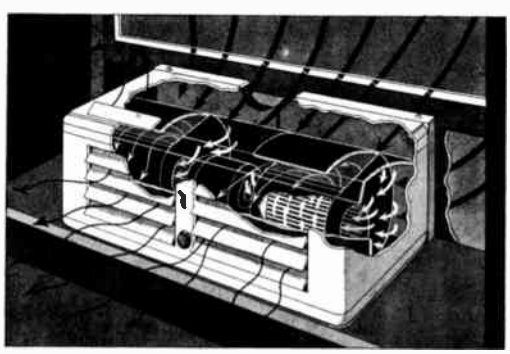


Figure 6

CONCLUSION

Some of the important factors for the improvement of the apartment house can be summarized as follows:

1. Provision for positively controlled ventilation with closed window conditions, acoustically.

2. Provision for double walls and suspended ceilings below double floors between apartments horizontally and vertically, respectively.

3. Elimination of closed courts or light walls.

4. Provision for acoustical treatment of all halls, both inside and outside of apartments.

5. Provision for vibration isolators under radio cabinets, pianos, etc.

6. Arrangement of rooms so that closets are situated between apartments along common walls.

From the standpoint of noise reduction it is always desirable to have all floors covered with carpets or rugs laid over heavy cushion fabric or mat. This provides acoustical absorption and also reduces the impact noises transmitted to the room below.

It is quite clear, then, that the owner, engineer and architect should include adequate provisions for sound proofing throughout all stages in the planning and building of a new apartment house.

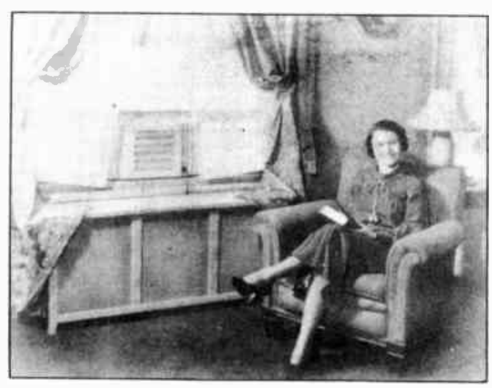


Figure 8

No.....................

RCA FOUNDED 1909

The Radio Club of America

11 West 42d Street

New York City

APPLICATION FOR MEMBERSHIP

TO THE BOARD OF DIRECTORS: ....19.......

I hereby make application for membership in THE RADIO CLUB OF AMERICA, and agree, if elected, that I will be governed by the Constitution as long as I continue a member.

....
(Signature in ink)

NOTE—Applicants are urged to read extracts from the Constitution on the back of this blank.

Answer All Questions

NAME (Print in Block Letters) Mr..

Address—

* Home ☐..

* Business ☐..

** Indicate by X address to which notices should be sent.*

General and technical education; degree held, schools or colleges attended:....................................

....

What radio or technical organization do you belong to? ....

....

What particular branch of the radio art are you interested in? ....

....

Date and place of birth..

Occupation ....

References:

....

....

....

Do not fill in below this line

Receipt acknowledged ....Elected ....

Deferred ....Advised of election..

Remarks ....

....

Proceedings

of the

Radio Club of America

Incorporated

July, 1936

Volume 13, No. 2

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11 West 42nd Street + + New York City

The Radio Club of America, Inc.

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PROCEEDINGS
of the
RADIO CLUB OF AMERICA

Volume 13 July, 1936 No. 2

COSMIC CYCLES AND RADIO TRANSMISSION

BY

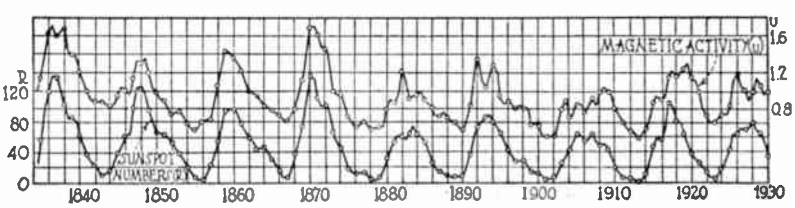
HARLAN TRUE STETSON, PH. D.

Delivered before the Radio Club of America
June 3, 1936

Everyone connected with radio transmission knows that there are certain times when communication is superior and other times when it is much inferior. Listeners to oroadcast programs are aware of the fact that conditions change from time to time in the behavior of receiving conditions quite beyond the control of the receiving set. Variations in the amount of static and periods of fading often exhibit themselves at longer and shorter intervals depending upon conditions in the upper atmosphere that are receiving an increasing amount of study. Investigations extending over a decade or more leave little doubt in the minds of scientists who have probed these problems that there are cosmic effects taking place in the ionosphere through which radio waves are propagated. The study of these cosmic effects, the cycles which they undergo, and the ultimate causes, is a new and coming field in science which is likely to be of greatest importance not only from the scientific standpoint but from the point of view of the whole radio and communication industry.

The most obvious periodic change in the conditions affecting radio wave communication is that produced by the successive passing of daylight and darkness as the earth rotates upon its axis. Certain high frequencies invariably give better performance by day than by night. The region of the radio spectrum known as the broadcast band shows better performance in darkness than in daylight. The increase in the range of broadcasting stations is very noticeable a few hours after sundown. Even so, this diurnal cycle does not perform the same day after day but is subject first of all to wide variations due to seasonal effects of cosmic origin. The changing declination of the sun due to the combination of the earth's orbital motion and the inclination of the axis of rotation to the plane of its orbit is, of course, directly responsible for the inequality of daytime and darkness during the summer and winter seasons. The radio operator who knows his communication range must be provided with charts giving available ranges for different frequencies at different seasons of the year. Since the sun is the principal source of ionization of the earth's atmosphere that makes radio wave propagation possible, we can see that anything which influences the quantity, duration, or character of solar radiation must inevitably produce corresponding results in the radio business.

The results of studies carried on in the last decade indicate that the well known solar cycle of approximately 11 1/4 years' duration has its counterpart in radio wave propagation. As will be shown later, quantitative measurements of field intensities appear to show that during periods of marked solar activity such as is generally indicated by the appearance of large numbers of sunspots, the seasonal effect is more than counteracted. When one, therefore, endeavors to search out the cause of behavior of cosmic cycles he is at the same time anticipating the prediction of long range forecasting of conditions in the ionosphere upon which radio communication depends.



Curve of sun-spot numbers and earth's magnetic activity.
"Reproduced from 'Earth, Radio and The Stars' by H. T. Stetson. McGraw, Hill Book Company, Inc."
Figure 1

When asked if astronomical bodies other than the sun produce any effect upon radio communication lines, it already appears safe to answer this question in the affirmative. There is, for example, what appears to be direct evidence that the moon as well as the sun has an effect on the ionosphere. The same appears to be true of meteoric showers. Whether or not the planets can have any effect directly on the electrical condition of the earth's atmosphere or indirectly through causing disturbances in the solar atmosphere which in turn affects the ionizing power of solar radiation must be for the moment left open for question.

To approach the problem of the possible effect of cosmic cycles upon changing conditions in the conductivity of the earth's upper atmosphere from which radio waves are returned, it is well to examine various known sources of atmospheric ionization excluding those of terrestrial origin such as radio activity in the earth's crust. Skellett of the Bell Telephone Company has conveniently summarized extra terrestrial sources of ionization in the following table.[1]

Sources of Ionization	Energy Received by the Earth, Ergs per Square Centimeter per Second
Ultra-violet light from the sun	28.35
Meteors during meteoric shower (a.m.) up to	2.4
Ultra-violet light from the stars (approximate) . .	0.014
Cosmic rays	0.00031
Meteors - average normal day: a.m.	0.00024
p.m.	0.00012
Ultra-violet light from the full moon	0.000044

An examination of this table indicates that the ultra-violet light of the sun is by far the principal ionization agent to be considered; amounting to 28.35 ergs per square centimeter per second. Skellett estimates that meteors during the dates of particular showers take second place. It is perhaps a bit surprising to see the low value of the energy available for ionization that can be attributed to the high penetrative radiation of which we have heard so much, commonly called cosmic rays. It should also be noted that the lowest value of all in the table is attributed to the ultra-violet light from the full moon. This is important to keep in mind in connection with any explanation for a lunar effect on radio communication of which more will be said.

If we view a cross section of the earth's atmosphere from its surface out, some 10 kilometers must be allowed for the troposphere with which classical meteorology is concerned. Above this troposphere lies the isothermal region with a temperature of -55°C. Here the stratosphere begins. Judging from the latest scientific results of Stevens and Anderson in Explorer II, it appears that in the lower regions of the stratosphere we encounter the beginnings of the ozone level. Spectroscopic observations of sunlight at Captain Steven's record altitude of 72,395 feet above sea level indicated that at this altitude some 25 per cent of the total amount of ozone was below the observers.(2) From this level up, therefore, the amonnt of ozone must show considerable increase probably extending to the altitude of the ionized layers. From 50 to 80 kilometers the ionization of the atmosphere appears to be sufficient to reflect very long waves and to this region has been ascribed the D-layer. From 80 to 120 kilometers we arrive at an electron density great enough to turn back waves of broadcast frequency. The mean height of this E-layer, commonly referred to as the Kennelly-Heaviside layer in honor of onr American and English engineers, will vary with the change from day to night conditions and from season to season, as well as with the solar cycle or any other cause that may affect the population or density of its electron or ionic content.

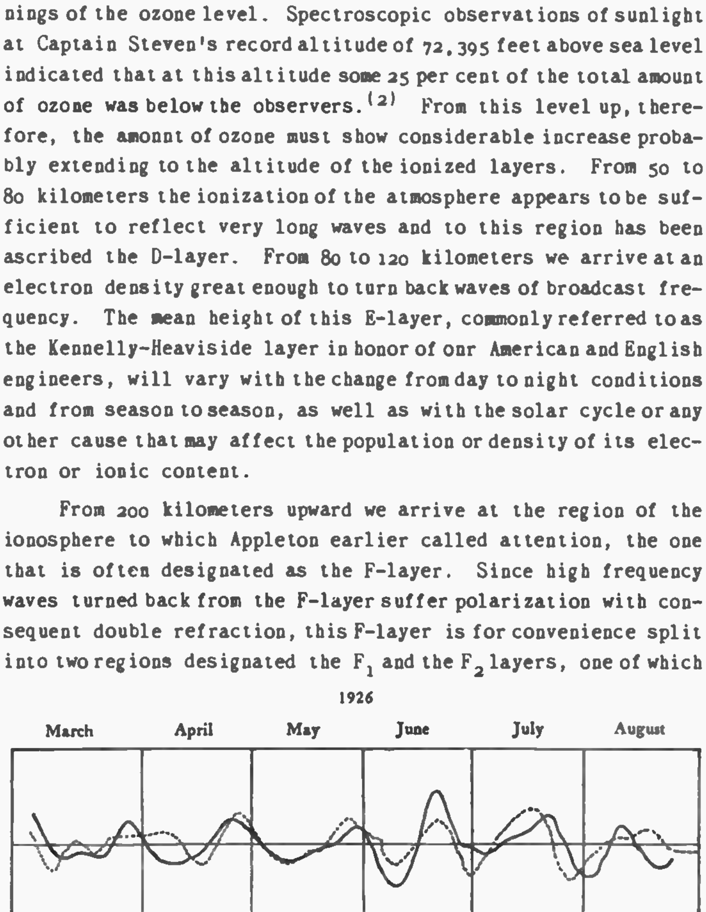
From 200 kilometers upward we arrive at the region of the ionosphere to which Appleton earlier called attention, the one that is often designated as the F-layer. Since high frequency waves turned back from the F-layer suffer polarization with consequent double refraction, this F-layer is for convenience split into two regions designated the F_1 and the F_2 layers, one of which turns back the ordinary and the other the extraordinary wave resulting from the double refraction.

Any cosmic phenomena, therefore, which cyclicly change the degree of ionization or the electron density will affect the success of the transmission of radio waves which depend upon the momentary constitution of the ionosphere. The radio, therefore, gives us a new tool for exploring the upper atmosphere to the height of 200 or 300 kilometers.

Two different methods may be employed for getting information concerning ionic variations. One of these is the direct reflection method such as has been used by Appleton and others in England and at the United States Bureau of Standards, Cruft Laboratory, and elsewhere in the United States. The elapsed time from the time of emission of a radio impulse and its subsequent reception at an adjacent point is recorded on the oscillograph and gives a measure of the virtual height from which the transmitted wave is reflected or turned back to earth. These virtual heights are, of course, calculated on the assumption that the radio wave is propagated with the velocity of light and cannot be taken as direct height on account of some uncertainty as to the variation of velocity with the conditions encountered. However, these virtual heights which vary widely from day to night show a definite seasonal trend as the degree of ionization at the respective levels from which known frequencies are reflected change with the duration and intensity of sunlight on the upper atmosphere.

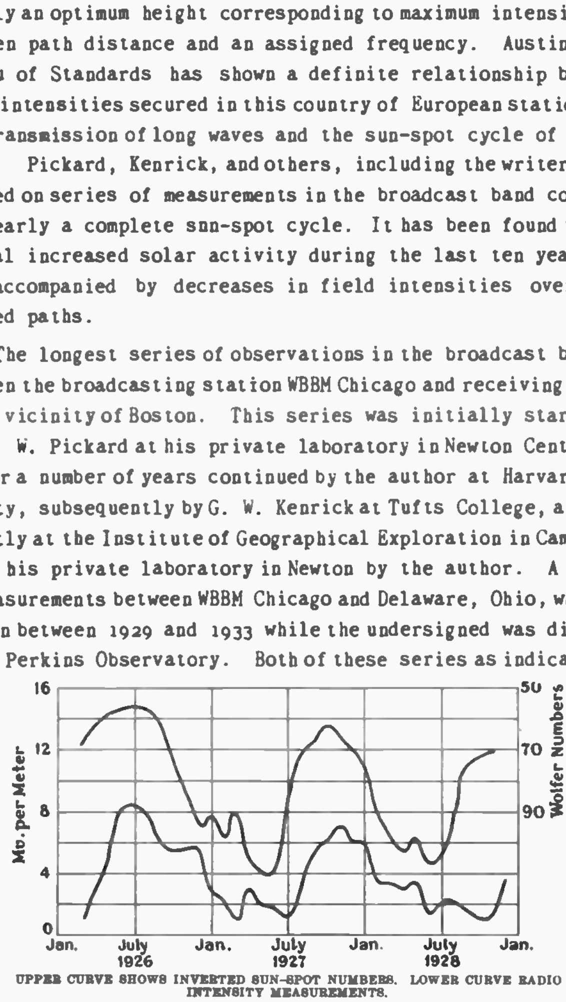
Another method of utilizing radio for obtaining variations in layer heights is through the measurement of field intensities at fixed points from the transmitting station since there is presumably an optimum height corresponding to maximum intensity for a given path distance and an assigned frequency. Austin at the Bureau of Standards has shown a definite relationship between field intensities secured in this country of European stations for the transmission of long waves and the sun-spot cycle of 11 1/2 years. Pickard, Kenrick, and others, including the writer, have carried on series of measurements in the broadcast band covering now nearly a complete snn-spot cycle. It has been found that in general increased solar activity during the last ten years has been accompanied by decreases in field intensities over prescribed paths.

The longest series of observations in the broadcast band is between the broadcasting station WBBM Chicago and receiving points in the vicinity of Boston. This series was initially started by Mr. G. W. Pickard at his private laboratory in Newton Center and was for a number of years continued by the author at Harvard University, subsequently by G. W. Kenrick at Tufts College, and more recently at the Institute of Geographical Exploration in Cambridge and in his private laboratory in Newton by the author. A series of measurements between WBBM Chicago and Delaware, Ohio, was carried on between 1929 and 1933 while the undersigned was director of the Perkins Observatory. Both of these series as indicated in



CURVE SHOWING CORRELATION OF SUN-SPOTS WITH RADIO RECEPTION
DOTTED CURVE, THE INVERSE OF SUN-SPOT NUMBERS; FULL CURVE, RELATIVE INTENSITY OF RADIO RECEPTION ON TRANSATLANTIC, SOUTH AMERICAN AND CONTINENTAL RECEPTION.

Figure 2



UPPER CURVE SHOWS INVERTED SUN-SPOT NUMBERS. LOWER CURVE RADIO INTENSITY MEASUREMENTS.

Figure 3

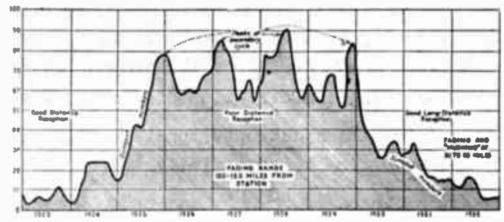
the graph have shown cyclical fluctuations that appear to reflect the degree of solar activity as measured by the number of sun-spots.

It has been found in the Chicago - Delaware series that if the sun-spot numbers are restricted to the central zone, corresponding to the position of spots near the sun-earth line, a somewhat closer correspondence is found between the graph of radio field intensities and solar activity than when spots for the whole disc of the sun have been included. This would appear to indicate that variations in the ultra-violet light are not alone responsible for the radio effects. While reception from a distance of several hundred miles appears to vary inversely in its field strength as compared with sun-spot numbers, the quality of reception at points nearby a broadcasting station may behave contrarily. This is usually explained on the grounds of interference between the sky wave and the ground wave. At times of sun-spot minima with a consequent ascent of the Kennelly-Heaviside layer, it may happen that the sky wave is reflected so well that it interferes at intervals with the ground wave at distances within 50 miles of the sending station, thus seriously hampering good reception. With the lowering of the Kennelly-Heaviside layer accompanying increasing solar activity, the sky wave ceases to interfere as effectively with the ground wave, thus we may actually get better reception within local areas at times of sun-spot maxima than at times of sun-spot minima.

In the case of distance reception, on the other hand, the lower Kennelly-Heaviside layer is less conducive to long distance transmission in the broadcast band, and even nighttime conditions at times of great solar activity simulate the usual effects of daytime reception in the broadcast band. It seems highly desirable that more of the relatively inexpensive equipment for measuring field intensities be established at strategic points throughout the country if we are to obtain more complete data for studying cosmic effects upon transmission in the broadcast band of the radio spectrum.

There is room for a difference in opinion concerning the exact mechanism whereby sun-spots produce changes in the ionization of the reflecting layers. The author has discussed at length in another place(3) the relative merits of the hypotheses of ultra-violet flares accompanying the formation of the spots and the theory that the spots themselves are centers of electronic or some kind of corpuscular emission from the sun-spot centers. Observations made during times of total solar eclipses show the immediate screening effect of the sun's ultra violet radiation in reducing the degree of ionization at a given height or the effective change in the Kennelly-Heaviside layer accompanying the optical shadow as the moon passes in front of the sun. Conflicting results give some evidence for a theory of ionization by corpuscular radiation advanced by Chapman. Evidence concerning this has likewise been discussed at length elsewhere. It would appear not improbable that superimposed upon the effects of ultra-violet radiation which may or may not be locally increased with the formation of sun-spot zones there is still room for a corpuscular hypothesis which is used so effectively by Stormer in brilliant studies of auroral forms.

The recent coincidences of eruptions in the sun observed at the Mount Wilson Observatory with time of complete fade-outs of transoceanic reception noted by Dellinger(4) is at least corroborating evidence for an intimate relationship between particular events on the sun and prompt response in the ionosphere. That the period of fade-outs has in many instances indicated intervals of 54-days or approximately twice the synodic period of solar rotation at the equator is especially puzzling and so far without ex-



Curve of sun-spot numbers of the last solar cycle showing effect upon radio reception.

Figure 4

planation. It is most urgent that those engaged in radio operations and especially those in charge of continuous operations should cooperate with the Bureau of Standards in expediting reports of all such fade-outs whether it be suspected as a world-wide event or not. With the continued acquisition of information as to these periods of remarkable change in communication characteristics, it will be possible to more readily correlate corresponding solar events.

The relatively definite cycle of solar activity makes it possible to make general predictions concerning radio transmission. It is interesting to note that, as exhibited in the curve, many secondary fluctuations in the sun-spot numbers are reflected in the performance of radio reception. A period of fifteen months pointed out by the writer(5) has many times shown its counterpart in the curve of field intensities. The reason for the fifteen-month period in solar fluctuations or even for the 11 1/2-year period in sun-spot activity is not yet known. When we have obtained some evidence for the fundamental cause of these solar fluctuations, we shall have gone a long way to a more accurate prediction of the cosmic conditions affecting the radio industry. Meanwhile, it is perhaps worthwhile to reflect upon such hypotheses as have been considered as a possible basis of the fluctuations in the solar cycle.

Numerous attempts have been made from time to time to account for periodicities in the appearance of sun-spots on the basis of the various planetary periods. The sidereal period of Jupiter most nearly approximates that of the fundamental cycle in solar activity. Jupiter's period, however, is 12 years and not the mean sun-spot period of 11.25 years. The sun-spot period, however, is far from constant as in certain instances the interval between maxima has been as little as 8 years and in one case as long as 17 years. An exhaustive examination by E. W. Brown(6) made 36 years ago suggested at that time that a combination of the period of Jupiter with that of Saturn (29 1/2 years) showed a correlation with the sun-spot period. Later, however, discrepancies cast doubt in the same author's mind upon the validity of these earlier conclusions. One difficulty with all planetary theories for explaining the appearance of sun-spots is that the tidal action of the planets is too small to appear to be significant in causing eruptions in the solar atmosphere on gravitational grounds. If one were to suppose, however, that the planets are at different electrical potentials, then there is perhaps a fresh basis for attack on the sun-spot theory from the planetary viewpoint.

As has been pointed out by the writer(5), the fifteen-month secondary period in solar activity coincides very closely with the combined periods of Mercury (88 days) and that of Venus (225 days). If electric charges on the planets exist, it is not easy to separate their effect from the gravitational effect since both would follow the inverse square law. There are only a few thinkable instances where planetary perturbations would make it possible to differentiate an assigned value for gravitational mass from a similar value confused with a possible effect due to electro-static charges. Some investigations are now being made to consider the results of combining forced period oscillations with assumed values of natural periods of oscillation of a solar atmosphere.

Current opinion among astrophysicists is against any planetary theory to account for solar activity. Preference is being given to periodic disturbances in the internal structure of the sun that presumably give rise to disturbances in the outer atmosphere thus producing the sun-spots. Application of the hydrodynamic principles to the formation of solar vortices by Bjerknes[7] have yielded some plausible explanations for the formation of sun-spots at the extremities of tubular disturbances extending below the surface of the solar photosphere. No explanation, however, on such grounds has been forthcoming for eleven-year periodicity. Bjerknes shows, on the basis of his deductions, how spots may be expected to migrate in latitude as the cycle progresses.

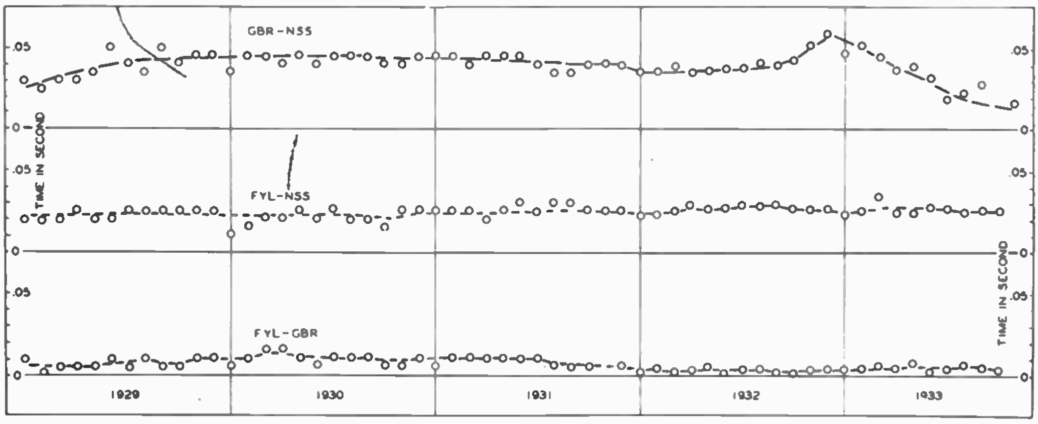
Notable correspondences between the occurrence of sun-spots and the disturbances in the earth's magnetic field have been on record for more than a century. Investigations in the performance of radio communication in recent years have shown disturbances concomitant with magnetic storms. Many radio engineers, therefore, have supposed that investigation of cycles in radio transmission with cycles in terrestrial magnetic activity is a more profitable field for exploration than the correlation of radio phenomena with solar activity. It is the opinion of the author, however, that solar disturbances are the primary cause for both magnetic variations and radio disturbances. Since any variations in the sources of ionization of the upper atmosphere of the earth would affect the ionic and electronic density of the ionosphere, it is easy to see that the electrical currents due to the rotation of the ionosphere would correspondingly vary. The current set up by motion of and in the ionosphere would immediately be reflected in changes in the magnetism induced in the earth by such variations. If we were to regard the ideal case of an ionosphere in equilibrium rotating with the earth, then the electric current produced by the motion of the electronic shell would induce magnetism in the earth symmetrical about the geographical poles. The compass needle at any given moment, therefore, would point in the direction determined by the resultant of the permanent magnetic field of the earth, whose axis lies in the direction of the earth's magnetic poles, and the field due to the induced magnetism determined by the geographical poles.

Any cosmic effects which would change the degree of this ionospheric shell would, therefore, vary the component due to the induced magnetism. In this way a well known diurnal change in the declination of the compass can be explained since the degree of ionization in the upper atmosphere is different for the illuminated and unilluminated halves of the globe. Any sudden disturbances on the sun which would upset the distribution of ions in the ionosphere might, therefore, be expected to be accompanied with corresponding disturbances in the earth's magnetism.

Herein also would appear to lie an explanation for a lunar period in terrestrial magnetic variations as has been previously hinted. In our quantitative studies of radio reception which show so unmistakably the effects of solar activity, considerable evidence exists for a lunar tide in the ionosphere. If such tides that correlate with the moon actually exist, the changed distribution of electrons with the position of the moon should result in variations in the magnetism induced in the earth, hence the lunar cycle in both magnetic activity and radio field intensities. Curves of field intensities both between Chicago and Boston, and Chicago and Delaware, Ohio, show systematic changes with the hour angle of the moon.

Through the courtesy of Professor Mimno of the Cruft Laboratory at Harvard University it has been possible to examine the percentage of reflection of waves of 3492.5 kilocycles frequencies from the E-layer and as has been reported[8] from about 10,000 hours of observations included in the material examined, there was an increase in the percentage of the time of reflection from the E-layer from 12 to 22 per cent as the moon passed from conjunction with the sun to a position a little past full. Even making allowances for a suspected seasonal correction there remained an 8 per cent increase of reflections in the E-layer as the difference between the hour angle of the sun and the moon increased from 0 hours to 14 hours. A corresponding decrease in the percentage of reflections accompanied the change in hour angle differences from 14 hours to 24 hours. We may summarize this by saying that near full moon there is a tendency for an increase in ionic density on the night half of the earth's atmosphere thus favoring increased numbers of reflections from the E-layer. At new moon, on the other hand, any effect which the moon may have, has probably been lost in the solar effect on the daily half of the earth's atmosphere. Whether the lunar effect on the ionic or electronic density at a given level is due to gravitational or other sources has not yet been determined.

Perhaps one of the most striking results from our investigations of variabilities in radio transmission is the apparent variation of time elapsed in the propagation of waves utilized for the intercomparison of time signals between observatories on either side of the Atlantic.[9] After making all reasonable allowances for variations in lag, the actual computed times indicate 100 per cent variation in the effective velocity of the 17 kilocycle waves between 1929 and 1934 over the Annapolis - Rugby path. Similar but less drastic variations occur from intercomparisons between Annapolis and Bordeaux, and Bordeaux and Rugby. While the assumed great circle routes over which the waves are propagated are not vastly different between the United States and



MONTHLY MEAN VALUES OF APPARENT TIME OF PROPOGATION OF 17-KILOCYCLE TRANS-ATLANTIC TIME-SIGNAL

Figure 5

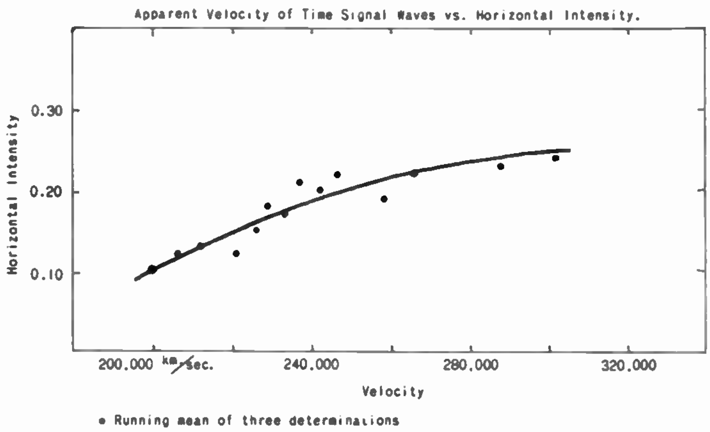


Figure 6

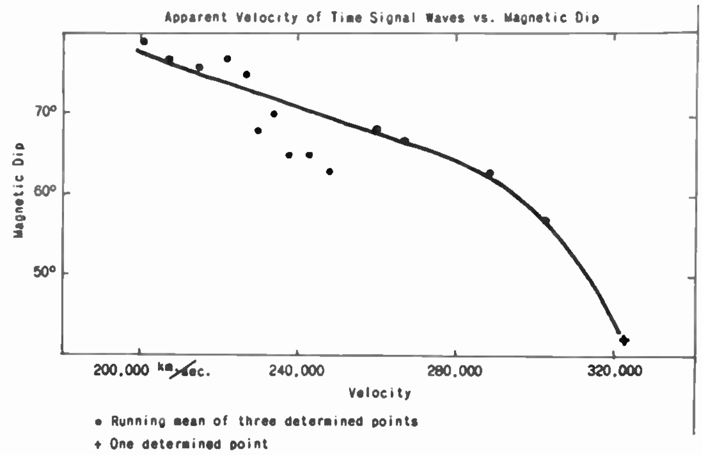


Figure 7

France, as compared with England, the more northern route may be subject to greater irregularities on ac ount of its higher mean magnetic latitude. This has led to a careful study of all available material which might show a relationship between the effective velocity of time signal waves and the routes over which they are presumably propagated.

From a study of some 20 different inter-observatory comparisons distributed throughout the world the mean effective velocity of propagation shows a striking correlation with the value of the horizontal intensity of the earth's magnetic field$^{(H)}$. The calculated velocities have been found to range from less than 200,000 km. per second for a value of 0.08 to a velocity of 300,000 per second, approximately that of light, for values of H exceeding 0.20. A corresponding study of propagation velocities and the values of magnetic dip give consistent results, the highest velocity of 300,000 km. per second corresponding to a dip of 61° and the lowest velocity that of 200,000 km. for a dip of 83°. The relationship between the apparent velocities and both H and dip are represented in the accompanying graphs. The scattering of individual points is probably due chiefly to the fact that the great circle route between the points intercompared in many cases does not correspond to the actual path pursued by the radio wave.

From the foregoing discussion, it appears that the relation of cosmic phenomena to radio communication is to become a subject of increasing importance. That definite cycles in certain cosmic phenomena such as solar activity, the period of solar rotation, and the period of the moon's revolution about the earth, exert their effects upon the characteristics of radio communication appears to be established upon reasonable evidence. As we are able to discover the underlying cause of these cycles in cosmic phenomena we may actually hope in the future to predict with some degree of accuracy the performance of radio communication over various paths at various frequencies through coming years. Difficulties now encountered in communication, due to cosmic causes, may be overcome through cyclical changes in assigned frequencies when a more thorough understanding of cosmic phenomena shall make evident the cure for the present maladies.

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PROCEEDINGS
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Volume 13 — October, 1936 — No. 3

IGNITION DISTURBANCES

BY

LESLIE F. CURTIS*

**Delivered before the Radio Club of America
September 16, 1936**

INTRODUCTION

This paper is a revision of one presented before the Institute of Radio Engineers at Rochester, New York, in November 1933, and explains in detail the origin of interference in the circuits of a typical battery ignition system. It supplements the author's paper "Electrical Interference in Motor Car Receivers," PROC. I.R.E., 20, p.674; April, 1932.

In discussing the suppression or reduction of ignition disturbances or interference in the operation of radio receivers it seems desirable to summarize what little is known or surmised about the nature of these disturbances and their sources. While theoretical analysis unsupported by experimental evidence will not, in general, be found a complete basis on which to proceed in the development of satisfactory means for the elimination of ignition interference, the lack in the literature known to the writer of any detailed treatment of the problem seems to him to justify the analysis which is here given.

GENERAL CONSIDERATIONS

It has been customary to analyze the response of any tuned antenna system to interference or strays in terms of the selectivity of the system and its acceptance of a band of frequencies falling within the side bands reproduced by the complete receiver.

In the analysis of the response to a spark transmitter, either as a desired or as an interfering signal, where the individual wave trains follow each other at very short time intervals, this procedure yields useful results.

However, with ignition interference, the ignition sparks occur at such relatively infrequent intervals that the disturbance from any one ignition cycle has disappeared before the following one begins.

In the case of an eight-cylinder motor running at 3600 r.p.m., only 240 ignition sparks per second are required, each one of which persists for only a few microseconds. Thus the disturbance, which gives rise to the radio interference, occurs within an extremely short portion of the ignition cycle, often in the form of a single discreet impulse and is thus wholly unlike the conventional spark transmitter.

With respect to the radio receiver itself, the time constant of the tuned antenna circuit may be of the order of 10×10^{-6} sec. Thus at the end of one ignition cycle, $\frac{1}{240}$ sec., any disturbance is reduced to $\epsilon^{-\frac{10^6}{10 \times 240}}$ times its initial value, or, for all practical purposes, zero.

It is reasonable to assume that no auxiliary car circuit, such as a resonant wire, rod, or frame, has any greater time constant than the above. Therefore no circuit is likely to be present in the field of the disturbance which will sustain oscillations until the next spark occurs. Obviously, at low engine speeds, the possibility of overlapping of the effects of successive sparks is even further avoided.

In this case it is simpler to deal with the individual damped wave trains than to resolve them into equivalent carrier and side bands.

I shall therefore discuss the theoretical nature of the wave fronts produced by a typical battery-ignition cycle, and then summarize the logical methods of reducing the response in the receiver to these impulses.

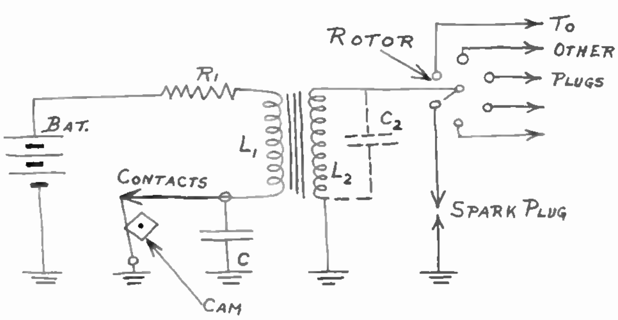

FIG. 1.

Figure 1 shows the portions of a battery ignition circuit which are usually considered in a superficial explanation of its operation. The constants which determine the interference transients include the obvious ones shown in the figure plus the distributed and lumped constants of all the wiring and the voltage-cur-

*Chief Engineer, United American Bosch Corp.

TABLE I

PHASE	TRANSITION	CIRCUIT CONDITION	DIAGRAM	PARAMETERS	CURVES	CAR-RADIO INTERFERENCE	DISTANCE INTERFERENCE
	Contacts close	Zero current	Fig. 1.			No	No
1		Primary current increases	Fig. 1.	Fixed		No	No
	Contacts start to open		Fig. 2.		Fig. 3. Fig. 5.	No	No
2		Voltage increases across contacts	Fig. 2.	Variable K	Fig. 3. Fig. 5.	No	No
	Contact spark		Fig. 2.		Fig. 5.	No	No
3		Current transferred to condenser	Fig. 2.	Variable arc	Fig. 4. Fig. 5.	No	No
	Spark out	Steep L-T wave front	Fig. 2.		Fig. 5.	?	No
4		L-F oscillation	Fig. 6. Fig. 8.	Fixed	Fig. 7.	No	No
	Rotor gap breaks down	Rectangular H-T wave front	Fig. 8.		Fig. 9.	Yes	?
5		L-F oscillation H-F oscillation	Fig. 8.	Fixed	Fig. 9. Fig. 10.	Yes	?
	Repetition of above	Rectangular H-T wave front	Fig. 8.		Fig. 9. Fig. 10.	Yes	?
5a		L-F oscillation H-F oscillation	Fig. 8.	Fixed	Fig. 9. Fig. 10.	Yes	?
	Plug gap breaks down	Rectangular H-T wave front	Fig. 8.		Fig. 10.	Great	?
6		H-F oscillation H-T arc at plug	Fig. 8. Fig. 12	Variable arc	Fig. 11.	Great	Yes
	H-T arc goes out	Steep H-T wave front	Fig. 12.		Fig. 11.	?	No
7		H-F oscillation L-F oscillation	Fig. 6.	Fixed	End of Fig. 7.	?	No
	Contacts close						

NOTE:- L-T Low-tension
H-T High-tension
L-F Low-Frequency
H-F High-Frequency

rent characteristic of any arc or spark. These will be discussed in detail for each portion of the complete ignition cycle.

The analysis of the ignition cycle may be divided into separate periods or phases which follow each other in definite sequence. During two of these phases, the complete differential equations expressing the phenomena involve variable parameters, and a general solution is not practical. The results may be approximated qualitatively, however, by segregating portions of the circuit and by assuming that the audio, radio, and super-radio frequency components are confined to relatively simple parts of the complex circuit.

Having developed an approximate qualitative analysis, we may then estimate the magnitude of the interfering components, - at least to the extent of predicting the most troublesome ones.

IGNITION PHASES

In each complete ignition cycle there are eight distinct phases, between which the circuit parameters change abruptly. During these phases the parameters may also change slowly, as will be described. At each abrupt change the approximate analysis is transferred to a different portion of the circuit. Table I summarizes these changes.

PHASE I

At the start of and during phase 1 no interference is produced since the primary current is then building up logarithmically at a relatively slow rate, limited by the inductance and the resistance of the primary circuit alone.

The other phases may give rise to radio interference and will, therefore, be discussed in some detail.

PHASE 2

Precisely how long phase 2 persists is rather uncertain since we have insufficient data on the voltage necessary to strike a spark across an exceedingly small gap thereby initiating phase 3. The explanation which is offered is, therefore, largely speculative, but leads to definite conclusions with regard to the wave fronts associated with the primary condenser circuit.

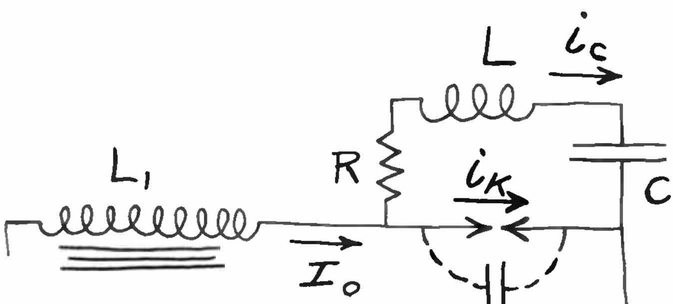

FIG. 2.

During phase 2 the current carried by the primary of the coil may be assumed to be constant, since it does not have time to change. Attention is, therefore, confined to that portion of the low-tension circuit between the coil and ground as indicated in Figure 2.

At the instant the contacts begin to separate, there is practically no voltage across them and before an arc or spark can exist, the voltage must be built up to a critical value. At the same instant all the coil current flows into the capacitance of the contacts and their mountings, as a displacement current, since the inherent inductance of the condenser circuit prevents an instantaneous building up of current in it.

If we could have geometrically perfect contacts 0.4 Cm. in diameter and separated by 0.0001 Cm., their capacitance would be approximately 100 mmf. As the gap is never uniform down to these small dimensions, the actual capacitance is appreciably less, but still large enough to be a factor in this study.

From the geometry of typical low-tension condenser circuits, it is estimated that their inductance may be between 20×10^{-9} and 500×10^{-9} henry. Let it be assumed that it is 100×10^{-9} henry.

If the contacts were instantaneously opened until their capacitance was 100 mmf. and thereafter remained fixed, the voltage across them would follow a damped radio-frequency wave of angular velocity

$$\omega \cong \frac{1}{\sqrt{100 \times 100 \times 10^{-21}}} = 316 \times 10^{6} \text{ radians/sec.}$$

and $f \cong 50 \times 10^{6}$ cycles/sec.

During this time the low-tension condenser would receive no appreciable charge.

Actually, however, the capacitance of the contacts varies from infinity, when they are closed, to a negligibly small value, when they are opened an amount sufficient to prevent a spark, so that the oscillation starts at a relatively low frequency and rapidly but smoothly increases to a very much higher frequency than indicated.

The initial rate of separation of the contacts will vary from 2.0 to 25.0 Cm./sec., depending on the engine speed and the cam shape. The capacitance between the contacts will, therefore, vary inversely as these rates and inversely with the time measured from the beginning of the opening.

For a complete determination of the wave forms involved, a step by step method is indicated, integrating over increments of time (Δt) during which the contact capacitance is assumed constant.

The terminology used is as follows:

- R effective radio-frequency resistance of condenser circuit.
- L effective radio-frequency inductance of condenser circuit.
- C effective radio-frequency capacitance of condenser circuit.
- K variable capacitance of contacts and mountings.
- I_o coil current.
- i_c condenser current.
- i_k displacement current to contacts.
- t time.
- ω angular frequency.
- α attenuation constant.

The differential equation for the circuit is then:

$$\frac{1}{K}\int (I_o - i_c)dt = Ri_c + L\frac{di_c}{dt} + \frac{1}{C}\int i_c dt$$

or

$$\frac{I_o}{K} = R\frac{di_c}{dt} + L\frac{d^2 i_c}{dt^2} + \left(\frac{1}{C} + \frac{1}{K}\right)i_c$$

Integrating, the current in the condenser circuit is:

$$i_c = \frac{I_o C}{C + K} + \varepsilon^{-\alpha t}(A_1 \sin \omega_1 t + B_1 \cos \omega_1 t)\Big]_0^{\Delta t}$$

where $\alpha = \dfrac{R}{2L}$

and

$$\omega_1 = \frac{\sqrt{\frac{4L(C + K)}{CK} - R^2}}{2L} \cong \frac{\sqrt{4LK - R^2K^2}}{2LK}$$

and A_1 and B_1 are constants of integration. A new zero of time is taken for each integration.

The displacement current across the contacts is $I_o - i_c$ or

$$i_K = \frac{I_o K}{C + K} - \varepsilon^{-\alpha t}(A_1 \sin \omega_1 t + B_1 \cos \omega_1 t)\Big]_0^{\Delta t}$$

The voltage across the contacts is $e_K = \Sigma \dfrac{i_K \Delta t}{K}$

These currents and the contact voltage, until a spark passes, may be expected to approximate the form indicated in Figure 3.

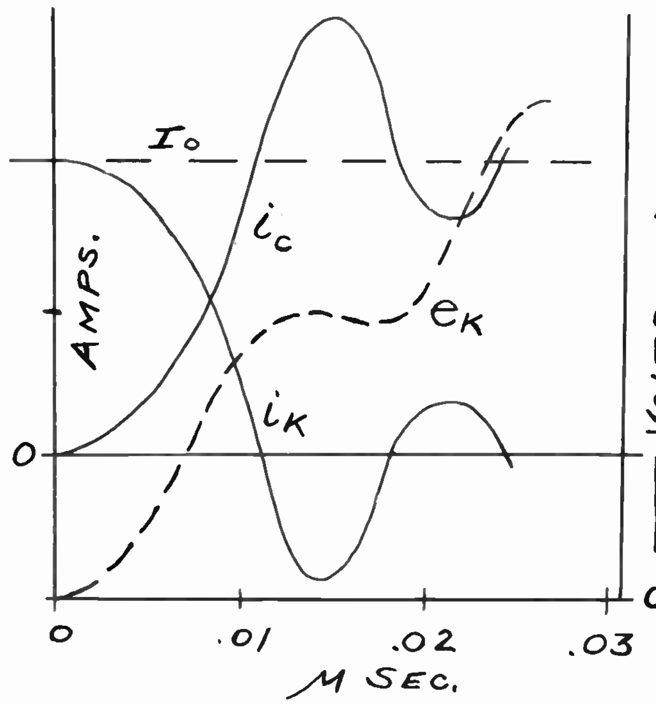

FIG. 3.

The interference, if any, produced by phase 2 will be discussed with that of phase 3, since it is of the same nature.

As is sometimes the case, if the voltage across the contacts does not build up to the critical breakdown value, depending on their separation, no primary spark occurs at all. Then the complete transfer of the current to the condenser circuit is accomplished as indicated in the figure. The total time required may be of the order of 0.02 to 0.05 microsecond. The major portion of the current is transferred in a fraction of this time with an exceedingly steep wave front. The following damped oscillation is at ultra-radio frequency.

Almost always, particularly at low engine speeds and with non-abrupt cams, the voltage wins in its race with cam opening, and a spark passes between the contacts. Phase 2 is then terminated and phase 3 begins.

PHASE 3

During phase 3, the contacts have usually opened to such an extent that the contact capacitance is negligible, particularly in view of the fact that the voltage required to sustain an arc is less than that to strike it. The voltage across the contacts is then determined by the arc characteristics of the contact material. For tungsten, which is ordinarily used, Anderson and Kretchmar*, give the voltage, current and gap relations as in Figure 4.

*University of Washington Research Papers.

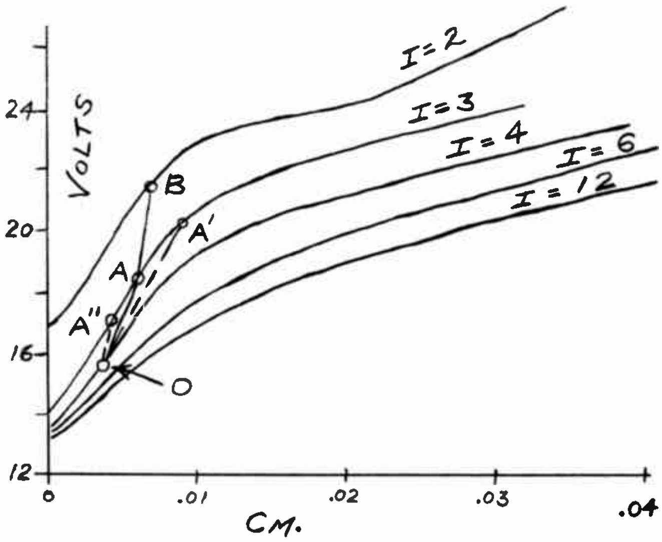

FIG. 4.

At the start of phase 3 the contacts are already open a small amount and a small current is flowing in the condenser circuit from phase 2. At a fixed engine speed the contact opening is approximately proportional to the time. We may, therefore, utilize Figure 4 drawn to a time scale for the applied voltage. The family of curves may also be identified by condenser currents rather than arc currents, the sum of the two being equal to the coil current.

When the arc is struck we are at some point 0 on the diagram. At a time Δt later we may be at point A. The voltage has increased at a rate $\frac{\Delta E}{\Delta t}$.

The differential equation for a step by step solution of the condenser current is

$$e_K = E + \frac{\Delta E}{\Delta t}\cdot t = Ri_c + L\frac{di_c}{dt} + \frac{1}{C}\int i_c dt$$

We may take increments of Δt small enough so that $\frac{\Delta E}{\Delta t}$ is constant and the solution for the current in the condenser is $i_c = \left(\frac{\Delta E}{\Delta t}\right)C + \varepsilon^{-\alpha t}(A_2 \sin \omega_2 t + B_2 \cos \omega_2 t)\Big]_0^{\Delta t}$

where α is as before, and $\omega_2 = \frac{\sqrt{4LC - R^2C^2}}{2LC}$

and A_2 and B_2 are the constants of integration.

Observing the initial conditions we may then solve for the current at the assumed time. If the calculated current change is not as assumed, we must correct our diagram by trial and error, but we may finally make an estimate of the proper change in voltage and plot it against the correct time on a curve.

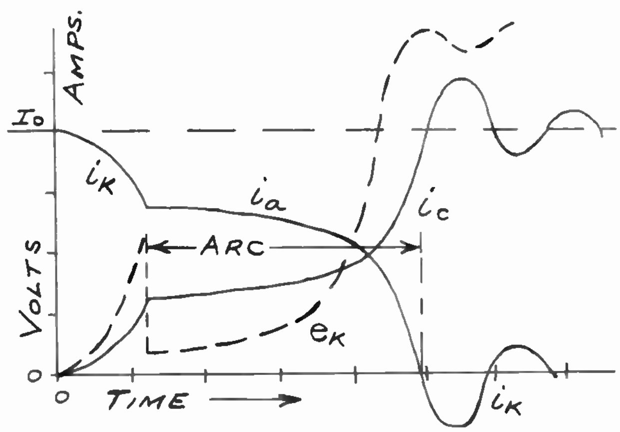

FIG. 5.

The complete solution is tedious but it is readily seen by inspection that, as the condenser current and contact voltage increase, they do so at increasingly rapid rates.

The voltage required to sustain an arc of less than 1.75 amp. is in excess of 100 volts.

Therefore, the current in the arc is sucked out and the voltage rises steeply. When this has occured the circuit oscillates as in phase 2 at a very high frequency determined by the now small distributed capacitance of the contacts and the inductance of the condenser circuit. This is rapidly damped and the total primary current is finally established in the condenser circuit.

An attempt to represent (not to scale) the complete spark and pre-spark phases is given in Figure 5.

All of this lengthy explanation is preliminary to the statement that whether a contact spark occurs or not, the coil current is finally transferred from the contacts to the low-tension condenser circuit with a very steep wave front. Any oscillations are at high radio frequencies, certainly well above the standard broadcast band. (0.5 to 1.6 M.C.)

The voltage wave front is also steep, but since the contacts are usually well shielded, and the maximum voltage is only a few hundred volts, this seldom gives rise to interference.

If, however, the condenser is not located within the distributor, the magnetic coupling of the condenser circuit to other circuits in the car may produce disturbing voltages as the result of the rapid change in current (say 0 to 4 amps.). The whole arc phase may cover a total time of perhaps 0.001 sec., but the most of the current is transferred during the last few microseconds. The voltage induced in any other circuit is, of course, $e = m\frac{di_c}{dt}$ where m is the mutual inductance of the condenser-contact loop and the circuit in question.

PHASE 4

During phase 2 and 3 the coil current was assumed constant. For phase 4 the effective circuit is as in Figure 6.

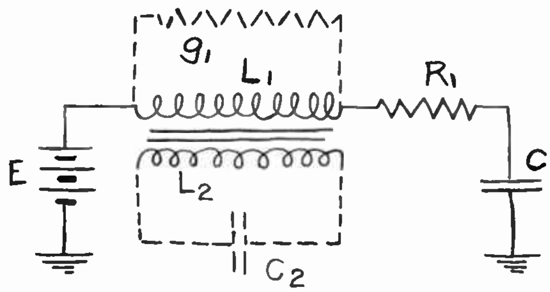

FIG. 6.

The primary current follows a logarithmically damped oscillation at an audio rate depending on the primary constants and those reflected from the open secondary thru a very high percentage of coupling between the windings. Figure 7 represents the case where no secondary spark occurs.

The frequency of this oscillation is usually between 2000 and 4000 cycles per second. The damping depends mainly on the coil losses other than the ohmic resistance of the winding.

If the secondary lead to the distributor were open, the secondary voltage would ordinarily rise to 30,000 or 40,000 volts, although at high cam speeds, which limit the time for the primary current to build up, it is often considerably less. The effect of even this high voltage is usually neg-

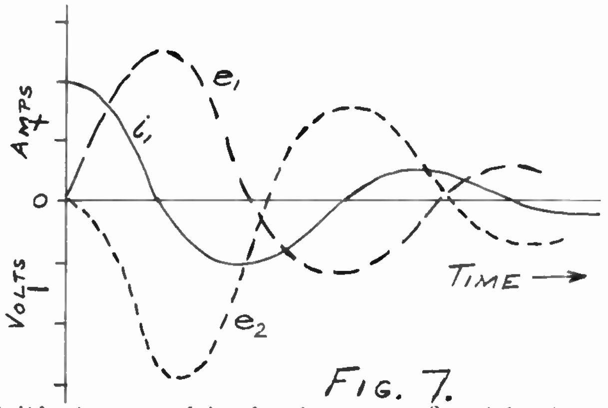

FIG. 7.

ligible because of its low frequency. Certainly there is no radiation at this frequency. If the antenna were coupled electrostatically to the high-tension terminal of the coil, there might be a slight response due to the induced transient at the tuned frequency of the first circuit in the receiver. This is rather a remote possibility.

There is probably no interference from magnetic coupling to the current in the primary circuit.

PHASE 5

The portion of the circuit to be considered during phase 5 is shown in Figure 8. It includes the constants of the high-tension cables, rotor gap, and spark plug. Actually the cable constants may be distributed, but as a first approximation they are here considered lumped.

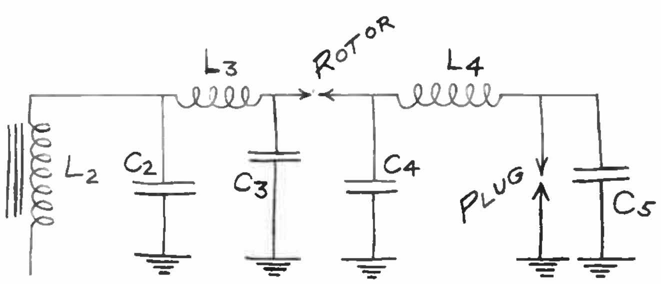

FIG. 8.

As the voltage rises at the secondary terminal of the coil L_2, capacitances C_2 and C_3 are charged at an audio rate with no appreciable drop across the cable inductance L_3. When the voltage reaches the breakdown value for the rotor gap, (depending on its length), a spark passes, partly discharging C_3 and charging C_4 to the same voltage. Except for the slight energy loss in the spark itself this would be instantaneous.

The circuits on both sides of the rotor gap respond to damped oscillations until the voltages are uniformly distributed at each end of the cables L_3 and L_4. The wave forms of the voltages across these several capacitances are shown in Figure 9 with corresponding subscripts.

Had the lead inductances and capacitances been uniformly distributed we should have had traveling waves along both of the cables with reflections at the terminals. During these oscillations both coil and spark-plug cables would be effectively open at each end, except for the slight lumping of capacitance at the coil terminals. Thus the lowest frequency at which the individual cables would oscillate would be approximately $f_1 = \frac{1}{2 l_o \sqrt{LC}}$ where l_o is the length corresponding to the formula for the familiar half-wave oscillation. In addition to this fundamental, all its harmonics, both odd and even, would also be present in decreasing proportion.

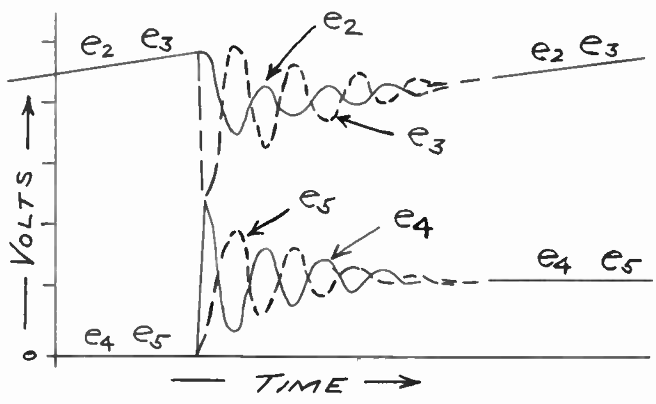

FIG. 9.

The capacitance and inductance of open high-tension leads spaced an average distance of two inches from ground are estimated to be

$$C = 4 \times 10^{-12} \text{ f/ft. and}$$
$$L = 300 \times 10^{-9} \text{ h/ft. respectively.}$$

The lowest frequency of oscillation in cable of this sort is

$$f_1 = \frac{450 \times 10^6}{l_o}$$

The cable connected to the coil would probably oscillate at a lower frequency because of the added capacitance of the secondary winding. The cable between the spark plug and the rotor would act practically by itself, with minor irregularities.

The capacitance and inductance of shielded 7mm. high-tension leads are estimated to be

$$C = 30 \times 10^{-12} \text{ f/ft. and}$$
$$L = 130 \times 10^{-9} \text{ h/ft. (ungrounded) or}$$
$$L = 65 \times 10^{-9} \text{ h/ft. (grounded)}$$

The fundamental frequencies for these conditions would be slightly lowered but not decreased in magnitude. In fact, because of the increased storage of energy in the larger cable capacitance the resulting spark would be "fatter".

As soon as the difference in voltage between e_3 and e_4 is again sufficient to break down the rotor gap, another spark passes. These cycles are continued until the voltage e_5 is large enough to jump the gap at the spark-plug electrodes. Phase 5 is then terminated and phase 6 begins.

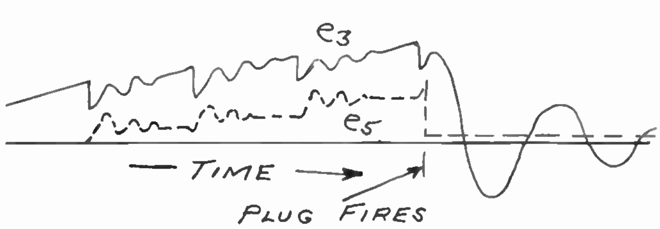

FIG. 10.

The voltages e_3 and e_5 (fundamental only) are indicated in Figure 10 for several successive rotor sparks terminated by plug breakdown.

PHASE 6

The spark at the plug almost short circuits it so that the voltage at its terminal drops almost instantaneously from, say, 6000 volts to only a few volts. During the first half cycle the rotor gap again breaks down and remains conducting, thereby making the circuit continuous from the plug to the coil.

A damped oscillation is now produced in the system which is much more violent than any of those preceeding it. Insofar as this oscillation is concerned, the cable is effectively shorted at the plug end and open at the coil end. Therefore, the frequencies present will be the fundamental

$f_1 = \frac{1}{4l_0\sqrt{LC}}$ and all of its odd harmonics.

This oscillation is accompanied by radiation from the length of the cable and also by the steep wave front at the plug which may be electrostatically coupled to radio circuits in the car.

These effects are by far the most serious of any during the ignition cycle. The radiation may directly affect an antenna on the car or at a distance. The steep wave front may produce oscillation in any auxiliary circuit on the car such as a choke rod, battery lead, poorly grounded hood, etc., to which it is coupled and which will radiate at the frequency to which it is tuned.

Suppose a circuit of lumped constants R, L and C to be coupled to the disturbance. The response is

$i = \varepsilon^{-\alpha t}(A \sin \omega t + B \cos \omega t)$

where $\alpha = \frac{R}{2L}$, $\omega = \sqrt{\frac{1}{LC} - \frac{R^2}{4L^2}}$

and A and B are constants of integration. These depend on the coupling as well as upon the circuit constants. The circuit oscillates at the one frequency to which it is tuned.

If the circuit constants are uniformly distributed, the response to the same wave front is

$$i = \Sigma_n \varepsilon^{-\alpha t}(A_n \sin n\omega t + B_n \cos n\omega t)$$

where n depends on the type of oscillation, that is, half or quarter wave, as fixed by open or grounded terminals. Such circuits may then radiate to antennas on the car or at a distance.

The response in the antenna circuit to *one* of the damped oscillations, either in the high-tension cable or in an auxiliary circuit, is

$i = \varepsilon^{-\alpha t}(A_n' \sin n\omega t + B_n' \cos n\omega t) + \varepsilon^{-\beta t}(M \sin st + N \cos st)$

where s is the angular frequency to which the receiver is tuned.

The response in the car antenna by virtue of direct coupling to the steep wave front is $i \cong \varepsilon^{-\alpha t} M' \sin st$

Both of the above expressions have components at the signal frequency to which the receiver will respond. The damped oscillation of the auxiliary circuit may not have been within the usual side band region of the receiver, but because of the transient character of the wave, antenna current at

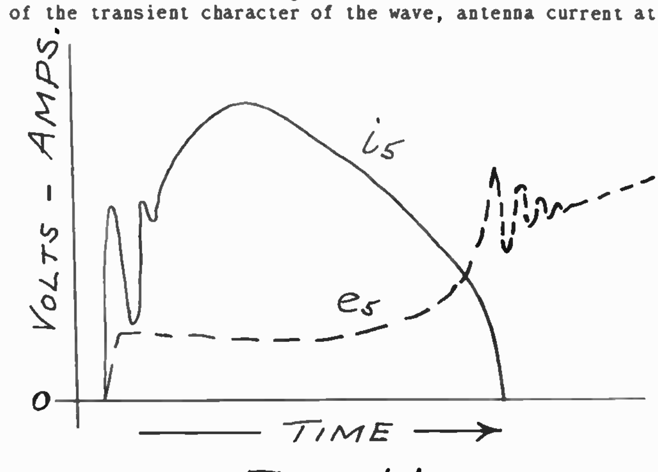

FIG. 11.

the signal frequency is produced. The audio-frequency output of the receiver is determined by the envelope of the signal-frequency wave and by the transient response of the loud speaker.

Before the oscillating current at the plug dies away the energy stored in the magnetic field of the coil and in the low-tension condenser establishes a comparatively steady arc current in the secondary circuit thru the plug to ground.

The rate of current rise is limited by the leakage inductance of the coil but it is essentially slow. Energy is then dissipated in all the circuits associated with this transfer and the current in the arc drops slowly and without interference components as indicated in the center portion of Figure 11.

However, the arc at the plug has a negative current-voltage characteristic, that is, the drop across the arc increases with decreasing current. The essential circuit diagram is shown in Figure 12.

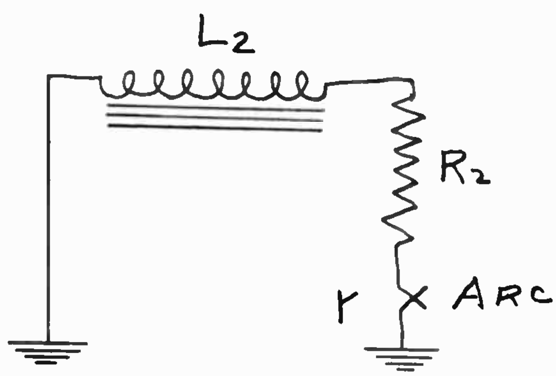

FIG. 12.

The secondary current dies out at an increasingly rapid rate as indicated in the latter part of Figure 11. The expression for the current is $i = I\varepsilon^{\frac{-R-r}{L}t}$ where r is the effective arc resistance which increases for small currents. Under certain plug conditions the current and voltage wave fronts are quite steep but it is very probable that the effects in the cable system are minor compared to those during the formation of the spark.

PHASE 7

When the high-tension arc goes out the cable is effectively open at plug, coil and rotor. These separate cable sections oscillate at the frequencies and with the damping found during rotor breakdown until the voltage is uniformly distributed along each section. However, since the wave front is not as steep as during breakdown, the effects can not be as great in magnitude.

After the arc goes out the remaining energy in the complete coil system produces a low-frequency damped oscillation of the same nature as in phase 4 but at lower values of current and voltage. Since these oscillations are at audio frequency, no further interference exists. Following phase 7, phase 1 is repeated.

SUMMARY OF INTERFERENCE SOURCES

Table I summarizes the interference phases discussed and gives references to the circuit diagrams and curves. The effects are important in the following order:-

1. Quarter-wave oscillations along the complete high-tension cable producing radiation.
2. A rectangular wave front of voltage at the spark plug which may induce oscillations in car antenna or auxiliary circuits coupled thereto.

3. Half-wave oscillations along the separate plug and coil cables producing radiation.
4. A rectangular wave front of voltage at the rotor gap which may induce oscillations in other circuits.
5. A steep wave front of current in the low-tension condenser circuit.

IGNITION INTERFERENCE SUPPRESSION

The most common method of reducing ignition interference is, of course, the use of spark-plug and rotor resistor type "suppressors". These resistances slow up the initial discharge of capacitances associated with the gaps and thereby reduce but do not eliminate the steep wave fronts. They also reduce the value of the gap currents and more or less effectively eliminate any oscillations in the high-tension cables.

If the car antenna or an auxiliary circuit is electrostatically coupled to the voltage at the spark plug, the suppressor is not a cure-all. If the suppressor has appreciable capacitance, either between its own terminals or to ground, there is still a sharp voltage wave front when the plug fires and an additional possibility of interference from the impulse discharge of the capacitance limited only by the low resistance of the arc. The resulting high current may be magnetically coupled to other circuits on the car.

The interference may be kept low by reducing the coupling from the sources of the disturbances to all other circuits and semi-isolated masses of metal. It is not sufficient to prevent direct coupling to the antenna. Inadequately grounded shielding of high-tension or car antenna cables may act as auxiliary circuits and should be avoided.

By-pass condensers with short sturdy leads may be used for effectively grounding circuits which can not be directly grounded.

The best method of interference reduction is the segregation of circuits. If all the high-tension leads and the distributor could be placed under a shield, well spaced from the leads, and well grounded at all its joints, radiation and electrostatic coupling would both be eliminated.

Some progress has been made in the use of band-pass antenna filters designed to prevent all voltages outside the desired band of frequencies from reaching the tuned antenna coil of the receiver. These filters do not eliminate components already existing in the pass band but reduce to some extent the impulse excitation of the first tuned circuit in the receiver.

The satisfactory reduction of ignition interference obviously depends on the proximity of the antenna to the car and the sensitivity of the receiver. Suppressors are said to be completely effective for reducing the radiation to conventionally located antennas for home receivers. Many standard-band receivers have been installed on cars having carefully located ignition systems where at the most only a single suppressor in the distributor lead was required. However, these same cars may have been bad radiators in the ultra short-wave bands. The final outcome, if all classes of service are to be protected, is up to the car designer.

DISCUSSION

As was indicated briefly by Mr. Curtis, the methods for reduction of the interference with radio reception commonly used in automotive ignition apparatus might or might not serve usefully to suppress interference in other frequency bands than the American broadcast band. Discussion on this point has been provided by Mr. H. A. Wheeler and Mr. L. C. F. Horle both of whom reported the making of measurements of the tendency toward interference of automotive ignition equipment in the high frequency and ultra-high frequency bands up to 70 megacycles. In both of the reported investigations in the high-frequency bands, the measurement of the tendency toward interference was made by the determination of the receiver operating conditions required to give a known peak voltage within the receiver circuits when the antenna of the radio system was placed within the interference field closely adjacent to the automobile. In the case of the data reported by Mr. Wheeler, which is reproduced in Figure 13, the values there given are determined from the degree of amplification in the radio receiver required to give a certain fixed peak noise output voltage across the speaker coil of the measuring receiver; while the data reported by Mr. Horle relates the high-frequency amplifier grid bias and hence, something closely approximating the logarithm of the amplification of the measuring receiver to the fre-

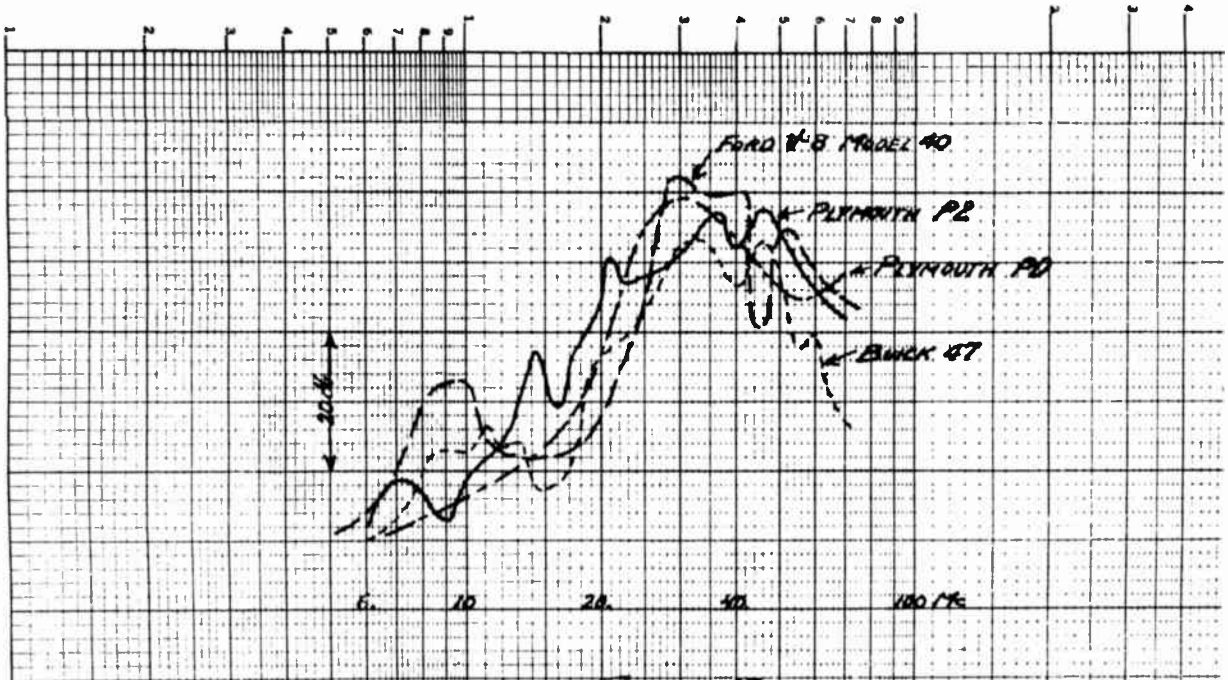

Figure 13

quency of measurement for a given peak noise output at the high-frequency amplifier output terminals. This data is shown in Figure 14. It is of interest to note in both cases that the interfering field intensities appear to be far higher in the high-frequency band than in the usual broadcasting band. This, it is believed, follows quite logically from Mr. Curtis' analysis of the phenomena operating in conventional ignition systems, and, most importantly, points to the need for especially careful consideration of whatever suppression means are to be employed if the frequencies above the normal broadcasting band are to be protected from ignition interference.

It will be observed from the Figures 13 and 14 how definitely the high frequency interference may be expected to operate against satisfactory radio reception in those bands commonly used for international broadcasting. It will be noted also how relatively great is likely to be the interference experienced in the 30 to 42 megacycle band within which already considerable aural broadcasting is being done in the United States and in the upper reaches of which doubtless far more will be done in the next few years.

While no spacific data was offered to the point, it was reported that some, but limited, experience indicates that the use of conventional resistance suppressors at the spark plug terminals and distributor terminals, while being very effective for the elimination of interference within the broadcast band, serves most indifferently, if at all, for the elimination of the interference in the higher frequency bands; thus, further emphasizing the need for the careful analysis of ignition circuits, such as are suggested in Mr. Curtis' paper, in order that the tendency toward interference may be reduced at its source, and the possible need also for the development of suppression devices which will be effective in the high-frequency bands as well as the low-frequency bands.

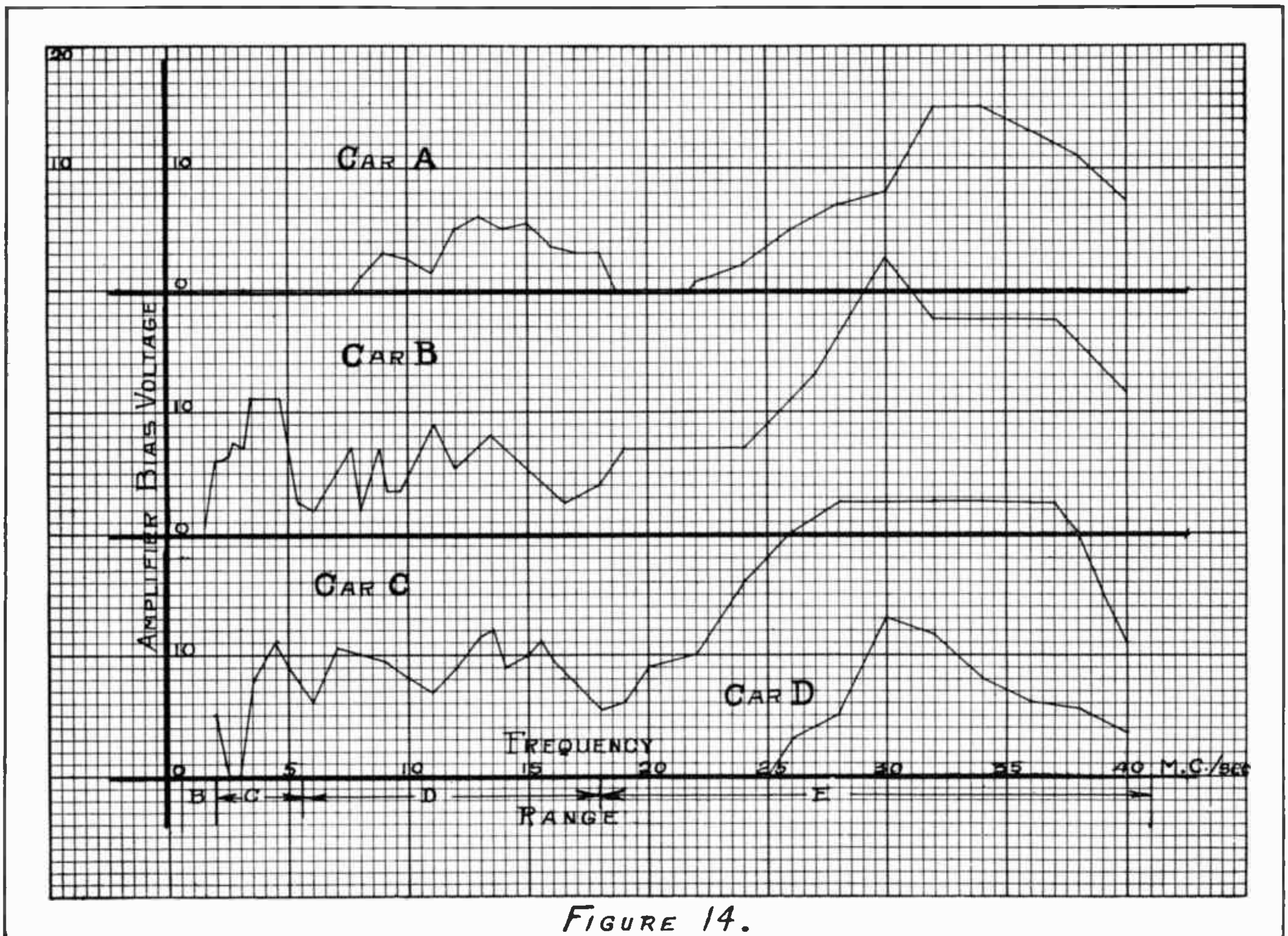

FIGURE 14.

BOOK REVIEW

MAKING A LIVING IN RADIO

by

Zeh Bouck

If "Making A Living In Radio" contributes nothing else than a removal of a veil of hallucination enshrouding the field of radio, it is worthy of publication. Radio is shown to be a cold, hard, practical field of a highly competitive nature in which the bonanza fruits of its early inception are bygones and wherein only fruits of difficult attainment remain - as in other stabilized, highly competitive industries.

Though the treatment is necessarily brief, the phases of servicing, operating, engineering, administration, sales, entertainment and writing are discussed with a cold-blooded and calculating introspection. The feeling of circumspection which it leaves with the reader is such as to discourage any except those who are intensely and genuinely interested.

Throughout the book the author stresses the fact that, as in any such competitive field, the need for as good an education as personal circumstances will permit is an ever growing one - regardless of what phase of radio the reader may contemplate choosing for his vocation. Certainly if one will heed its "road-markers" of circumspection, he will have traveled a long way towards "keeping his feet on the ground".

Lloyd West.

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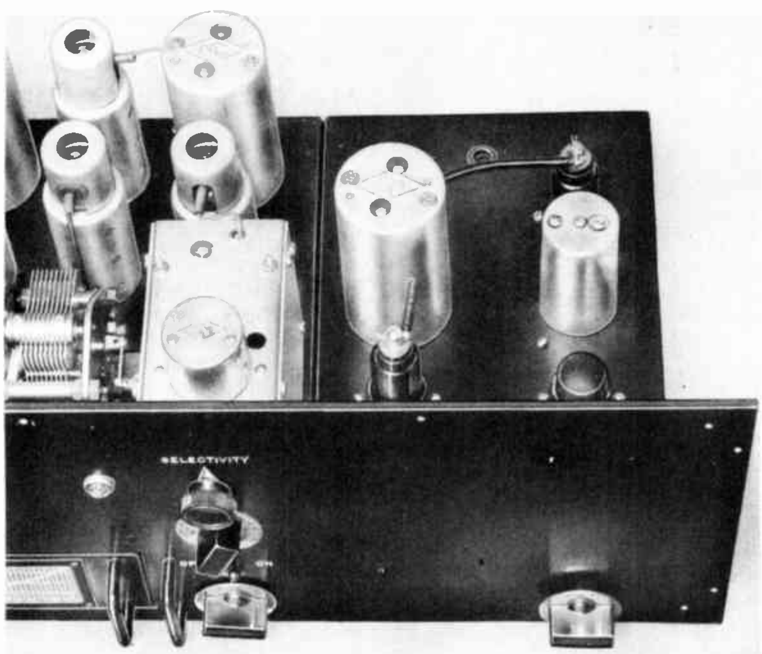
SOME ASPECTS OF INTERFERENCE AND NOISE REDUCTION IN COMMUNICATION TYPE RECEIVERS

BY

JAMES J. LAMB*

Delivered before the Radio Club of America
May 26, 1936

Only one of the fantastic features of radio communication in its present state is the continual compounding of complications that it inflicts upon itself both as the price of and as the result of progress. Each new technical advance that we make for the purpose of increasing the reliability and economy of operation of existing facilities through the reduction of interference discloses new possibilities which attract more users and give rise to further expansion and thus increase the load on the available frequency spectrum and, of course, threaten worse interference than before. This, then, requires further technical advances which result in more uses and more interferences - and so around the circle. And, while we are thus happily chasing ourselves in our own private vicious circle, the advances of civilization in other fields of endeavor are busily concocting new gadgets capable of bigger and better noise interference, such as the newest type of electric razor which has such excellant coverage -- all-wave and a block wide -- and likely to go off at hours of the day or night all out of keeping with normal shaving schedules. Worse yet, we are not content with complicating life for ourselves in our own field but we also must show outsiders how to use radio frequency equipment for purposes other than communication, with the result that we find ourselves being ungratefully bitten by diathermy "shadows" and the like.



Top view of the silencer section of the receiver using the circuit shown in Figure 9 on Page 35.

Viewed from this possibly pessimistic point of view, it might seem that technical progress in our chosen field is in the nature of a perpetual penance laid upon us, and that we must needs run as fast as we can in a modern sort of wonderland just to stay where we are. The way of the first pioneer may be hard; but sometimes we almost envy the young Marconi in that he had only to break through natural static, while man-made "static" was then still many years in the future.

But then if there were no complications such as these, perfection would have been too easily obtained and there would be no need to-day for the army of engineers, amateurs, experimenters and research workers still striving for further progress. And that would make the lot of many of us most unfortunate.

The problems of interference from man-made signal and noise interference concentrate in the amateur bands in a fashion which needs no detailed description. Hence some of the practical aids which have been applied in amateur communication receivers should be of general interest.

The general problem being interference, whether

*Technical Editor, QST

from undesired signals or noise, the attempts at solution take the form of improvement in selectivity; that is, selectivity in the broad sense of discrimination against everything but the desired signal. While selectivity is ordinarily considered as related only to the frequency characteristic of the receiver, in this instance it will be considered also in relation to the amplitude and phase characteristics of the receiving system. It may be permissible to distinguish three forms of selectivity: Frequency selectivity, amplitude selectivity and phase selectivity.

FREQUENCY SELECTIVITY

Perhaps the most widely used device for obtaining controllable high selectivity in communication receivers is the quartz crystal filter, used in the intermediate amplifier of superheterodyne receivers of the single-signal type. Two types of crystal-filter circuits are in general use, one of fixed sharpness of resonance with controllable symmetry, and the other of variable sharpness of resonance, also with controllable symmetry. This latter variable band-width type, which is adaptable to both c.w. telegraph and 'phone reception,, will be discussed here.

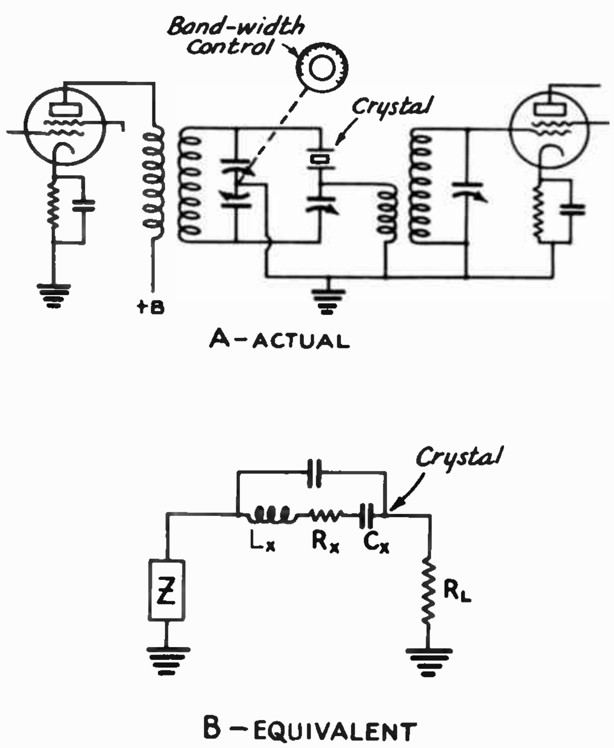
HOW CRYSTAL FILTERS WORK

Figure 1 illustrates the actual and equivalent circuits of a typical variable-selectivity quartz-crystal filter.[1] The crystal resonator is connected in a bridge circuit as shown in A, a two-section symmetric condenser forming two arms of the bridge, in parallel with a variable condenser which is used for adjustable tuning of the secondary of the input transformer. This arrangement gives an impedance stepdown of approximately 4 to 1 at the input. The primary of this transformer has approximately three times the inductance of the secondary, to which it is closely coupled, and is untuned. The primary of the output transformer is of such inductance and coupling as to match the output impedance to the series-resonance resistance of the crystal which is approximately 2500 ohms.

The crystal and rejection condenser in series with it form the other two arms of the bridge, the crystal providing the coupling.

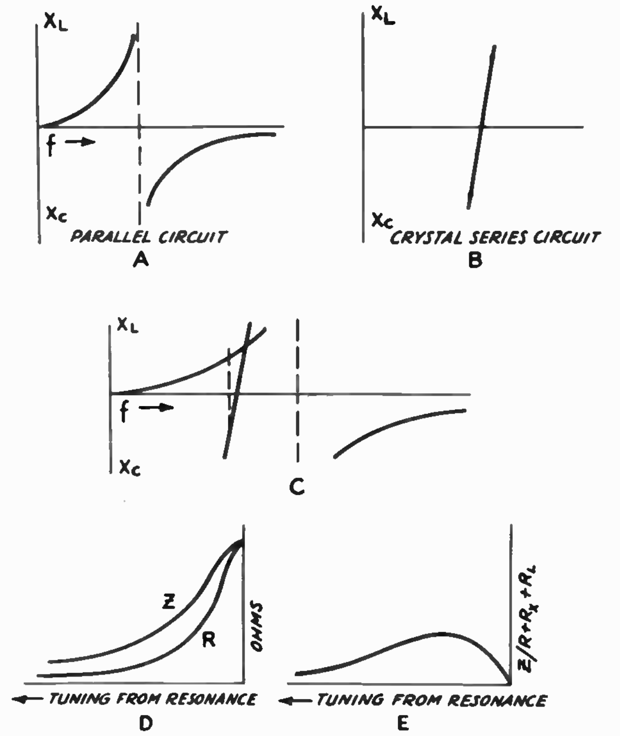
VARIABLE BAND-WIDTH ACTION

The equivalent series combination contains one half of the input circuit (across one section of the symmetric condenser) as well as the crystal and the primary of the output transformer. Series resonance occurs in this circuit when the capacitive and inductive reactances are equal. Reactance variation of R_L remains negligible over the range of operation, so that the resonant frequency of the complete circuit depends on the reactances of the crystal and Z. The parallel-tuned circuit, Z, is therefore the variable, tunable over a resonant frequency range near the crystal's frequency by adjustment of the band-width control. This means that the reactance curves for the parallel circuit, as shown in Figure 2, will be shifted along the frequency scale as the condenser is adjusted. Now, since the crystal has an extremely high inductance-capacitance ratio as a series circuit, its reactance curve is very steep, as shown in Figure 2-B. Hence the resonance frequency of the parallel circuit can be changed over a considerable range with but negligible effect on the resonance frequency of the complete circuit, as illustrated by the combined curves of A and B in Figure 2-C. With the reactance component of Z tuned out by the opposite reactance of the crystal, the variation in tuning of the parallel circuit by the band-width control will introduce, practically, only the varying resistance component



A, actual circuit of variable band-width crystal filter with adjustable rejection; B, equivalent circuit for illustrating variable band-width action.

Figure 1



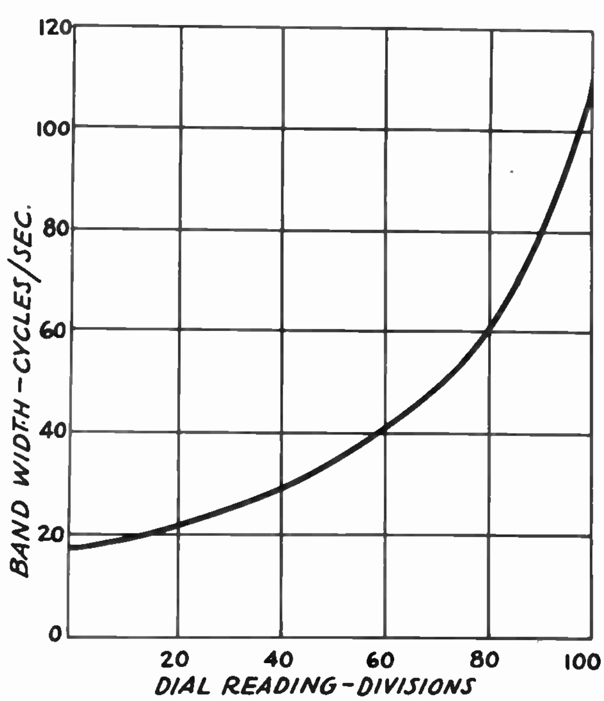
A, reactance curves of parallel-tuned circuit which is in series with the crystal; B, reactance curve of the series crystal; C, combined reactance curves of the parallel circuit and series crystal; D, impedance (Z) and resistance component (R) curves of the parallel circuit with tuning from resonance; E, variation in output voltage with tuning from resonance.

Figure 2

of parallel impedance in series with the crystal. The amount of this resistance determines the Q and, hence, the selectivity of the series circuit.

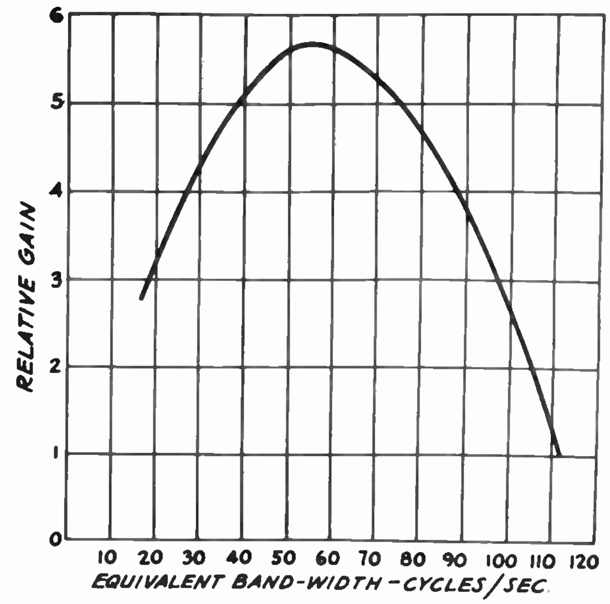
The voltage applied to the amplifier following the filter will depend on the current in the crystal series circuit. With this circuit resonant, the input voltage and series resistance will determine the current and, consequently, the output voltage. Now, both the input voltage and the series resistance are dependent on adjustment of the parallel-tuned circuit. Since the primary of the input transformer is not resonant, the voltage induced in series with the secondary will be comparatively constant over the small range required. Hence, the voltage applied to the series circuit, across the secondary (Z), depends on the secondary impedance. The impedance of the parallel circuit as it is detuned will change as shown by Curve Z of Figure 2-D, which curve also represents the voltage applied to the complete series circuit. Curve R of this figure illustrates the variation in the resistance component of the parallel impedance Z. The resonance current through the complete series circuit is dependent on the applied voltage and the total series resistance. This current, and hence the output voltage, will be represented by the ratio of Z to the total resistance of the series circuit and will vary with adjustment of the band-width control as illustrated in Figure 2-E.

It is evident that the maximum band-width (minimum selectivity) and minimum gain occur simultaneously with the input circuit tuned to resonance. An intermediate value of selectivity and maximum gain occur with the parallel circuit slightly detuned. This maximum gain condition (which occurs where the resistance and reactance components of the parallel circuit are approximately equal) is referred to as the adjustment for "optimum selectivity". Minimum band-width (maximum selectivity) and a lower value of gain occur with the parallel circuit detuned further from resonance. Experimental verification of the variation in selectivity by operation of the band-width control is shown is Figure 3. Variation in gain with selectivity is shown by the curve of Figure 4. Data for these curves were obtained by measurements on an early type single-signal receiver using this filter circuit.[2]



Experimental curve of the crystal-type S.S. receiver showing variation in band-width with tuning of the parallel circuit (Tuned to inductively reactive side of resonance).

Figure 3

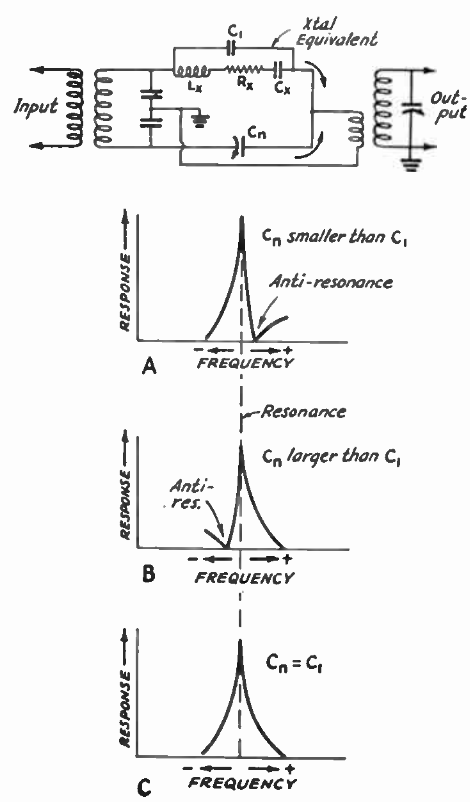


Variation in gain of the filter circuit with adjustment of the band-width control. (Cf E of Fig. 2)

Figure 4

REJECTION ACTION

As is well known from the equivalent circuit of the quartz crystal, the crystal is normally anti-resonant for a frequency approximately 1/2-percent higher than its resonant



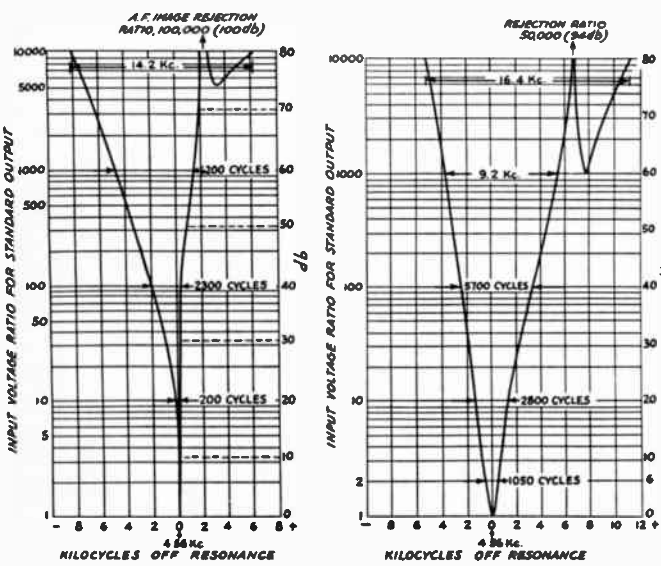
Illustrating the adjustable rejection action of the crystal filter, as used to eliminate heterodyne interference.

Figure 5

frequency. This results from the reactance of the shunt capacitance of the electrodes resonating with the inductive reactance of the crystal network at a frequency slightly above the latter's natural frequency. In the bridge arrangement of the crystal filter, this normal behavior is modified to shift the anti-resonant or rejection frequency to different values, both above and below resonance, within a limited range. The operation is illustrated by Figure 5.

The diagram of this figure shows the filter circuit with the crystal in its electrical equivalent form. Voltage is applied through the condenser C_n in anti-phase to the voltage operating on the crystal circuit. This will be recognized as similar to the neutralizing action for bridge circuits.

Now it might appear that C_n serves only to balance out C_1. However, in this instance C_n does not serve simply to neutralize the effect of the capacitance C_1, and thus to prevent unselective transmission past the crystal, but rather, as C_n is varied from minimum to larger capacitance the anti-phase voltage serves to make the effective shunting reactance of C_1 vary from its normal capacitive value, through zero, to a slightly minus capacitive value, when the effect is as if inductance were substituted for C_1. In the latter condition, the shunt reactance having changed sign, the complete crystal network is effectively in parallel resonance (or is anti-resonant) for a frequency below the crystal's natural frequency. Thus, while having maximum response to the desired-signal frequency, the circuit can be adjusted to reject an interfering signal having a carrier frequency in the range from several kilocycles above to nearly the same amount below crystal resonance. The rejection is most pronounced with the band-width control at optimum selectivity, but remains highly effective at minimum selectivity, as shown by the curves of Figure 6. These curves are made from measurements on a standard HRO receiver using this type of variable band-width filter.

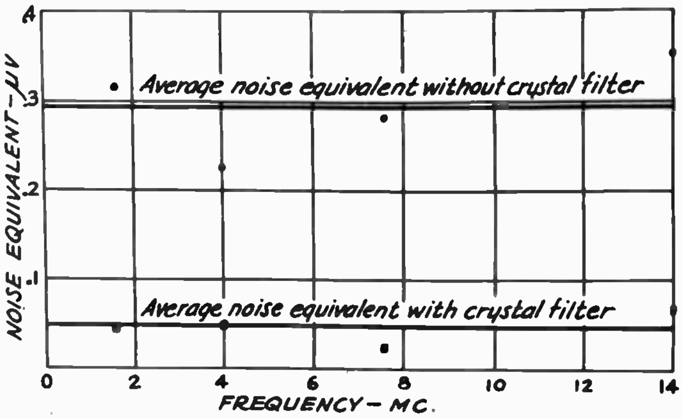


Maximum and minimum selectivity curves of a S. S. type receiver using the variable band-width crystal filter. Rejection adjustments are for two different interfering frequencies. Rejection is also effective on other side of resonance in both cases.

Figure 6

FREQUENCY SELECTIVITY AND NOISE

In addition to discriminating against undesired radio signals, the high-selectivity crystal filter also discriminates against noise, especially noise of the "hiss" type which consists of overlapping wave trains of noise pulses which are of amplitude comparable to that of the signal, or of smaller amplitude. As has been shown, particularly by V. D. Landon in a paper presented at the annual convention of I. R. E. in May, 1936, the peak and r.m.s. value of this type of noise varies as the square root of the band-width in a particular receiver. That is, the noise power is reduced in direct proportion to the reduction in band-width, and the effective voltage sensitivity of a receiver for c.w. signals is, therefore, increased as the square root of the ratio of reduction in band-width. Experimental verification of this improvement is shown in Figure 7, which is plotted from mea-



Improvement in receiver noise equivalent with increased selectivity. Straight superhet values are for an equivalent band-width of approximately 6600 cycles, crystal filter values for e.b.w. of approximately 50 cycles. (N. E. varies inversely as square root of e.b.w. ratio)

Figure 7

sured data taken on an early "single signal" type receiver, the noise being that of the receiver itself. The upper mean curve is for conventional superheterodyne selectivity with equivalent c.w. band-width of approximately 6600 cycles. The lower curve is for optimum crystal filter selectivity, the equivalent band-width being approximately 50 cycles. The ratio of improvement in effective voltage sensitivity is of the order of 20 db. At maximum selectivity of the filter (band-width approximately 20 cycles) the improvement would be approximately 50 db, while at the minimim filter selectivity (band-width of approximately 120 cycles) the improvement would be around 15 db.

AMPLITUDE SELECTIVITY

While the high-selectivity circuit discriminates against "hiss" type noise in the manner just described, the behavior of the receiver is markedly different under the influence of high-amplitude noise pulse excitation.

As has been pointed out by V. D. Landon in the paper referred to above, the ratio of peak to effective values for "hiss" noise voltage remains constant at a crest factor of approximately 3.4, regardless of the receiver bandwidth, both peak and effective values being reduced equally as the band-width becomes smaller. When, however, the noise excitation is of a staccato nature and the discrete noise pulses are of short duration as compared with the time separation of successive pulses so that the wave trains do not overlap, this peak-to-effective ratio or crest factor varies with band-width, being greater for large band-widths and becoming smaller as the band-width decreases. The explanation of this is, of course, that the individual wave trains generated within the receiver circuits by the noise pulses increase in duration and thus, the effective value increases relative to the peak value as the band-width is reduced through the improvement of circuit selectivity.

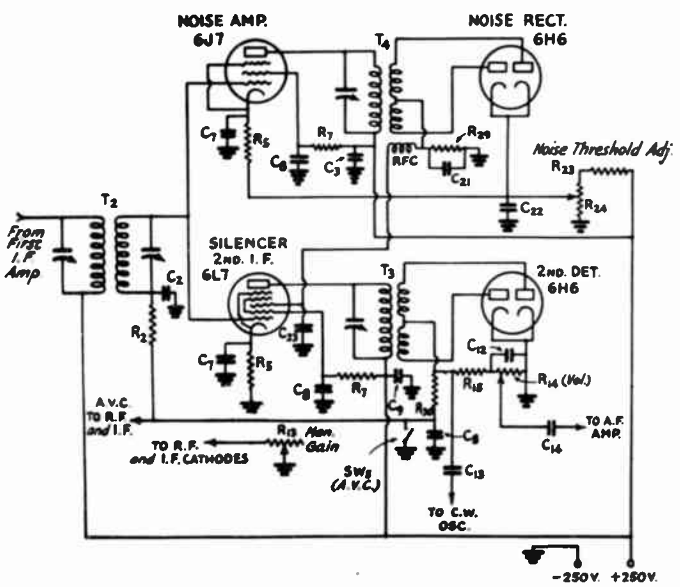
While high selectivity may perhaps be effective against this intermittent type of noise so long as it is of relatively small amplitude as compared to the signal, it becomes impotent when the action of the narrow-band filter circuit increases the duration of the individual wave trains and thus raises the effective value of the noise to a point where it becomes comparable with that of the desired signal. Incidentally, it is a happy fact that "hiss" noise voltage at the receiver's input circuit is generally of low amplitude, while high-amplitude noise is characteristically of the intermittent type.

Because of this, some means other than frequency selectivity must be employed in the case of noise pulses to bring down the effective value of the noise relation to the signal. And, in fact, the very characteristics of this type of noise suggest the method of its amelioration. Thus, since this type of noise is characterized by the fact that it occurs at relatively infrequent pulses - as compared with the frequency of other types of noise and as compared with the audio signal - the reduction of its effectiveness in interfering with the signal should result from its elimination through making inoperative the entire receiving system at the instant of its impingement. Indeed, it would be most effective to make the receiving system momentarily ineffective <u>just before</u> and during the impact of the noise pulse; and, indeed, this may be done by subjecting the signal to some delay in the receiving system while employing the noise pulse itself without delay in making the output portions of the receiving system inoperative for just sufficient duration as to wipe out the influence of the noise pulse. In practice, however, it has been found sufficient to provide for the effective reduction of the amplification of some one element of the radio receiver to substantially zero during all or part of the time during which the noise pulse would otherwise make itself troublesome.

For accomplishing this, two different circuit arrangements [3, 4] have been devised and are shown in Figures 8 and 9. In both of these the desired silencer action is obtained by providing several additional elements to an otherwise conventional superheterodyne circuit. Thus, a special noise amplifier stage is employed including its own noise rectifier which, in Figure 8, feeds a biasing voltage to the silencer tube which itself is in the chain of amplifier tubes between the second detector and the intermediate frequency amplifier. Normally it acts as a portion of the I. F. amplifier but, on the impressing of a noise pulse on the noise amplifier and rectifier, becomes momentarily inoperative and thus protects the second detector from the influence of the noise pulse. It is, of course, essential that the silencer operation occur for only such values of noise pulse amplitude as exceed the signal amplitude, lest the signal itself interfere with its own free transmission through the system of the receiver. To provide for this requirement the "Noise Threshold Adjustment" is provided in the form of a manually controllable bias on the noise amplifier and on the noise rectifier.

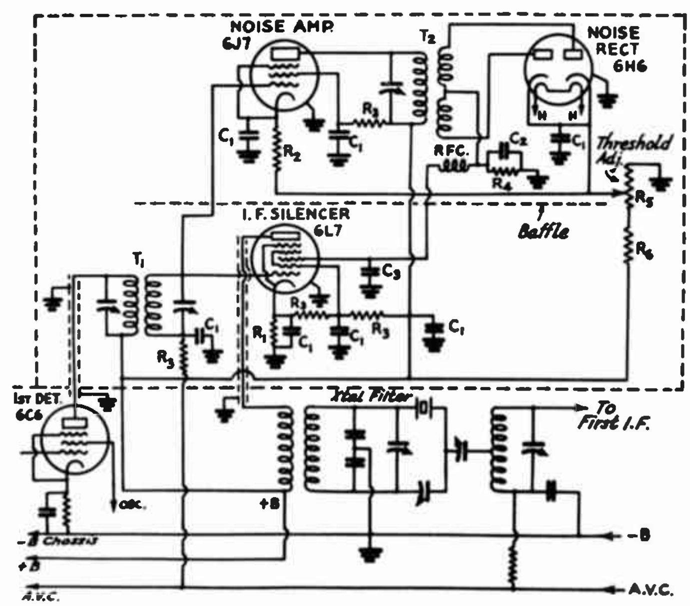
Such an arrangement as this provides very effectively for the reduction of the noise due to noise pulses of all kinds, such as result from the operation of other electrical equipment in the region of the radio receiving equipment. It is highly effective in connection with all radio telegraphic reception and for certain kinds of radio telephone reception but it must be admitted that for truly high fidelity radio transmission, its usefulness is markedly limited. Where, however, intelligibility - as in amateur and commercial non-public radio service - is the primary requirement it serves most effectively to convert transmissions which would be otherwise quite unuseable to perfectly useable transmissions; and thus it provides much of value to many types of radio communication.

The simultaneous use of these two methods of noise suppression, one useful in the reduction of the troublesome effectiveness of the hiss type of noises and the other useful in the reduction of the effectiveness of the pulse type of noise at the same time, suggests itself immediately and, indeed, has been found of great usefulness. It is, however, to be noted that in order that both expedients may be effective, it is essential that the silencer



Noise-silencing circuit applied to second i.f. stage of communication-type superhet receiver.

Figure 8



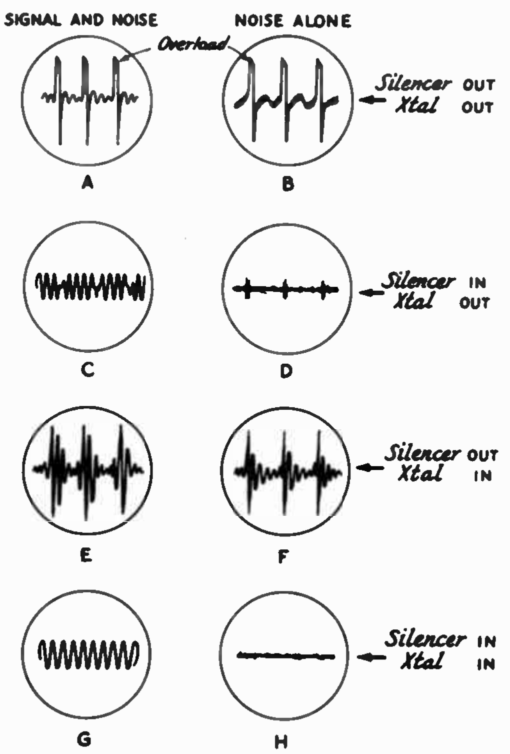
Silencing circuit applied between first detector and crystal filter of a S. S. type receiver.

Figure 9

arrangements precede the crystal filter in the chain of amplification comprising the radio receiver since, if the order is reversed, the noise pulse when impressed on the crystal filter will, thus, be converted from its original form of that of a pulse of high amplitude and short duration to one of long duration and only little decrement and thus give it something of the characteristics of the signal itself and make it highly effective in interfering with the signal. When used in the proper order, however, in which the silencer circuit wipes out the noise pulse before it can be offered to the crystal filter, a most effective combination results.

The circuits of such an arrangement are shown in Figure 9 in which the silencer circuits follow immediately on the output of the first detector of a conventional superheterodyne type of receiver and in which the output of the multifunction silencer-amplifier tube feeds the crystal filter directly.

The effectiveness of this combination of noise suppression arrangements can, of course, be best appreciated by listening to its operation in the reception of signals. It has, however, been found possible to show by oscillographic analysis the wave forms resulting from its operation and thus provide some visual evidence of its effectiveness. This is indicated by the wave form reproduced in Figure 10 of which the four traces shown on the left-hand column are those of the combined signal and noise under different conditions of noise suppression, while those shown in the four traces in the right hand column are those of the noise alone. Thus, in Figure 10, A and B, are shown the untreated noise and noise-signals which are characterized by the fact that the noise amplitude is not only so great as to vastly exceed the signal amplitude but so great as to cause actual overloading of the receiver circuits as indicated. In Figure 10D is shown the effect of the operation of the silencer circuits from which it will be evident how thoroughly effective these circuits are in reduction of the impulse type of noise. Figure 10C shows, similarly, how relatively free of the impulse type of noise is the signal as the result of the operation of the silencer circuits.



Oscillograms of c.w. beat note output obtained with the S. S. receiver using the silencer circuit ahead of the filter. (Beat-note c.w. reception with spark) interference.)

Figure 10

On the other hand, it will be noted from Figure 10, E and F, and their comparison with A and B, how markedly the crystal filter builds up the impulse noises so as to mask the signal completely. And in Figure 10G and H is shown the result of the operation of both the crystal filter and the silencer circuits. From these it will be noted that not only does the silencer circuit almost completely eliminate the influence of the impulse noise but the crystal filter does, to a surprising degree, fill in the "hole" in the wave form made by the operation of the silencer circuit.

EDITOR'S NOTE

At this point in Mr. Lamb's dissertation, an extended demonstration of the operation of his arrangements in the reception of long distance short wave signals was made. Locally generated interference was provided in the interest of having a constant and controllable source of noise; and reception under the several conditions of noise suppression, as indicated in Figure 10, was accomplished. It is to be reported that precisely as suggested by Mr. Lamb's oscillographs, the overpowering intense impulse type of noise was so far suppressed by the action of the silencer circuits as to make otherwise unintelligible telegraph signals easily and comfortably readable. Similarly, it was noted how the action of the crystal filter alone converted the staccato noise pulse into clear, rounded, bell-like tones that persisted for easily appreciable periods and thus even more completely masked the signal than did the original noise itself. And, finally, it was shown how the combination of the silencer circuits followed by the crystal filter resulted in signals which, as far as the ear of the auditor could tell, were completely free of all of the originally overpowering noises.

DIVERSITY RECEPTION

In addition to its well-known ability to mitigate fading effects, diversity reception also offers possibilities for improving effective selectivity in reception, particularly for reducing heterodyne interference in the reception of amplitude modulated signals. This may be considered as a species of phase selectivity, as has been pointed out by the mathematical analysis of J. Robinson ("The Elimination of Inter-Station Interference", Wireless Engineer, April, 1935). If two spaced antennas connected to separate receivers are used for the reception of amplitude modulated waves and the audio-frequency outputs of the two receivers are combined, the signal outputs from the two detectors will, in general, add up arithmetically, but beat-frequency heterodyne products from an interfering carrier will add up vectorally. Hence, heterodyne interference in even a simple diversity system will, at worst, be the same as in a single receiver and may be reduced to zero when the beat note outputs of the two final detectors are equal in amplitude and 180° different in phase. This phase relationship may be controlled by the spacing and the directivity of the antennas or perhaps by adjustable phase-shifting networks in the receiver proper.

While the writer has been occupied during the last year in the development of a diversity receiving system of this kind[5] and will doubtless have reports of its performance to make some time in the not too distant future, no specific report can be given at this time. It is felt, however, that this mode of interference reduction should be here given mention.

This brief discussion of methods for the reduction of noise and interference would not, however, be complete without some reference to at least one other method that seems to have tremendous possibilities. More specifically, it should be pointed out that in such methods as are here discussed, the problem which is faced is that of differentiating between the interfering noise and a signal which are largely identical in characteristics and thus give little latitude for differentiation: so small a latitude that minor operational improvements are viewed as major technical accomplishments. Indeed, as long as the nature of the signal and the noise continue to be so nearly identical, just so long will the solution of the noise suppression problem be exceedingly difficult of attainment. There are then, two general directions in which further work in this direction may take: either the magnitude, nature of the noise, or both, may be so modified as to improve the operation of radio systems; or the magnitude, nature of the signal, or both, must be so modified as to bring about improved operation. Much work is being done by many in all of these directions.

Those who are associated with the sources of interference, i.e., the electrical and automotive industries, have become keenly conscious of their responsibility in this matter and are achieving some success in the elimination of many sources of noise interference with radio.

The broadcast radio receiver manufacturers are recognizing the fact that carelessly installed antennas can be expected to do little more than to invite noise interference and are, therefore, educating the radio-listening public to the advantages to be found in elevated antennas connected to the receiving set through transmission lines isolated from the sources of noise interference.

The operators of broadcasting stations are, to such a degree as economic and regulational considerations make it possible, slowly but surely building the power level of broadcasting stations upward against the noise level which their signals encounter.

And, of course, such expedients as have been here described are contributing their share to the reduction of the effect of noise interference.

There is thus left one major direction of this work as yet little investigated and only recently brought to the general attention of the technicians in the radio field. That is, the possibilities of other types of radio transmission in the interest of providing for the easier and more effective differentiation between the noise and the signal. Outstanding in this direction is the recently announced work of Major E. H. Armstrong in the development of a practical system of frequency modulation.[6] The writer is as yet too unfamiliar with the work in this direction to justify an opinion on its ultimate possibilities, but he can't resist the desire to point out that perhaps, with the passing years, we will all come to view the work here discussed and reported as having been of far less importance than the mere facing of the fact that since we can do so relatively little in changing the nature of the noise to which our radio signals are subjected, we might best do all we can to change the nature of the signal.

Above all, engineers in our field must view new developments, no matter how radically they may transgress our preconceived notions, with open minds. If our present methods are actually obsolete, and amplitude-modulated transmissions are due ultimately to be thrown "out the window", we should be the first to realize the situation and get into action. Not to be the first to try, nor yet the last to adopt, may be an adage satisfactory to the philosophical way of thinking. But an obstinate engineer can never tell how long he can remain reactionary without missing the bus entirely.

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RADIO INTERFERENCE

BY

ALLEN W. HAWKINS*

Delivered before the Radio Club of America

April 9, 1936

"Radio Interference" is the term applied to all electrical radiations which interfere with radio reception. Radio interference can be classified in three groups: interference resulting from atmospheric and other natural causes; interference by radio transmitters; and interference arising from the operation of electrical equipment commonly termed "man-made static". It is the purpose of this paper to discuss the latter class, and to discuss it mainly from a stand-point of locating the sources of interference.

The sources of "man-made static" can be classified in two groups. The first group includes high voltage or high frequency apparatus such as x-ray equipment, diathermy machines and vacuum tube bombarders. The second group includes all other equipment giving rise to interference radiations through sparking and arcing.

The first class of interference sources such as x-ray apparatus need not be discussed at any length because that type of equipment is confined to relatively few areas and because of high levels of such interference produced by it the sources are easily located.

The second class of interference sources, those arising from the arcing and sparking can develope in any piece of electrical apparatus or circuit wiring. Thus unless we retire to the wide open spaces, we are surrounded at all times with potential sources of radio interference. It is no wonder, then, that most radio interference comes from some sparking wire, motor commutator, oscillating thermostat, or something else that makes an electrical arc. Since new types of appliances are constantly appearing on the market, there are always new sources of interference being called to the attention of the radio trouble-shooter.

Quite a large proportion of these appliances are not inherently producers of interference. It is only because of loosened contacts, improper adjustment, or faulty installation that they become troublesome to the radio listener.

It may be interesting to consider the actual figures of radio interference found in a large portion of the State of New Jersey. This summary contains only cases where complaints were made by individuals to the local power company. These figures are probably typical for any other geographical area of similar distribution of industrial, residential, and rural areas.

The total number of radio interference cases investigated in 1935 was 1990 distributed as to the source of interference, as follows:

Electric power lines and equipment	11%
Other Utilities (trolleys, telegraph, telephone, traffic lights, railroad signals, etc.)	3%
Radio set defects	21%
Complainants' wiring & appliances	17%
Other appliances in neighborhood	20%
Unknown responsibility (noise disappeared)	28%

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From these figures we see that in 72% of the cases the source of the interference was found and in the other 28% the interference disappeared before a location of the source could be made. So let us consider only the cases where the trouble was found.

Thus a new tabulation follows:

Electric power lines and equipment	16%
Other utilities	4%
Radio set defects	29%
Complainants' equipment	23%
Other appliances in neighborhood	28%

By adding the percentages of complaints due to radio set defects, 29%, to those due to complainants' household equipment, 23%, it is evident that the trouble-shooter will find the trouble in the complainant's establishment in about 52% of the cases he is called on to investigate. In any case, adding 28% for other appliances in the neighborhood, the source of the trouble will be in or near the complainant's house 70% of the time.

Thus, the radio trouble-shooter should visit the complainant's house first. If the interference is not in evidence at the time, he will be able to gain helpful information from the complainant regarding the disturbance. The trouble-shooter should try to ascertain the approximate time of day the annoyance occurs, wave bands covered, volume level and characteristics of the noise itself. From this information he will know when to come back, what sort of apparatus to bring with him and something about the character of the interference.

Now that the trouble-shooter is well started on his job, let us consider what sort of a problem he faces.

A spark discharge can, of itself, produce any and all frequencies of the radio spectrum. The circuits of conductors connected to it become the tank and antenna system. The resonant characteristics of these conductors determines the frequency or frequencies of interference. The area of disturbance as well as the frequency depends upon the size of this resonant network. Resonance of this network can be either broad or sharp although it is usually very broad. This network can be of almost any size since it may include building wiring and the vast extent of power lines. In spite of this threat to all radio reception in any large area the interference is usually limited to an area of a city block or less.

Interference usually leaves a building where it originates through the electric service wires, in spite of the fact that the iron service tube or pipe should be an excellent radio frequency shield. From the service wires the interference may pass along the street on the power lines and be radiated therefrom. It may pass through power transformers from secondary to primary or vice versa. The interference may be picked up on adjacent power circuits and carried to other areas. A noise originating on a power line is usually of high intensity and covers large areas.

In most cases the interference in power lines will be of highest volume level near the source and for some distance will be without apparent standing waves. At sufficient distance from the source, however, standing waves usually occur, tapering off gradually at considerable distances from the area of intense interference.

Now let us consider the method of locating the trouble.

The two most practical attacks on the problem are the "hot and cold method" and the "cut and try method".

In using the "hot and cold method", it is assumed that the interference is loudest near its source. The trouble-shooter simply makes a survey of the interference zone from the standpoint of intensity, the point of highest intensity indicating the location of the source.

The "cut and try method" is based on the theory that the interference will vanish when the electrical potential is taken off the source. In using this method, all electrical circuits in the area of interference are de-energized one at a time. When the circuit that feeds the source is de-energized, the noise will disappear. Parts of this circuit can then be de-energized one at a time to further localize the trouble. This method is carried on until the source itself is discovered.

The "cut and try method" is usually limited in use to one building because it is not expedient to interrupt the electric service to many consumers at once.

In the average case of radio interference both methods are used. The "hot and cold method" is used first to locate the source of interference within a comparatively small area, then the "cut and try method" is used to more definitely locate it.

When the trouble is apparently on an overhead power line, the vicinity of the interference is found by the "hot and cold method". Then a lineman is sent up to inspect the suspected poles. The greatest amount of noise interference will occur when he climbs the pole from which the noise originates.

When the trouble is found by the "hot and cold method" to be in a building, then the noise is located by de-energizing the building circuits one at a time.

These two methods usually suffice to locate sources of radio interference. In some cases, however, they fail. In these cases the noise may not be loudest at its source. De-energizing a circuit may make the noise disappear in certain locations, but only because the circuit network is detuned by the switching operation.

The radio trouble-shooter must be wary of these possibilities. He should recheck himself continually to avoid being led on a wild goose chase. Above all, the trouble-shooter should not allow himself to form any premature theories concerning the interference. These theories are apt to bias his observations to such an extent that his location of the trouble will be difficult.

Permit me my frankness when I say, that most radio engineers I have met have made incorrect judgments in cases of radio interference because they have formed premature theories.

The use of a loop antenna as a direction finder of the source of interference is rarely effective. The loop will simply indicate the direction of the nearest conductor of the interference. The loop antenna is usually used, however, because of its easy portability and to indicate what wires are carrying the troublesome interference.

The character of the noise cannot be depended on to identify the source, because a variety of sources can produce the same sort of interference. However, a thermostatic noise from a heating pad or a fish bowl heater can easily be recognized by its regular intermittant character. Motor commutation usually produces a cyclic noise which is recognizable.

By way of illustration let me take an actual case of radio trouble-shooting.

Leaving the complainant's house, we walk north along the street. The noise falls off in intensity so we walk back south, past the complainant's house. As we continue south the noise increases and then falls off. Then we return to the point of maximum intensity and use the loop antenna to pick out the electric service wire giving rise to the greatest intensity of noise. We enter the house to which the service is attached. The main switch is opened and the noise disappears. As the woman of the house is assuring us that no appliances are being used, the maid suddenly departs to the attic and the noise stops. When the maid returns, she acknowledges that she had left a heating pad operating in her bed. And thus another source of interference is discovered.

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FACTORS RELATING TO FAITHFUL REPRODUCTION

BY

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Delivered before the Radio Club of America
October 15, 1936

The subject of phonograph or radio reproduction is necessarily a very broad one. For this reason the following paper concerns itself only with aural compensation and volume expansion as aids in obtaining the desired result. Demonstrations will be given of various factors involved and a very useful piece of laboratory equipment in connection with this work will be described and demonstrated.

Before discussing aural compensation as applied to an electric phonograph or radio receiver, it would probably be well to digress a moment and consider the general subject of sound and hearing.

A large amount of research work has been carried on by different experimenters in connection with sound at different levels and its effect upon the human ear. General reference to books and papers on the subject gives one the impression that although some discrepancies exist between results obtained by different experimenters, there is, in the majority of cases, quite close agreement with the work of Dr. Harvey Fletcher of the Bell Laboratories. For this reason we have, in our work on aural compensation, used data presented by Dr. Fletcher on his paper "Loudness, Its Definition, Measurement and Calculation"[1].

Listening carefully to the radio set of a few years ago, we have all undoubtedly noticed that a reduction in volume from a rather loud level to one which could be used in the average apartment without causing annoyance to ones neighbors was accompanied by a change in musical balance. At the louder level the low frequency instruments such as the bass viol, bass drum and tuba were in proper balance with the middle and high frequency instruments. At the lower level, however, the low frequency instruments were no longer in balance and the music had a harsh or tinny characteristic. This effect could also be obtained at an outdoor orchestral performance or band concert if one were to change his position from one close to the instruments to a position a hundred feet removed from them. The high frequency instruments will decrease in volume much less rapidly than the tuba and other low frequency instruments. The reasons for this phenomenon are very clearly given in Dr. Fletcher's paper.

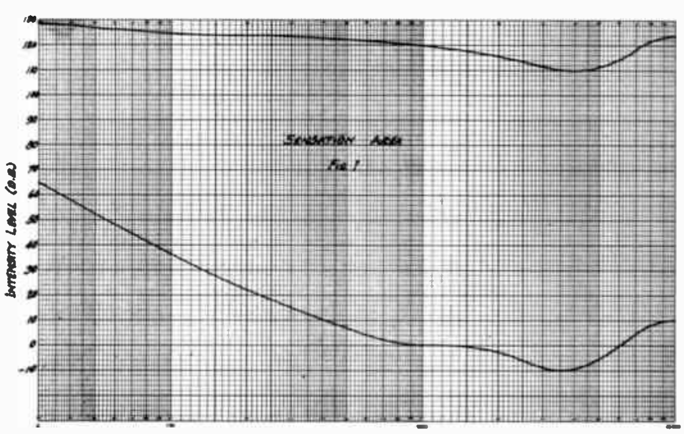

Figure #1 illustrates the frequency response and intensity range of the average human ear from the threshold of hearing to the threshold of feeling. These two thresholds were determined at different frequencies by means of pure tones applied to head phones worn by the observers. Many different readings were taken at various fixed frequencies and for many different observers. From these, it was possible to plot the lower curve which shows the threshold of hearing for the average individual. It will be noted that the curve is plotted with 1,000 cycles as the 0 d.b. level. For reference purposes the input level to the ear canal at this frequency was found to be 10^{-16} watts per square centimeter. It is easily seen by reference to the curve that for a frequency of 100 cycles, the same intensity level in the ear would require an input 35 d.b. higher than for 1,000 cycles. Conversely, the input in the 3000 - 4000 cycle range would need to be 10 d.b. less than that at 1000 cycles to produce equivalent effect. The upper curve was determined in much the same manner with the exception that the limiting factor became the point at which the observer felt actual pain. This curve, which is called the threshold of feeling, represents the highest level the average individual can withstand for

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any length of time without actually suffering pain as a result. It will be noticed in this case the input at 100 cycles need be only 5 d.b. higher than that at 1000 cycles to produce the same result. In other words at the threshold of feeling the curve of the average ear is essentially flat.

Between the two thresholds, it then became possible to plot a family of curves representing the average characteristic at different levels. For convenience these curves were determined in 10 d.b. steps at 1000 cycles beginning with the threshold of hearing and carrying up to the threshold of feeling. Figure #2 shows this family of

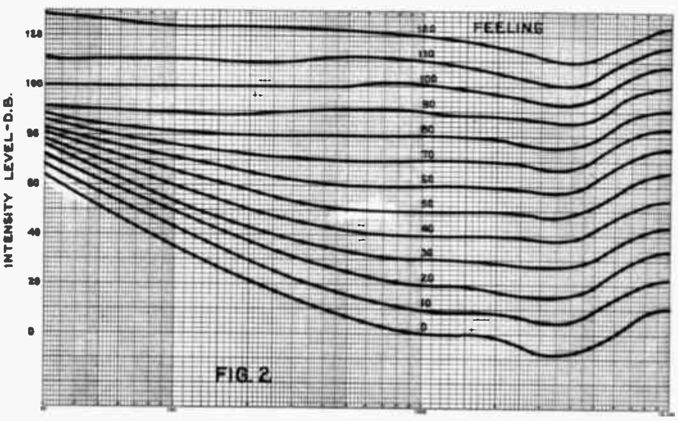
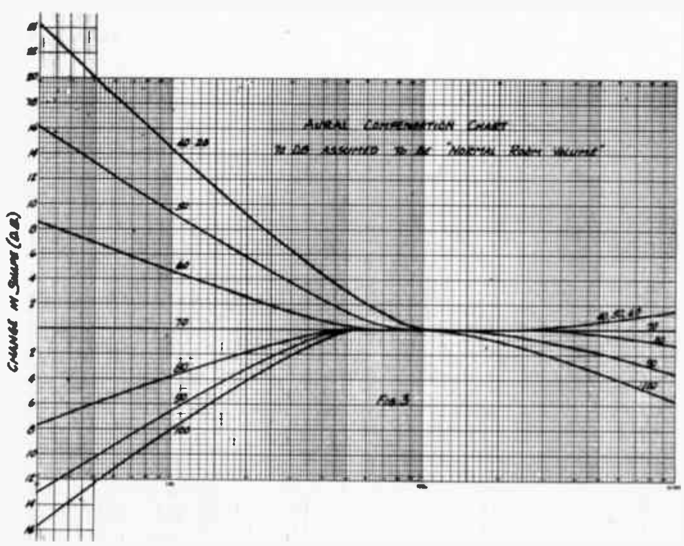


FIG. 2

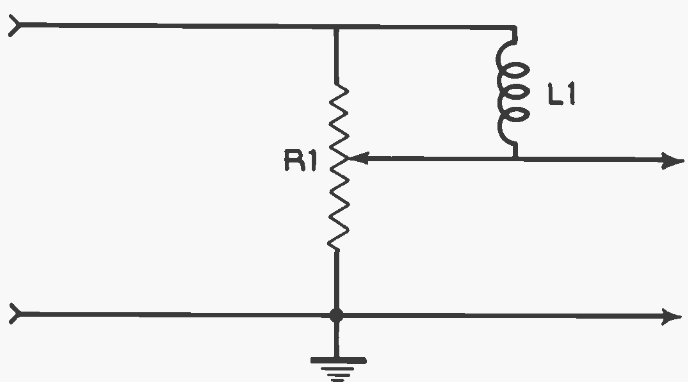
curves. An observation of these curves indicates immediately that the range between 500 and 5000 cycles is practically flat for almost any level but that for frequencies below 500 cycles, the input to the ear must be increased over the 1000 cycle level to maintain proper balance. These curves then show us that some form of low frequency compensation is necessary if we are to maintain proper balance over very wide changes in volume. It is, of course understood that the volume changes referred to are the changes in average level from one value to another rather than the normal variations in dynamic level which are automatically taken care of by the conductor of the orchestra.

It has been indicated by Dr. Fletcher's work and borne out by tests with calibrated volume controls in observers' homes, that the average listening level in a quiet location is approximately 70 d.b. above reference level or threshold of hearing. With this level as a basis and assuming that a desirable sound characteristic has been obtained at this level, it then becomes possible to plot a new series of curves from those shown in Figure #2 and which are shown in Figure #3. For convenience, the 70 d.b. level is shown as flat since we have already assumed that the subject level was satisfactory and thus any irregularities in the curve must be carried through to the other levels. From this family of curves, we can readily determine the amount of compensation necessary as the sound level is changed above and below the 70 d.b. average level. For instance, a decrease in average level of 10 d.b. at 1000 cycles must be compensated for by a decrease of only 5 d.b. in response at 100 cycles. Similarly, a decrease in average level of 20 d.b. at 1000 cycles must be compensated for by a decrease of only 10 d.b. at 100 cycles. The higher frequencies are not affected as much but it is desirable that the response be decreased only about 8 d.b. at 10,000 cycles for each 10 d.b. decrease in average level at 1000 cycles. Conversely, an increase of 10 d.b. in average level above the 70 d.b. level at 1000 cycles must be accompanied by an increase of only 6 d.b. in the 100 cycle level. An increase of only about 2 d.b. at 10,000 cycles should accompany this 10 d.b. increase in level. These curves may thus be used directly to determine the amount of compensation necessary at all levels above and below the average reference level of 70 d.b.

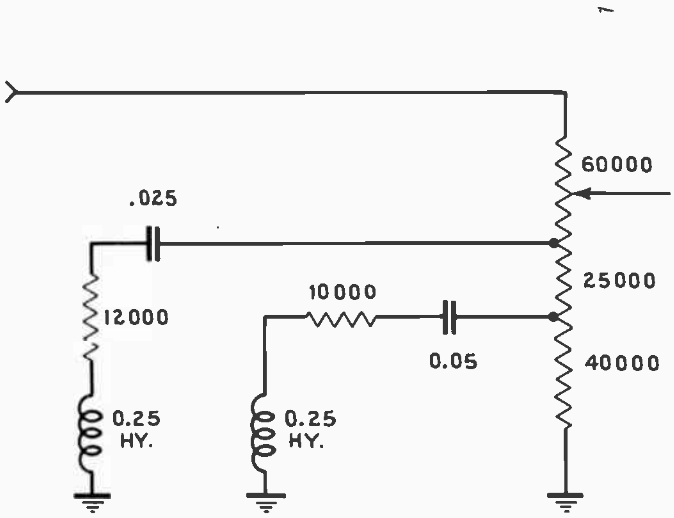


Having determined that some form of compensation is necessary, and with curves available showing the amount required for the different levels, we are faced with the requirement of finding ways and means of accomplishing the desired result.

One method which has been used in the past with good results on low impedance phonograph input systems is shown in Figure #4. In this, the volume control, which is of the potentiometer type, has an inductor L_1 connected between the slider and the high end of the volume control. The size of the inductor will, of course, be determined by the resistance of the volume control with which it is used. In the particular case illustrated, the value of L_1 was approximately 30 millihenries and the volume control resistance was 60 ohms. With these values, the shunting effect of L_1 is almost negligible at 1000 cycles. At 100 cycles, and at mid point on the volume control, the shunting effect is approximately 2 to 1 or in other words, we have a rising characteristic 6 d.b. higher at 100



SIMPLE COMPENSATION CIRCUIT
FIG. 4



TYPICAL COMPENSATED VOLUME CONTROL
FIG. 5

cycles than at 1000 cycles at mid volume setting. This method of compensation is not adequate for better grade phonographs or radio receivers, but has been used with considerable success on lower priced instruments.

The next method, shown in Figure #5, has been found quite adequate for present day requirements. It will be noted in observing the diagram that there are two fixed taps on the resistance element in addition to the slider. These taps are used in conjunction with the proper shunting networks to give the proper amount of low and high frequency compensation for the different levels. The value of total resistance used is of necessity determined by the particular application. The positions of the taps are determined by the gain of the amplifier following the control and bear such a relation to the total resistance that normal listening level occurs at approximately the first tap down from maximum volume setting taking into account the compensation networks. The diagram indicates the values used for one particular compensated control which is being used in the instrument for demonstration. The shunt networks may have values changed as required to give the particular curve desired at normal level as well as the degree of compensation needed to fulfill the curves of Figure #3. A smaller capacitor will have the effect of increasing the low frequency response with reference to 1000 cycles whereas a larger capacitor will cause the opposite effect, other values remaining constant. The control and its attendant circuits are designed such that the change in level between taps is 20 d.b. When the slider has reached the second tap, no further compensation occurs. A further improvement can be shown by the addition of a third tap to the control, but this can be applied only to the higher priced instruments. The amount of high frequency compensation is determined by the size of the inductors used. Increasing the size of the inductor increases the high frequency response and conversely decreasing its size decreases the amount of tip up at the high frequency and as the volume is decreased.

A third type of tone compensation, which has been used in the past, is the automatic bass compensator. In this circuit, an amplifier system of the variable gain type is employed in parallel with the regular audio channel. By virtue of its variable gain characteristic and due to the fact that its response is limited to frequencies below 1000 cycles, it can be made to increase the overall low frequency response as the volume level is decreased. Due to the added cost and nicety of balance required for operation, this particular system has not had wide application in receivers enjoying quantity production. A similar arrangement has been applied to the radio receivers for restricting the amount of high frequency reproduction with a decrease in antenna signal voltage.

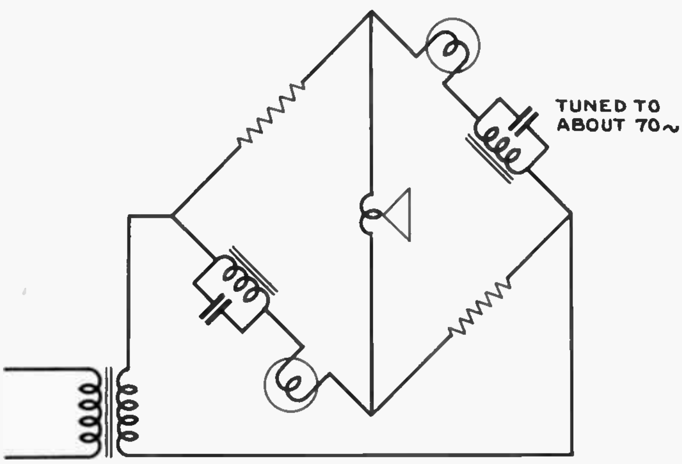
This covers briefly the subject of aural compensation and some possible methods of accomplishing the desired results. A short demonstration of this feature will follow as a means of showing aurally exactly what this means to you and me.

The next portion of this discussion will center around volume expansion as applied to both phonograph and radio reproduction. When volume expansion is mentioned to the average person, he immediately wants to know what advantage a set employing this system has over one which is not so equipped. To better understand the reasons for the existance of volume expansion, and to point out its advantages, let us consider changes in volume level of the orchestra, limitations in the reproduction of these changes as introduced by the phonograph or radio system and restoration of the original by means of an automatic volume expansion system.

By extensive tests, it has been well established that a large symphony orchestra has an available volume range of approximately 70 d.b. from a soft pianissimo to a heavy cresendo passage. If we were to try and cover this volume range on a phonograph record, we would immediately be faced with two definite limitations. The cresendo passage would, of necessity, require maximum movement of the cutter on the wax consistent with groove spacing and other mechanical limitations. With this as the top limit, the movement of the cutter on pianissimo passages would be so microscopic that when played back, the only apparent result would be surface noise or record scratch. Obviously, these passages must be brought above the surface noise level, in order to be heard and the cresendo passage must be limited as regard cutter travel, for the reasons outlined above. In doing this, compression of the 70 d.b. volume range occurs and we find a total volume range on latest records of approximately 50 d.b. When one considers that 20 d.b. difference represents a change in voltage or needle velocity of 10 to 1 and that 50 d.b. is 320:1 and 70 d.b. is 3200:1, I believe, it is readily apparent why the expression the director tried so hard to obtain during recording has been somewhat tempered by the time it is reproduced on an ordinary instrument. To illustrate by figures just what these limitations are mechanically on a phonograph record, let us consider that a recording is being made at about 100 grooves per inch. The maximum swing of the cutter point is thus limited to less than .010 inch if the grooves are not to touch each other. Assuming we could take this distance on loud passages, then the pianissimo passage, 70 d.b., lower than this would be approximately .000003 inch. This is much less than the microscopic structure of the record material itself and even an increase in the pianissimo passage of 20 d.b. still allows a swing of only .00003 inch. From these figures, I believe it is readily apparent that we are doing well to obtain the present volume range on the record alone.

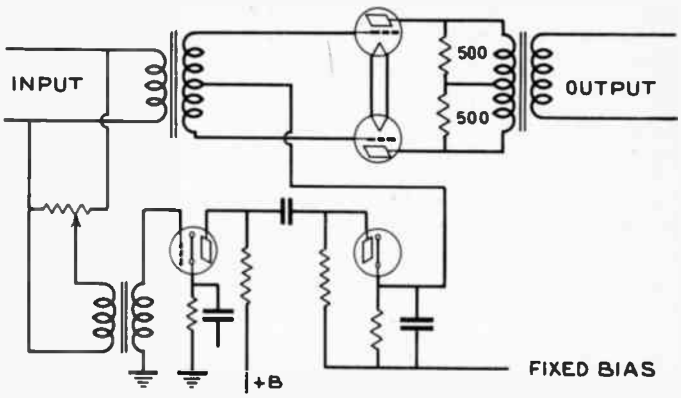
Limitations of a similar type are imposed by the broadcast station except for the fact that in this case line noises, hum and other extraneous disturbances require increasing the level on pianissimo passages and the danger of over modulation requires decreasing the level on cresendo passages. Present day high grade stations are able to cover a volume range of approximately 50 d.b. This is seldom used, however, since there is a definite desire on the part of the station owners and operators as well as program sponsors to obtain maximum listener coverage at all times by maintaining as high an average level of modulation as possible. Wide volume changes on present broadcast equipment would be satisfactory for people located close to the transmitter, but appreciation of it would be lacking by those located at some distance, particularly when the pianissimo passages were pushed aside by bursts of static.

The above discussion, I believe, indicates the desirability of obtaining some form of volume expansion as a means of restoring this loss of 20 d.b. in volume range particularly if full enjoyment of symphonic programs is to result. There are several ways of obtaining this result and the use of any one type is dependent upon the amount of expansion desired and cost permitted. With any of the systems which can be used at present, it is obvious that the result is a compromise since monitoring for all recording and broadcasting is accomplished manually and expansion in the phonograph or radio must take place automatically. Eventually, it may be possible to automatically compress the volume range during recording or broadcasting and so design the phonograph or receiver that its expansion curve is the counterpart of the compression curve. When this has been done, then the listener will be able to enjoy to a much greater extent, the various symphonic programs available. Until that time, however, we can obtain a great amount of pleasure even from the present volume expansion systems as applied to standard phonographs and broadcast receivers.



SYSTEM No.1
BRIDGE TYPE EXPANDER CIRCUIT
FIG.6

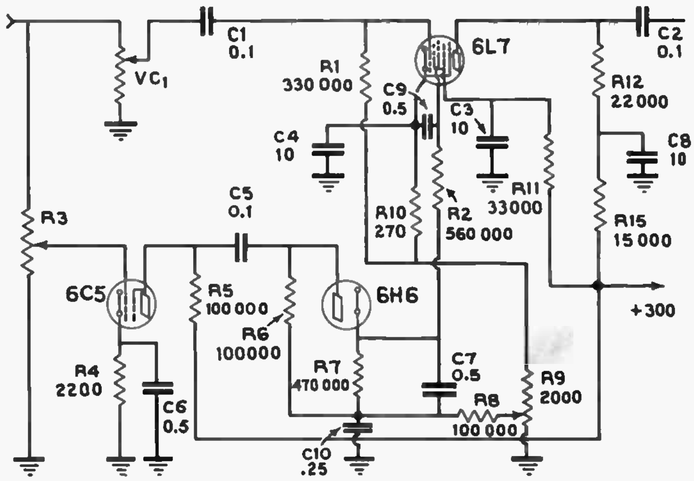
One relatively simple method, shown schematically in Figure #6, has been used the past year on a commercial broadcast receiver.[2] It employs low voltage electric lights of special design in a balanced bridge circuit across the output system. At low signal levels, the bridge is nearly in balance and very little signal gets through to the loudspeaker. As the audio signal increases, the resistance of the filaments in the lights changes and the bridge is thrown out of balance. This allows more than the direct increase in signal strength to be applied to the voice coil of the loudspeaker resulting in a degree of volume expansion dependent upon the strength of the applied signal. Earlier forms of this device imposed power output limitations upon the audio system but improvement in thermal resistance characteristics of the filaments used has resulted in a system capable of 15 d.b. expansion without serious distortion due to change in load impedance across the output tubes.



SYSTEM No.2
PUSH PULL TRIODE, VARIABLE IMPEDANCE
EXPANSION CIRCUIT
FIG.7

A second system, shown in Figure #7 consists of push-pull triodes having transformer coupling in both input and output circuits.[3] The plate circuit of the tubes, which normally calls for a load impedance of 20,000 ohms for proper matching, is shunted to approximately 1000 ohms by means of two 500 ohm resistors connected across the primaries of the coupling transformer. A variable bias system, dependent upon the strength of the incoming signal for its action, is connected to the common grid return of the triodes. In this type circuit, the transfer of energy between these tubes and the next audio stage is dependent upon the effective plate impedance of the tubes and the resistance of the load. Since the tube impedance is dependent upon the grid bias, there is a proportionally greater transfer of energy to the 500 ohm circuit at low grid bias than at high grid bias. In this way it is possible to use this circuit for volume expansion with very low distortion. Its main disadvantage, as far as commercial phonographs or broadcast receivers are concerned is its greater cost as compared with systems which do not require as many tubes and do not use transformer coupling.

The third system, shown schematically in Figure #8, has been used by RCA Victor for the past year for phonograph reproduction. Since this is the system to be demonstrated shortly, I believe it would be well to discuss the functions of the various parts. It will be noted that the incoming audio signal branches at VC_1. One branch



RCA VOLUME EXPANSION CIRCUIT
FIG. 8

goes to the #1 grid of the 6L7 variable gain amplifier tube and the other branch terminates in the degree of expansion control R_3. The 6L7 tube has previously been adjusted by means of variable resistor R_9 to operate on the proper portion of its characteristic for the amount of volume expansion and audio gain required. In the particular circuit, the tube gain is about 2.5 and the plate current is about 1 milliampere in the zero signal condition. As signal is applied and with R_3 set as maximum, amplification of the signal occurs in the 6C5 and rectification takes place in the 6H6. This rectified voltage, the value of which is determined by the strength of the incoming signal appears across R_7 and is thus impressed upon the #3 grid of the 6L7 through the time delay circuit composed of R_2 and C_9. The polarity of this voltage is opposite to that which is already present on the #3 grid and serves to reduce the effective voltage on this grid. This increases the voltage gain in the 6L7 and a variable gain amplifier results, the gain of which is entirely dependent upon the strength of the incoming signal. This system provides a relatively cheap volume expander capable of increasing the volume range 15-18 d.b. if so desired. The amount of expansion is easily varied by means of R_3 and in this way the desired result can be obtained on almost any type of phonograph or radio program.

A discussion of this subject must of necessity include some mention of the cabinet and its relation to the overall musical balance. The broadcast type cone loudspeaker radiates from both sides of the diaphragm. When this type speaker is mounted on a flat baffle and away from corners of the room or other cavities, very little low frequency resonance is present. When, however, the flat baffle is folded back to form a cabinet for the loudspeaker an immediate change takes place in the low frequency balance unless special precautions are taken. The cavity behind the loudspeaker cone serves directly in reinforcing the low frequency response and "boomy" reproduction results. To many people this type of reproduction is entirely pleasing and unless some "boom" is present they feel the set is not properly designed. To the music lover, however, this "boom" is highly objectionable since it is a type of musical balance, or unbalance, which never occurs in an orchestra.

There are many factors which govern the amount of cabinet resonance reproduced, among which the more important are: type of output system used; whether high or low impedance; frequency of resonance of cone suspension system; ruggedness and weight of wood used for the cabinet; depth of cabinet from speaker baffle to back opening and whether the back is open or closed. A brief discussion of each of these factors will enable us to understand more fully their direct effect upon reproduction.

If a high impedance output system is used, for instance one employing pentodes, changes in impedance of the plate load cause a proportional increase in voltage across the load due to the constant current characteristics of these tubes. A cone loudspeaker at its suspension resonance frequency presents a much higher impedance than at 400 cycles. For this reason it is desirable, if boominess is to be decreased, that the cone resonance be located below 70 cycles. Furthermore if a reduction in resonance voltage or output is desired at this frequency then a low impedance output system should be used. With either system the cone suspension resonance should never be located above 80 cycles in a console model since average cabinet resonance in this type cabinet occurs in the band between 100 and 150 cycles depending upon the cabinet depth.

Another factor directly connected with the amount of low frequency resonance effect is the weight of wood used for the cabinet. If thin woods are employed with very little bracing then at those frequencies where resonance occurs, or close to them, vibration of the cabinet sides results and undesirable responses occur. Heavy sides and bracing prevent this and as a result smoother reproduction of the low frequency portion of the music and voice range is obtained. If the depth of the cabinet is increased, cavity resonance occurs and even if the back is open there is an open organ pipe effect and undesirable responses result. For this reason it is highly desirable that the depth of the cabinet be restricted as much as possible consistent with good appearance. A back on the cabinet may or may not increase the resonance effect depending upon the cabinet design. In general the effect of adding a back will increase the undesired boominess unless special precautions are taken to acoustically ventilate the cavity.

There are many ways of overcoming this cavity effect to almost any desired extent, depending upon the additional cost of the apparatus. Some of these methods employ the back wave in the cabinet to advantage while others merely are concerned with getting rid of certain undesired effects of the back wave. In the absence of publications certain representative patents have been referred to where necessary for the technical material contained therein. One system for reducing cabinet resonance employs several speaker cones or other forms of diaphragms flexibly suspended in openings in the front of the cabinet. Early work on this arrangement was done by Mr. W. D. LaRue at the Victor Talking Machine Co. in Camden. Later developments have been made by Dr. H. F. Olson[4] at the RCA Mfg. Co.

Another system employs an acoustical labyrinth passage in the cabinet at the rear of the speaker for absorbing the back wave without undesired reaction upon the low frequency response of the speaker. In one form of apparatus the exit of the labyrinth has been employed to re-enforce the low frequency waves although, in such a case, best results have been obtained by making the labyrinth expand exponentially, thereby constituting a folded horn loading the rear of the diaphragm. Early work was done on the labyrinth acoustic baffle by Mr. Julian High[5] at Westinghouse Mfg. Co. Later work with a labyrinth baffle of the horn type loading the rear of the diaphragm has been done by Dr. H. F. Olson[6] in connection with high fidelity theatre installations and broadcasting monitoring speakers.

Still another system for overcoming cabinet resonance has employed one or more absorption chambers or wave traps tuned to the frequencies of troublesome resonant peaks. Early work on this arrangement was done by Carlisle of Westinghouse and later developments have been made by Dr. Irving Wolff[7] at the RCA Mfg. Co.

Another system being used this season employs a solid back on the cabinet and a very solid type of cabinet construction. Acoustic ventilation and re-enforcement of the low frequency end of the music and voice range is obtained by a series of pipes located in openings in the bottom of the cabinet. By determining the size and number of these pipes for a given cabinet, it is possible to extend the low frequency response of the over-all sound output one-half to three-fourths of an octave and at the same time to reduce the response from six to nine d.b. at the low frequencies of the voice range, around 100 to 120 cycles. Reference is made to developments by Thuras[8] at the Bell Telephone Laboratories, and to more recent work by C. O. Caulton of the RCA Mfg. Co.

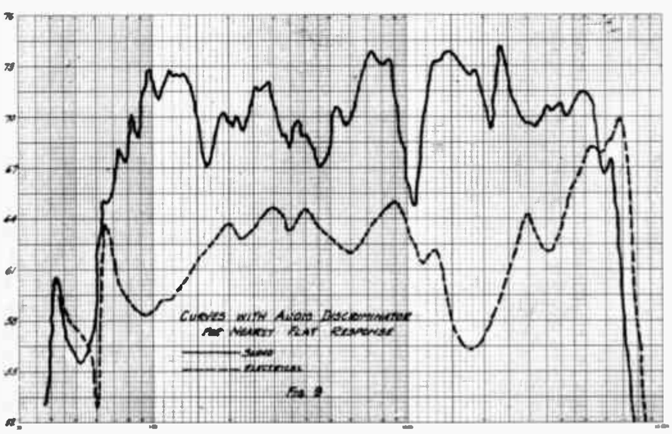
Further refinements for improving acoustic reproduction consists in an inclined speaker baffle in a cabinet. C. R. Garrett[9] and I did some early work on this for the purpose of reducing cabinet resonance. Another refinement consists in a high frequency beam spreader in front of the speaker cone, developed by Dr. Irving Wolff. For high fidelity work the double voice coil speaker, developed by Ringel and Olson, has been used for extending the high end of the range to 8,000 and 10,000 cycles.

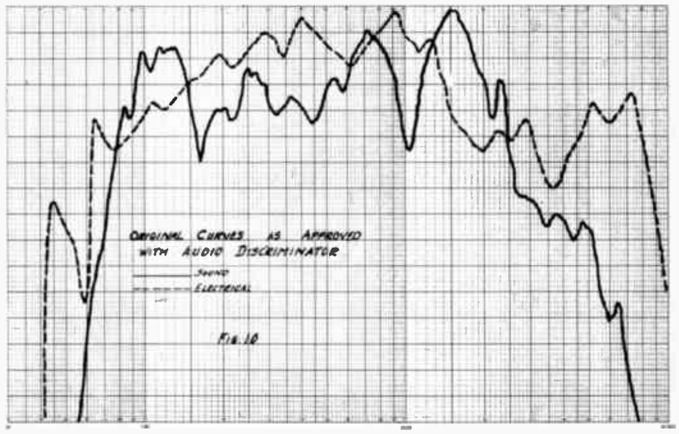
In the absence of technical publications on the above material, reference has been made to patents for convenient reference by those interested in obtaining further details.

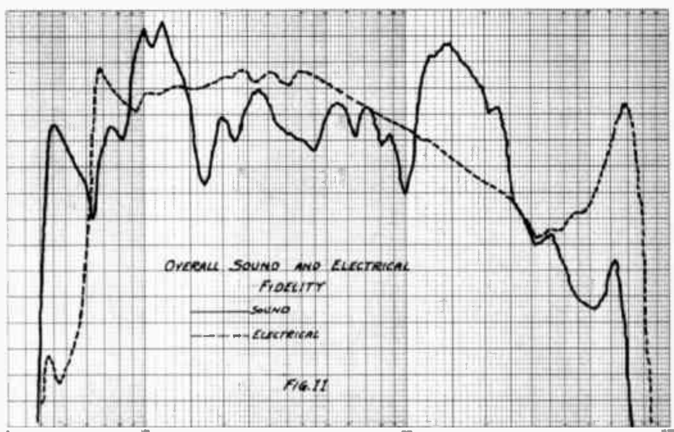
The next portion of the paper will deal with a piece of laboratory equipment which has been very helpful in determining the desired audio frequency characteristics for a given amplifier or input system to obtain pleasing sound output. It is called the audio frequency discriminator. Fundamentally, this device is a compound filter and amplifier provided with a system of controls which permit its frequency characteristics to be altered to almost any desired extent. This control of the frequency characteristic is effected by a division of the audio range of the amplifier, namely 20-10,000 cycles, into eleven filter bands the gain in each band being individually under control. The bands overlap at the sides and are so phased at these points that the combined overall response may be made substantially flat if so desired. The individual bands are slightly less than one octave in width at the overlap point and have a range of amplitude control averaging 12 d.b. up and down from the flat characteristic. A switch is provided which permits the operator to quickly transfer from the normal audio system to that which incorporates the discriminator. In this way it is very easy to compare an audio system which is being worked on with one having the desired characteristic and in this way determine the changes that are necessary to correct the former.

In addition to the eleven filter stages, the discriminator is provided with a continously variable high frequency cutoff filter. This filter is in no way connected with the band filters and allows a much finer control of the upper limit of the tonal range than does the band filter. It has essentially a vertical cutoff over its entire range from 3500 to 10,000 cycles. This cutoff filter may be switched in and out of the circuit as desired.

Figure #9 shows the overall electrical and sound curves of the phonograph being demonstrated with the discriminator adjusted for essentially flat response from 60 to 7,000 cycles. Listening tests on this instrument with this setting indicated that surface noise from a commercial standpoint was highly objectionable. As a result the discriminator controls were readjusted to give the curves shown in Figure #10. It will be noted that while the range remains essentially the same there is a definite tendency toward a trailing off characteristic above 4,000 cycles. The present recording system employed in Victor records has a slightly rising characteristic in this range and the overall result is one which is very pleasing from a musical standpoint; yet the surface noise is not objectionable. Having determined the desired characteristics, it then was necessary to provide the proper equalizing network to obtain similar performance without the discriminator. Figure #11 shows how closely it was possible to duplicate results. A brief demonstration of the use of the discriminator will be given at this point.







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HIGH FIDELITY RADIO RECEPTION

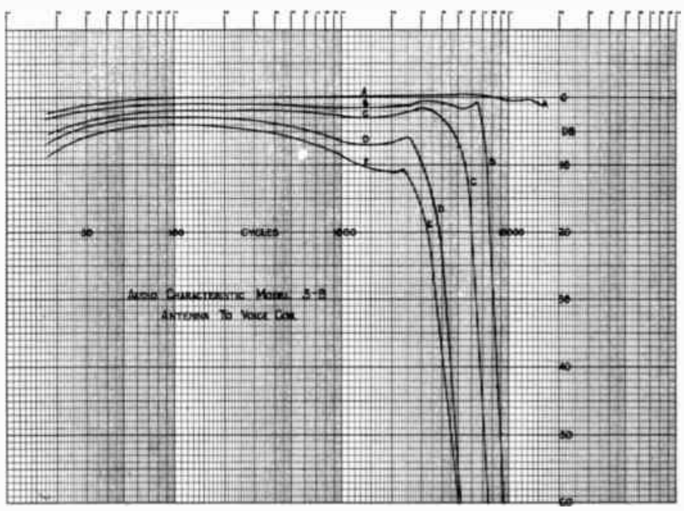
BY

LINCOLN WALSH*

Delivered before the Radio Club of America
December 10, 1936

A high fidelity system might be defined as a system of picking up sounds, transmitting them, and reproducing them so that they sound to the ear precisely like the original sounds. It can also be defined as a system which picks up the sounds, transmits and reproduces them, with all the original sinusoidal components present in their original proportion and phase relation but with the introduction of no new components. If the system does this, the reproduced sound duplicates the original and the system is an ideal high fidelity system. There remains, then the question of how closely a practical system must approach this ideal in order to be satisfactory and this can be answered only by the ear.

Perhaps more than any other sense, the sense of hearing is influenced by suggestion. If the listener has in his mind the conviction that his radio gives him an exact reproduction of the sound in the studio, his ear tells him that there is nothing wrong with the fidelity of his radio. But while the ear of such a listener tells him that all is well with the sound to which he is listening, even when he listens critically to judge the quality of reproduction, an inquiring observer would notice that the extent to which he listens casually to his radio bears a direct relation to its quality of reproduction. Actually, a person will tire quickly of listening to a receiver of poor tone quality, and yet not be conscious of its poor quality. Conversely the same person listening to a receiver of good quality may not be conscious that the quality is better, yet he will listen to it for much longer periods without tiring. It is a matter of common observation that there are many homes having midget receivers of obviously poor quality, whose owners are very proud of the quality and performance of their receivers, and yet those receivers are turned on only for special programs while in homes having receivers of better quality they operate almost continuously. Thus, while the ear is a highly uncritical instrument, it quickly tires of listening to voice and music which is distorted or lacking in important frequency ranges.



The musical instruments of today have evolved thru centuries of listening, and the test which determined their survival was whether or not the tone was suitable as judged by the ear. These centuries of experience have shown that sounds of all frequencies thruout the wide frequency band of audibility are requisite for musical expression and so, when we as radio engineers design radio broadcasting receivers so that they cut off or seriously attenuate frequencies below about 100 cycles and frequencies above 4500 cycles, we are not only undertaking to give a new interpretation to music but we fly in the very face of man's centuries of musical experience.

The human voice is one sound to which we are always listening, and this experience therefore provides an excellent basis for the ready evaluation of the fidelity of reproduction. Even a receiver of very poor fidelity gives a high degree of understandability, because, as is well known, a range of 300 to 3000 cycles is all that is necessary to give understandability. But to give the naturalness that is necessary if the reproduction is to be untiring to the listener, it is necessary to reproduce the entire audible range.

A particular instance of the importance of this is the reproduction of soprano voice. We all know how often the re-

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production of the voice of a soprano as heard over the radio is lacking in the qualities that make it a pleasure to listen to the singer in person. This results from the fact that the fundamental frequency is usually far higher than that of average speech, and the harmonics which give the voice its "color" are not reproduced by the receiver. Yet on a high fidelity system all the overtones are present, and the reproduction is natural and pleasant.

It is a fact, however, that sometimes a program sounds better if the highs are reduced by lowering the cutoff at the high end. This is invariably due to the presence of distortion or to a high background noise level. In the absence of distortion and noise, any normal ear will choose the highest available cutoff.

The ear like most human senses is subject to habit, and if a person is accustomed by habit to listening to a receiver with a low cutoff, he may not immediately react favorably to a high fidelity system. A receiver having a medium cutoff and a peak near that cutoff, may at first sound to such a person as if it has more highs than a truly high fidelity system, but, again, he will be found to tire quickly of listening to the receiver with a peak. But he will listen indefinitely to the high fidelity system without fatigue.

A listener in a comparison test between a high fidelity system, and a system having a lower cutoff, will very often choose the lower cutoff at first but if he takes time, sometimes as much as an hour or two of listening, he will inevitably choose the high fidelity system. Some listeners when first hearing a high fidelity system by itself think it tinny, some think it very bass, some think it has too much bass and too much treble, and lacking in middle register, because they are hearing tones they are not accustomed to hear in radio reproduction. But after listening for an extended period to good program material they like it. And that is the final, and the only reliable test.

These observations are the result of a systematic study of listener reaction started by the writer some nine years ago. They report the conclusions arrived at after a study of the reactions of something over one hundred listeners only a relatively few of which were radio engineers or serious students of music. On the basis of these observations the writer long ago undertook to develop receiver design details to supply the latent and all too little recognized desire for a much closer approach to complete fidelity in radio reproduction. The results of that work have been incorporated in a receiver typical of the especial arrangements which have been found necessary to meet this need and the performance of that receiver will be demonstrated. Before proceeding to the demonstration, however, it will doubtless be of interest to describe something of the design details and their specific purposes and functions.

THE HIGH FIDELITY RADIO RECEIVER

From the standpoint of present broadcast receiver design practise, high fidelity means extending the audio frequency range at both its ends. It has long been known that to have the tone balanced and most pleasant, the audio range of a system must be centered somewhere between 400 and 1000 cycles. If we extend one end of the range, the other must also be extended for best effect.

THE LOUD SPEAKER

At the low end, it has been found desirable to extend the range of the amplifier to below 30 cycles, notwithstanding the limited effectiveness of commonly available speakers in that range. To assist the speaker in this range the largest possible baffle must be used. Olney's work on accoustic labyrinths has pointed the way to improve low frequency speaker response where only limited baffle area is available. An electrical resonance in the circuits, or a mechanical resonance in the speaker - more commonly the latter - is sometimes used to provide a peak at about 100 to 130 cycles, which results in the "boom" of false bass response. In any program, there is always enough energy in the region of this peak to give the low pitched background which many consider to show good bass response. But this causes listener fatigue to develop very quickly and it is thus extremely undesirable to improve bass response by means of any such resonance, either electrical or mechanical. If speaker resonance must be countenanced it should occur at the lowest possible frequency, certainly below 50 cycles; the amplifier must be flat within 2 db down to 30 cycles; the baffle must be of corresponding size, or a suitable labyrinth must be employed; and any bass compensation that is employed must be entirely of resistance-capacity type, to avoid any bass resonance.

THE "TWEETER"

For the high frequency portion of the audio range an especially built "tweeter", which is quite similar to a standard 6 inch cone speaker has been found to supply the best practical solution to the problem presented by the need for efficient translation into acoustic energy of the high audio frequencies. Such a "tweeter" has been included in the demonstration receiver. It has a good sound pressure response curve up to 14,000 cycles, and it is not seriously down at 16,000 cycles. This is secured thru the use of an extremely light voice coil, a short tube connecting the voice coil to the cone, and a light paper cone, of rather low damping.

The "tweeter" is connected thru a small condenser directly across the plates of the push-pull output tubes, so that it cuts in gradually above 2000 cycles. Experience indicates that it is better to have a gradual transition from the low frequency speaker to the tweeter, than any abrupt transition as results from the use of sharp cutoff filters.

While speakers can be built that respond well up to 9000 cycles, the combination of a low frequency speaker and a "tweeter" shows itself to be far superior to any single speaker. One reason for this doubtless resides in the fact that the motion of a large cone diaphragm at high frequencies is made up of two sets of waves radiating out from the voice coil. One wave is longitudinal with respect to the paper of the cone while the other is a lateral wave in the paper, the former travelling at considerably higher velocity that the latter. Their propagation in the paper and their reflection at the edge of the cone determines the high frequency response, and the control of all of these factors is a far more complex and difficult problem than making two speakers of distinctly different proportions each with definite and supplementary characteristics. Additionally where efficient response up to 16,000 cycles is desired no single practical speaker has been found to serve at all satisfactorily.

ELIMINATION OF DISTORTION

Distortion is a very important factor about which volumes could be written. The problem starts at the RF amplifier. The signal voltages at the grid of this tube must be held low to prevent overload and harmonics that will modulate the carrier.

The converter in the demonstration receiver is a special circuit which has a very low noise level. The detector delivers an audio voltage only, and has no relation to the AVC system which is separate. The automatic volume control system holds the detector input voltage at 10 volts for all normal signal inputs, in order to avoid the distortion which occurs in a diode when operated at low voltage levels - within the "parabolic" range - and to avoid distortion in the last IF tube due to overload. which would occur if the diode had to be driven to high voltages.

Additionally, of course, the AVC system holds signal voltages thruout the receiver at such values that will avoid overloading any of the tubes such as the converter and the IF tubes.

It is not, perhaps, commonly appreciated to what degree the use of silicon steel in the interstage audio transformers introduces harmonics, particularly at low levels. But because of this fact the audio signal is carried thru resistance capacity coupled circuits, having no iron core devices anywhere, up to the input of the push-pull output tubes. At this point a push-pull transformer of special design, including a core of high-permeability alloy, which does not generate harmonics is employed, and which contributes greatly to the clean tone quality of the receiver.

All the audio amplification is provided by the use of low-mu triodes, which are the only amplifiers sufficiently free of distortion for a high fidelity system except, perhaps, as the newly developed degenerative circuit arrangements may make the multi element tube less unsuited to this field.

THE I. F. AMPLIFIER

The problems presented by the need for so designing the selective high frequency amplifier stages as to provide for high fidelity reception are basically impossible of solution since, under the American scheme of broadcast frequency allocation and assignment in which adjacent assignments differ by only ten K.C. and practically all frequencies so assigned are in simultaneous use, the requirement that interference-free reception be possible on any assigned frequency at any time and place, unavoidably limits the audio band width of reception to something less than five thousand cycles. Under these limitations there is little that the radio designer can do other than to provide a relatively narrow band width and pray that it will be found not too unacceptable. And, indeed this is precisely the direction in which the receiver designs of recent years have gone.

It is patently absurd, however, to so limit the fidelity of receivers so that only such painfully low fidelity is available to the listener who is located relatively close to his local transmitter or who may be in the high field strength area of a high powered transmitter and thus largely free of adjacent channel interference and interference from noise sources. And since the system under which our radio receivers are distributed to the purchasing public requires so complete a universality of usefulness there is obviously no solution but to give the receiver such variable selectivity as to provide, on the one hand, so narrow a band width as will allow of distance reception in areas of noise and interference and, on the other hand, to provide so great a band width as will allow of the reproduction of the entire audio range being broadcast by the best transmitting system.

In the demonstration receiver this is accomplished in the intermediate frequency amplifier since, as is usual in the superheterodyne type of receiver here largely resides the selectivity of the system. It has been found best to use two IF stages, including three double tuned IF transformers, the coupling between the primary and secondary circuits being varied by moving the secondary coil relative to the primary. At the position of minimum coupling, the coupling is considerably below critical, and the selectivity is at its highest. At the position of maximum coupling, position A in the figure, the resonance curve of the first two transformers becomes double peaked, with the peaks separated by about 35 kilocycles. The third transformer which feeds the detector is broadened, but not enough to show double peaks. The single peaked resonance curve of this latter transformer fills in the valley of the combined curves of the other transformers so that the resultant overall curve of the IF system is flat over a band of about 32 KC, thus permitting the unattenuated passage of sidebands corresponding to all audio frequencies up to 16,000 cycles. A second step of coupling, position B, is provided in the receiver in which similar conditions exist, with, however the band width reduced to about 16 KC corresponding to an audio band of 8,000 cycles. Positions C and D have band widths of 12 and 8 KC respectively, and the fifth position, E, is the position of minimum coupling, and passes about 5 KC. The audio bands corresponding to these are, respectively, 6,000, 4,000, and 2,500 cycles.

It might be well to point out in passing that the use of variable inductive coupling as here employed has certain advantages over other possible types of coupling that might be employed, wholly aside from the obvious advantages of economy of production, ease of production adjustment etc. It will, of course, be remembered that, in general, the peaks in the transmission characteristic of a pair of tuned and over-coupled circuits can be equal only when there is no loss in the coupling element and since mutual inductance is the one coupling element that is loss free, it is thus especially suited to this purpose. Another factor of interest here is that by varying only the mutual by the motion of one of the coils, the band width is varied while maintaining the midfrequency fixed.

R. F. AMPLIFIER

The radio frequency system has as its primary function the elimination of the image frequency, which in the demonstration receiver is 940 KC higher than the signal frequency. It serves also to eliminate such other signal frequencies, as might be brought in thru beating with harmonics of the oscillator. The RF system may, therefore be made broad enough to suit the widest band passed by the I. F. amplifier and need not be of variable band width. The antenna circuit of the demonstration receiver is double tuned, and double peaked, with peaks separated about 35 KC. The RF amplifier is single tuned and is broadened by the use of very fine wire in the winding, so that it just about fills in the valley of the double peaked antenna system. At the higher end of the broadcast band, the valley of the antenna circuit is less deep, and the RF stage is less sharp, so that the RF system as a whole is flat within 1 db over a band of about 35 KC.

This gives a system which will pass all frequencies up to 16,000 without any attenuation. When this system is tried on the air, the tone quality is all that might be expected. But after sunset, when the distant stations on adjacent channels begin to come in, 10 KC whistles are heard. If a filter is put in circuit to cut out 10 KC, the whistle disappears, but then the so called "monkey chatter" is heard. This chatter is the high pitched unintelligible sound which is due to the side bands of the adjacent channel beating with the carrier of the desired signal, in contradistinction to the more commonly experienced interference, known as "cross talk" which is the result of the adjacent channel side bands beating with their own carrier. It is of interest to note that in the receivers which have been built on the basis here discussed, crosstalk from the adjacent channel has been far less serious in creating interference than the adjacent channel chatter. As a matter of experience, it has been found that chatter is the ultimate limitation on fidelity.

It has been found however that, when receiving moderately strong signals from local stations, practically all the chatter can be eliminated by the use of an extremely sharp low pass filter, having its cutoff at about 7500 cycles.

The reason for this is not hard to find if it is remembered that workers in the acoustics and telephone fields have shown that most of the energy in speech and music lies in the frequency range below 2000 cycles, and that the energy drops progressively as the frequency under inspection increases. Thus it is found that there is very little energy in the region of 8000 cycles and higher although that little energy is of great importance in giving naturalness to the sound of which it is a part. Similarly, then most of the energy in the sidebands of a signal is in those frequencies differing by less than two thousand cycles from their carrier and hence by more than 8000 cycles from the carrier of the adjacent channel transmission. Thus, the resulting "monkey chatter" carries most of its energy in the frequencies above 8000 cycles and is thus subject to effective elimination thru the suppression of all frequencies of that order or higher.

THE AUDIO FILTER

To eliminate both the chatter and the whistle in the demonstration receiver, there is used a 4 section Campbell type filter, having 11 elements; 4 series inductances, 5 shunt capacities, and 2 mutual inductances. This filter is flat up to 7200 cycles, and is down approximately 30 db at 8000, 50 db at 9000, and 70 db at 10,000 cycles. These values of attenuation have been found necessary to receive local signals without chatter in the 7500 cycle setting of the fidelity control.

By means of the fidelity control in the demonstration receiver, the band width of the I.F. is varied and thru a simple mechanical linkage with the audio filter, the cutoff of the audio system is simultaneously changed to provide a cut off just below the cutoff of the IF system. This gives the highest possible audio band width for a given selectivity.

In practical use it has been found possible to receive signals as low as 10 millivolts without chatter on the 7500 cycle setting, but with all the superior fidelity implied by that high cutoff as compared with conventional receivers. Weaker signals may require that the cutoff be set at 5500 cycles, which provides better than usual fidelity. When very high selectivity is desired as in cases of high noise levels and in the reception of distant stations, the band is narrowed to 4000 or even to 2500 cycles. This latter band width appears pointlessly narrow but it must be remembered that many listeners want high selectivity and the ability to get distant signals for a short period after they buy their receiver and these circuit arrangements make that possible. Happily, however, after they experience the pleasure of good programs they want little more than the nearby stations, and these they may receive on the high fidelity ranges free of chatter or noise and with highest fidelity consistent with the conditions of transmission and reception.

Experience indicates that in daylight the local stations can usually be heard without whistles, chatter, or any noise on the 16,000 cycle setting, and indeed many stations are transmitting programs of a quality which shows a definite improvement when the cutoff is raised from the 7500 to the 16,000 cycle position. Especially well does this type of receiver operate when receiving the high fidelity stations which have 20 KC channel separation, and good lines and amplifiers, and which are therefore best listened to on the widest band.

FIDELITY OF BROADCAST TRANSMITTERS

It is often argued that the broadcasting stations themselves do not supply programs with such frequency band widths as to justify the use of high fidelity receivers. With this in mind the writer sometime ago made casual investigation as to the upper cut-off of the higher powered transmitters in the New York area and he was happily surprised to note the care which has been taken in the design of much of the broadcasting equipment to maintain the upper frequency limit of the equipment at such a high value as to provide for truly high fidelity transmission. It was found on inquiry, for example, that the studio and transmitter equipment of WJZ and WEAF is good up to 16,000 cycles per second as is that of WABC and WOR. The telephone lines connecting the studios and the transmitters of these several stations are reportedly not as good as the equipment. Thus, the line to the WEAF transmitter is reported to cut off at 7000, the WABC line, at 8500, the WJZ line at 10,000 and the WOR line at 11,000. These latter data on the cutoff frequency of the telephone lines are not to be viewed as inherently limiting the broadcasting system since, they doubtless result largely from economic and not technical factors, and can and doubtless will be raised as the demand for such improvement developes. It must, however, be admitted that there is little incentive for any broadcaster to assume the added burdens of cost and maintainance required by any further expansion of the band width of his transmission until and unless the receivers in the audience to his transmissions are capable of taking full advantage of that transmission.

Thus, once again we have arrived at the stage in the development of broadcasting where the transmitter leads the receiver and now awaits being overtaken. Thruout the history of broadcasting, first the receiver and then the transmitter has been the more thoroughly developed element of the system and in each step in the sixteen years of progress in radio broadcasting each improvement in the lagging element has placed it so markedly in advance of the other as to provide the incentive for major improvement in the other which in turn prompted further improvement in the first, and so on and so on around the widening spiral of progress.

It seems quite reasonable, therefore, to believe that not only will the next move in this continued progress be made by the receiver designer-in the direction here described in detail - and with the agreement of his commercially minded associates, of course - but that that next step will leave the broadcaster lagging once again only, however, to have him soon moving again in the direction of the ultimate perfection of broadcasting.

EDITOR'S NOTE

In the course of the delivery of Mr. Walsh's paper demonstrations were made in connection with a special program for the Club from radio station WQXR under the direction of Mr. John V. L. Hogan. This program included a variety of broadcasting material designed to show the superior effectiveness of the high fidelity reception through the transmission of a wide range of tones and special musical transmissions both with and without the use of cut off filters at the transmitting station. It can be reported that any cut off at the transmitter that tended to reduce its transmission band width below the maximum possible was quite easily evident in the reception as heard by those in attendance and it was of especial interest for most of the members of the club present to make comparisons between the reproduction of Mr. Hogan's voice and their memory of it as it so often when Mr. Hogan attends in person. The demonstration left no doubt as to the need for the transmission of the entire band width of the receiver and the transmitter where completely satisfactory fidelity is the aim. Additionally it should be pointed out that while this demonstration was made at Havemeter Hall, Columbia University at which previous experience has shown that the noise level is always objectionably high, the installation of a noise suppression antenna system, through the kindness of Messrs Amy and Aceves and King, left little noise interference to detract from the demonstration or from the enjoyment of the special musical program that was transmitted.

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PROCEEDINGS of the RADIO CLUB OF AMERICA

Volume 14 — March, 1937 — No. 1

FACTORS RELATING TO FAITHFUL REPRODUCTION

BY

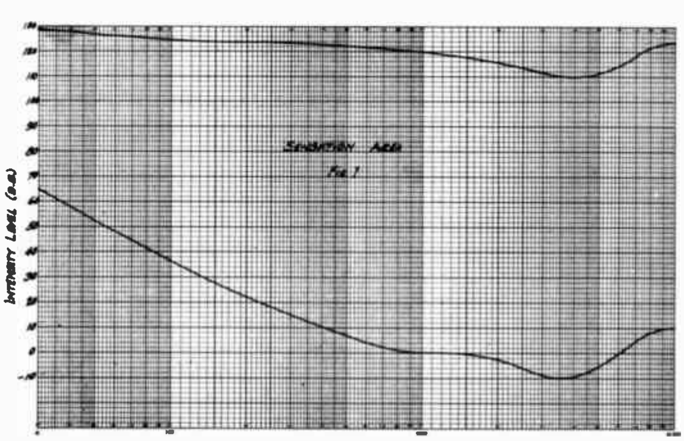
C. M. SINNETT*

Delivered before the Radio Club of America
October 15, 1936

The subject of phonograph or radio reproduction is necessarily a very broad one. For this reason the following paper concerns itself only with aural compensation and volume expansion as aids in obtaining the desired result. Demonstrations will be given of various factors involved and a very useful piece of laboratory equipment in connection with this work will be described and demonstrated.

Before discussing aural compensation as applied to an electric phonograph or radio receiver, it would probably be well to digress a moment and consider the general subject of sound and hearing.

A large amount of research work has been carried on by different experimenters in connection with sound at different levels and its effect upon the human ear. General reference to books and papers on the subject gives one the impression that although some discrepancies exist between results obtained by different experimenters, there is, in the majority of cases, quite close agreement with the work of Dr. Harvey Fletcher of the Bell Laboratories. For this reason we have, in our work on aural compensation, used data presented by Dr. Fletcher on his paper "Loudness, Its Definition, Measurement and Calculation"[1].



Listening carefully to the radio set of a few years ago, we have all undoubtedly noticed that a reduction in volume from a rather loud level to one which could be used in the average apartment without causing annoyance to ones neighbors was accompanied by a change in musical balance. At the louder level the low frequency instruments such as the bass viol, bass drum and tuba were in proper balance with the middle and high frequency instruments. At the lower level, however, the low frequency instruments were no longer in balance and the music had a harsh or tinny characteristic. This effect could also be obtained at an outdoor orchestral performance or band concert if one were to change his position from one close to the instruments to a position a hundred feet removed from them. The high frequency instruments will decrease in volume much less rapidly than the tuba and other low frequency instruments. The reasons for this phenomenon are very clearly given in Dr. Fletcher's paper.

Figure #1 illustrates the frequency response and intensity range of the average human ear from the threshold of hearing to the threshold of feeling. These two thresholds were determined at different frequencies by means of pure tones applied to head phones worn by the observers. Many different readings were taken at various fixed frequencies and for many different observers. From these, it was possible to plot the lower curve which shows the threshold of hearing for the average individual. It will be noted that the curve is plotted with 1,000 cycles as the 0 d.b. level. For reference purposes the input level to the ear canal at this frequency was found to be 10^{-16} watts per square centimeter. It is easily seen by reference to the curve that for a frequency of 100 cycles, the same intensity level in the ear would require an input 35 d.b. higher than for 1,000 cycles. Conversely, the input in the 3000 - 4000 cycle range would need to be 10 d.b. less than that at 1000 cycles to produce equivalent effect. The upper curve was determined in much the same manner with the exception that the limiting factor became the point at which the observer felt actual pain. This curve, which is called the threshold of feeling, represents the highest level the average individual can withstand for

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any length of time without actually suffering pain as a result. It will be noticed in this case the input at 100 cycles need be only 5 d.b. higher than that at 1000 cycles to produce the same result. In other words at the threshold of feeling the curve of the average ear is essentially flat.

Between the two thresholds, it then became possible to plot a family of curves representing the average characteristic at different levels. For convenience these curves were determined in 10 d.b. steps at 1000 cycles beginning with the threshold of hearing and carrying up to the threshold of feeling. Figure #2 shows this family of

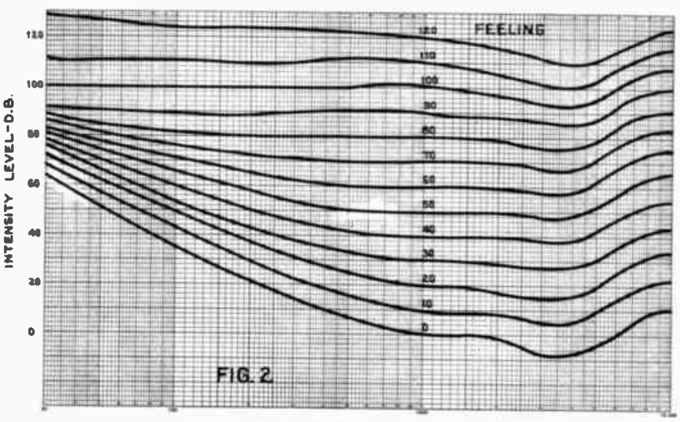
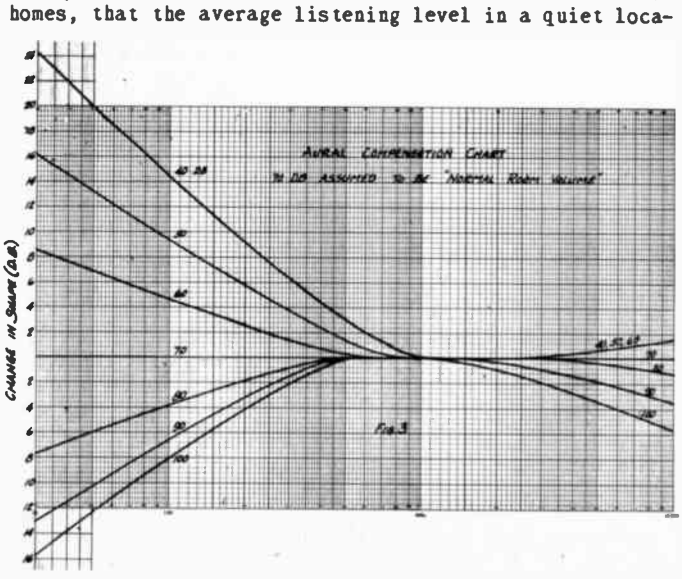


FIG. 2

curves. An observation of these curves indicates immediately that the range between 500 and 5000 cycles is practically flat for almost any level but that for frequencies below 500 cycles, the input to the ear must be increased over the 1000 cycle level to maintain proper balance. These curves then show us that some form of low frequency compensation is necessary if we are to maintain proper balance over very wide changes in volume. It is, of course understood that the volume changes referred to are the changes in average level from one value to another rather than the normal variations in dynamic level which are automatically taken care of by the conductor of the orchestra.

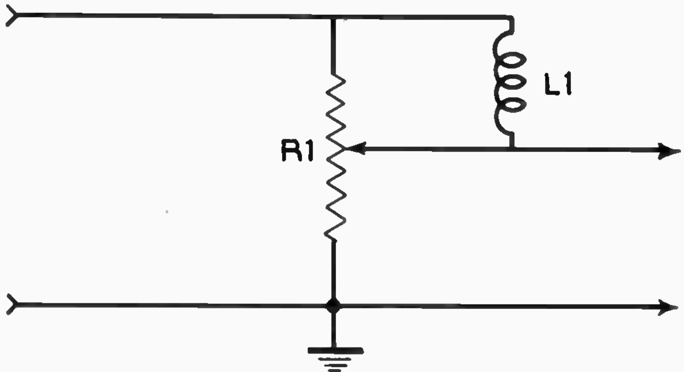
It has been indicated by Dr. Fletcher's work and borne out by tests with calibrated volume controls in observers' homes, that the average listening level in a quiet loca-



tion is approximately 70 d.b. above reference level or threshold of hearing. With this level as a basis and assuming that a desirable sound characteristic has been obtained at this level, it then becomes possible to plot a new series of curves from those shown in Figure #2 and which are shown in Figure #3. For convenience, the 70 d.b. level is shown as flat since we have already assumed that the subject level was satisfactory and thus any irregularities in the curve must be carried through to the other levels. From this family of curves, we can readily determine the amount of compensation necessary as the sound level is changed above and below the 70 d.b. average level. For instance, a decrease in average level of 10 d.b. at 1000 cycles must be compensated for by a decrease of only 5 d.b. in response at 100 cycles. Similarly, a decrease in average level of 20 d.b. at 1000 cycles must be compensated for by a decrease of only 10 d.b. at 100 cycles. The higher frequencies are not affected as much but it is desirable that the response be decreased only about 8 d.b. at 10,000 cycles for each 10 d.b. decrease in average level at 1000 cycles. Conversely, an increase of 10 d.b. in average level above the 70 d.b. level at 1000 cycles must be accompanied by an increase of only 6 d.b. in the 100 cycle level. An increase of only about 2 d.b. at 10,000 cycles should accompany this 10 d.b. increase in level. These curves may thus be used directly to determine the amount of compensation necessary at all levels above and below the average reference level of 70 d.b.

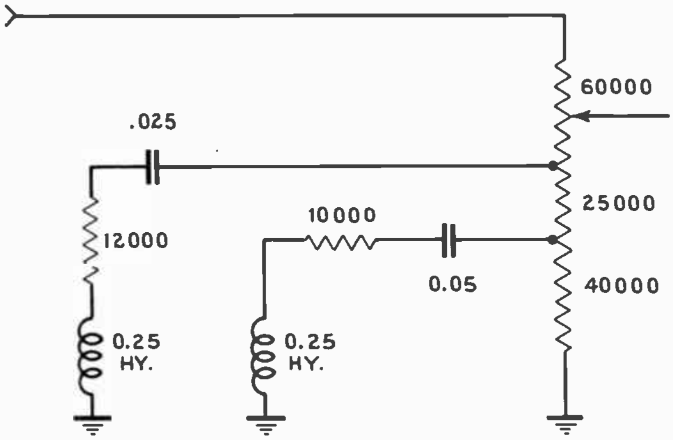
Having determined that some form of compensation is necessary, and with curves available showing the amount required for the different levels, we are faced with the requirement of finding ways and means of accomplishing the desired result.

One method which has been used in the past with good results on low impedance phonograph input systems is shown in Figure #4. In this, the volume control, which is of



SIMPLE COMPENSATION CIRCUIT
FIG. 4

the potentiometer type, has an inductor L_1 connected between the slider and the high end of the volume control. The size of the inductor will, of course, be determined by the resistance of the volume control with which it is used. In the particular case illustrated, the value of L_1 was approximately 30 millihenries and the volume control resistance was 60 ohms. With these values, the shunting effect of L_1 is almost negligible at 1000 cycles. At 100 cycles, and at mid point on the volume control, the shunting effect is approximately 2 to 1 or in other words, we have a rising characteristic 6 d.b. higher at 100



TYPICAL COMPENSATED VOLUME CONTROL
FIG. 5

cycles than at 1000 cycles at mid volume setting. This method of compensation is not adequate for better grade phonographs or radio receivers, but has been used with considerable success on lower priced instruments.

The next method, shown in Figure #5, has been found quite adequate for present day requirements. It will be noted in observing the diagram that there are two fixed taps on the resistance element in addition to the slider. These taps are used in conjunction with the proper shunting networks to give the proper amount of low and high frequency compensation for the different levels. The value of total resistance used is of necessity determined by the particular application. The positions of the taps are determined by the gain of the amplifier following the control and bear such a relation to the total resistance that normal listening level occurs at approximately the first tap down from maximum volume setting taking into account the compensation networks. The diagram indicates the values used for one particular compensated control which is being used in the instrument for demonstration. The shunt networks may have values changed as required to give the particular curve desired at normal level as well as the degree of compensation needed to fulfill the curves of Figure #3. A smaller capacitor will have the effect of increasing the low frequency response with reference to 1000 cycles whereas a larger capacitor will cause the opposite effect, other values remaining constant. The control and its attendant circuits are designed such that the change in level between taps is 20 d.b. When the slider has reached the second tap, no further compensation occurs. A further improvement can be shown by the addition of a third tap to the control, but this can be applied only to the higher priced instruments. The amount of high frequency compensation is determined by the size of the inductors used. Increasing the size of the inductor increases the high frequency response and conversely decreasing its size decreases the amount of tip up at the high frequency and as the volume is decreased.

A third type of tone compensation, which has been used in the past, is the automatic bass compensator. In this circuit, an amplifier system of the variable gain type is employed in parallel with the regular audio channel. By virtue of its variable gain characteristic and due to the fact that its response is limited to frequencies below 1000 cycles, it can be made to increase the overall low frequency response as the volume level is decreased. Due to the added cost and nicety of balance required for operation, this particular system has not had wide application in receivers enjoying quantity production. A similar arrangement has been applied to the radio receivers for restricting the amount of high frequency reproduction with a decrease in antenna signal voltage.

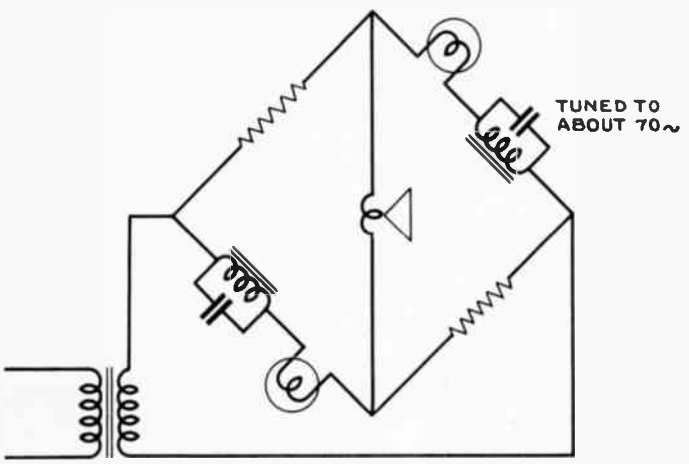
This covers briefly the subject of aural compensation and some possible methods of accomplishing the desired results. A short demonstration of this feature will follow as a means of showing aurally exactly what this means to you and me.

The next portion of this discussion will center around volume expansion as applied to both phonograph and radio reproduction. When volume expansion is mentioned to the average person, he immediately wants to know what advantage a set employing this system has over one which is not so equipped. To better understand the reasons for the existance of volume expansion, and to point out its advantages, let us consider changes in volume level of the orchestra, limitations in the reproduction of these changes as introduced by the phonograph or radio system and restoration of the original by means of an automatic volume expansion system.

By extensive tests, it has been well established that a large symphony orchestra has an available volume range of approximately 70 d.b. from a soft pianissimo to a heavy cresendo passage. If we were to try and cover this volume range on a phonograph record, we would immediately be faced with two definite limitations. The cresendo passage would, of necessity, require maximum movement of the cutter on the wax consistent with groove spacing and other mechanical limitations. With this as the top limit, the movement of the cutter on pianissimo passages would be so microscopic that when played back, the only apparent result would be surface noise or record scratch. Obviously, these passages must be brought above the surface noise level, in order to be heard and the cresendo passage must be limited as regard cutter travel, for the reasons outlined above. In doing this, compression of the 70 d.b. volume range occurs and we find a total volume range on latest records of approximately 50 d.b. When one considers that 20 d.b. difference represents a change in voltage or needle velocity of 10 to 1 and that 50 d.b. is 320:1 and 70 d.b. is 3200:1, I believe, it is readily apparent why the expression the director tried so hard to obtain during recording has been somewhat tempered by the time it is reproduced on an ordinary instrument. To illustrate by figures just what these limitations are mechanically on a phonograph record, let us consider that a recording is being made at about 100 grooves per inch. The maximum swing of the cutter point is thus limited to less than .010 inch if the grooves are not to touch each other. Assuming we could take this distance on loud passages, then the pianissimo passage, 70 d.b., lower than this would be approximately .000003 inch. This is much less than the microscopic structure of the record material itself and even an increase in the pianissimo passage of 20 d.b. still allows a swing of only .00003 inch. From these figures, I believe it is readily apparent that we are doing well to obtain the present volume range on the record alone.

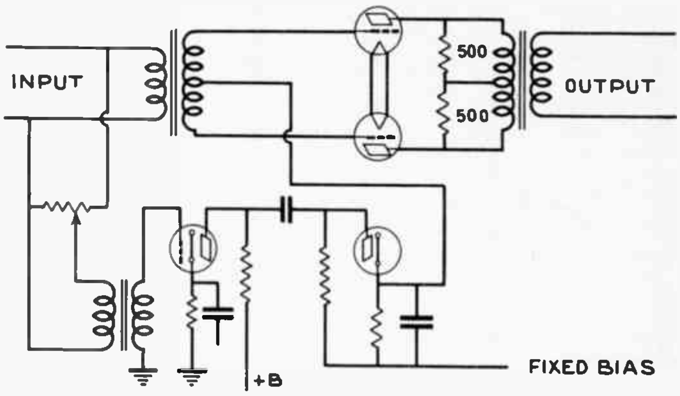
Limitations of a similar type are imposed by the broadcast station except for the fact that in this case line noises, hum and other extraneous disturbances require increasing the level on pianissimo passages and the danger of over modulation requires decreasing the level on cresendo passages. Present day high grade stations are able to cover a volume range of approximately 50 d.b. This is seldom used, however, since there is a definite desire on the part of the station owners and operators as well as program sponsors to obtain maximum listener coverage at all times by maintaining as high an average level of modulation as possible. Wide volume changes on present broadcast equipment would be satisfactory for people located close to the transmitter, but appreciation of it would be lacking by those located at some distance, particularly when the pianissimo passages were pushed aside by bursts of static.

The above discussion, I believe, indicates the desirability of obtaining some form of volume expansion as a means of restoring this loss of 20 d.b. in volume range particularly if full enjoyment of symphonic programs is to result. There are several ways of obtaining this result and the use of any one type is dependent upon the amount of expansion desired and cost permitted. With any of the systems which can be used at present, it is obvious that the result is a compromise since monitoring for all recording and broadcasting is accomplished manually and expansion in the phonograph or radio must take place automatically. Eventually, it may be possible to automatically compress the volume range during recording or broadcasting and so design the phonograph or receiver that its expansion curve is the counterpart of the compression curve. When this has been done, then the listener will be able to enjoy to a much greater extent, the various symphonic programs available. Until that time, however, we can obtain a great amount of pleasure even from the present volume expansion systems as applied to standard phonographs and broadcast receivers.



SYSTEM No. 1
BRIDGE TYPE EXPANDER CIRCUIT
FIG. 6

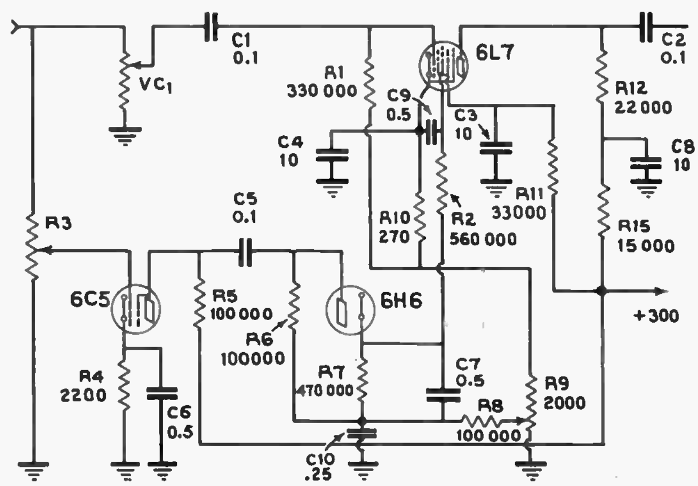
One relatively simple method, shown schematically in Figure #6, has been used the past year on a commercial broadcast receiver.[2] It employs low voltage electric lights of special design in a balanced bridge circuit across the output system. At low signal levels, the bridge is nearly in balance and very little signal gets through to the loudspeaker. As the audio signal increases, the resistance of the filaments in the lights changes and the bridge is thrown out of balance. This allows more than the direct increase in signal strength to be applied to the voice coil of the loudspeaker resulting in a degree of volume expansion dependent upon the strength of the applied signal. Earlier forms of this device imposed power output limitations upon the audio system but improvement in thermal resistance characteristics of the filaments used has resulted in a system capable of 15 d.b. expansion without serious distortion due to change in load impedance across the output tubes.



SYSTEM No. 2
PUSH PULL TRIODE, VARIABLE IMPEDANCE
EXPANSION CIRCUIT
FIG. 7

A second system, shown in Figure #7 consists of push-pull triodes having transformer coupling in both input and output circuits.[3] The plate circuit of the tubes, which normally calls for a load impedance of 20,000 ohms for proper matching, is shunted to approximately 1000 ohms by means of two 500 ohm resistors connected across the primaries of the coupling transformer. A variable bias system, dependent upon the strength of the incoming signal for its action, is connected to the common grid return of the triodes. In this type circuit, the transfer of energy between these tubes and the next audio stage is dependent upon the effective plate impedance of the tubes and the resistance of the load. Since the tube impedance is dependent upon the grid bias, there is a proportionally greater transfer of energy to the 500 ohm circuit at low grid bias than at high grid bias. In this way it is possible to use this circuit for volume expansion with very low distortion. Its main disadvantage, as far as commercial phonographs or broadcast receivers are concerned is its greater cost as compared with systems which do not require as many tubes and do not use transformer coupling.

The third system, shown schematically in Figure #8, has been used by RCA Victor for the past year for phonograph reproduction. Since this is the system to be demonstrated shortly, I believe it would be well to discuss the functions of the various parts. It will be noted that the incoming audio signal branches at VC_1. One branch



RCA VOLUME EXPANSION CIRCUIT
FIG. 8

goes to the #1 grid of the 6L7 variable gain amplifier tube and the other branch terminates in the degree of expansion control R_3. The 6L7 tube has previously been adjusted by means of variable resistor R_9 to operate on the proper portion of its characteristic for the amount of volume expansion and audio gain required. In the particular circuit, the tube gain is about 2.5 and the plate current is about 1 milliampere in the zero signal condition. As signal is applied and with R_3 set as maximum, amplification of the signal occurs in the 6C5 and rectification takes place in the 6H6. This rectified voltage, the value of which is determined by the strength of the incoming signal appears across R_7 and is thus impressed upon the #3 grid of the 6L7 through the time delay circuit composed of R_2 and C_9. The polarity of this voltage is opposite to that which is already present on the #3 grid and serves to reduce the effective voltage on this grid. This increases the voltage gain in the 6L7 and a variable gain amplifier results, the gain of which is entirely dependent upon the strength of the incoming signal. This system provides a relatively cheap volume expander capable of increasing the volume range 15-18 d.b. if so desired. The amount of expansion is easily varied by means of R_3 and in this way the desired result can be obtained on almost any type of phonograph or radio program.

A discussion of this subject must of necessity include some mention of the cabinet and its relation to the overall musical balance. The broadcast type cone loudspeaker radiates from both sides of the diaphragm. When this type speaker is mounted on a flat baffle and away from corners of the room or other cavities, very little low frequency resonance is present. When, however, the flat baffle is folded back to form a cabinet for the loudspeaker an immediate change takes place in the low frequency balance unless special precautions are taken. The cavity behind the loudspeaker cone serves directly in reinforcing the low frequency response and "boomy" reproduction results. To many people this type of reproduction is entirely pleasing and unless some "boom" is present they feel the set is not properly designed. To the music lover, however, this "boom" is highly objectionable since it is a type of musical balance, or unbalance, which never occurs in an orchestra.

There are many factors which govern the amount of cabinet resonance reproduced, among which the more important are: type of output system used; whether high or low impedance; frequency of resonance of cone suspension system; ruggedness and weight of wood used for the cabinet; depth of cabinet from speaker baffle to back opening and whether the back is open or closed. A brief discussion of each of these factors will enable us to understand more fully their direct effect upon reproduction.

If a high impedance output system is used, for instance one employing pentodes, changes in impedance of the plate load cause a proportional increase in voltage across the load due to the constant current characteristics of these tubes. A cone londspeaker at its suspension resonance frequency presents a much higher impedance than at 400 cycles. For this reason it is desirable, if boominess is to be decreased, that the cone resonance be located below 70 cycles. Furthermore if a reduction in resonance voltage or output is desired at this frequency then a low impedance output system should be used. With either system the cone suspension resonance should never be located above 80 cycles in a console model since average cabinet resonance in this type cabinet occurs in the band between 100 and 150 cycles depending upon the cabinet depth.

Another factor directly connected with the amount of low frequency resonance effect is the weight of wood used for the cabinet. If thin woods are employed with very little bracing then at those frequencies where resonance occurs, or close to them, vibration of the cabinet sides results and undesirable responses occur. Heavy sides and bracing prevent this and as a result smoother reproduction of the low frequency portion of the music and voice range is obtained. If the depth of the cabinet is increased, cavity resonance occurs and even if the back is open there is an open organ pipe effect and undesirable responses result. For this reason it is highly desirable that the depth of the cabinet be restricted as much as possible consistent with good appearance. A back on the cabinet may or may not increase the resonance effect depending upon the cabinet design. In general the effect of adding a back will increase the undesired boominess unless special precautions are taken to acoustically ventilate the cavity.

There are many ways of overcoming this cavity effect to almost any desired extent, depending upon the additional cost of the apparatus. Some of these methods employ the back wave in the cabinet to advantage while others merely are concerned with getting rid of certain undesired effects of the back wave. In the absence of publications certain representative patents have been referred to where necessary for the technical material contained therein. One system for reducing cabinet resonance employs several speaker cones or other forms of diaphragms flexibly suspended in openings in the front of the cabinet. Early work on this arrangement was done by Mr. W. D. LaRue at the Victor Talking Machine Co. in Camden. Later developments have been made by Dr. H. F. Olson[4] at the RCA Mfg. Co.

Another system employs an acoustical labyrinth passage in the cabinet at the rear of the speaker for absorbing the back wave without undesired reaction upon the low frequency response of the speaker. In one form of apparatus the exit of the labyrinth has been employed to reenforce the low frequency waves although, in such a case, best results have been obtained by making the labyrinth expand exponentially, thereby constituting a folded horn loading the rear of the diaphragm. Early work was done on the labyrinth acoustic baffle by Mr. Julian High[5] at Westinghouse Mfg. Co. Later work with a labyrinth baffle of the horn type loading the rear of the diaphragm has been done by Dr. H. F. Olson[6] in connection with high fidelity theatre installations and broadcasting monitoring speakers.

Still another system for overcoming cabinet resonance has employed one or more absorption chambers or wave traps tuned to the frequencies of troublesome resonant peaks. Early work on this arrangement was done by Carlisle of Westinghouse and later developments have been made by Dr. Irving Wolff[7] at the RCA Mfg. Co.

Another system being used this season employs a solid back on the cabinet and a very solid type of cabinet construction. Acoustic ventilation and re-enforcement of the low frequency end of the music and voice range is obtained by a series of pipes located in openings in the bottom of the cabinet. By determining the size and number of these pipes for a given cabinet, it is possible to extend the low frequency response of the over-all sound output one-half to three-fourths of an octave and at the same time to reduce the response from six to nine d.b. at the low frequencies of the voice range, around 100 to 120 cycles. Reference is made to developments by Thuras[8] at the Bell Telephone Laboratories, and to more recent work by C. O. Caulton of the RCA Mfg. Co.

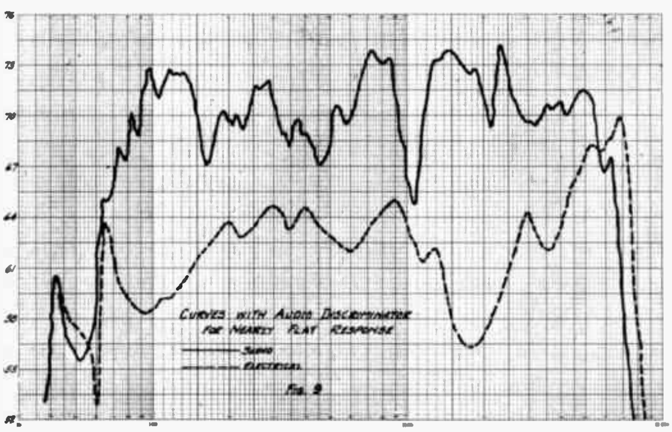
Further refinements for improving acoustic reproduction consists in an inclined speaker baffle in a cabinet. C. R. Garrett[9] and I did some early work on this for the purpose of reducing cabinet resonance. Another refinement consists in a high frequency beam spreader in front of the speaker cone, developed by Dr. Irving Wolff. For high fidelity work the double voice coil speaker, developed by Ringel and Olson, has been used for extending the high end of the range to 8,000 and 10,000 cycles.

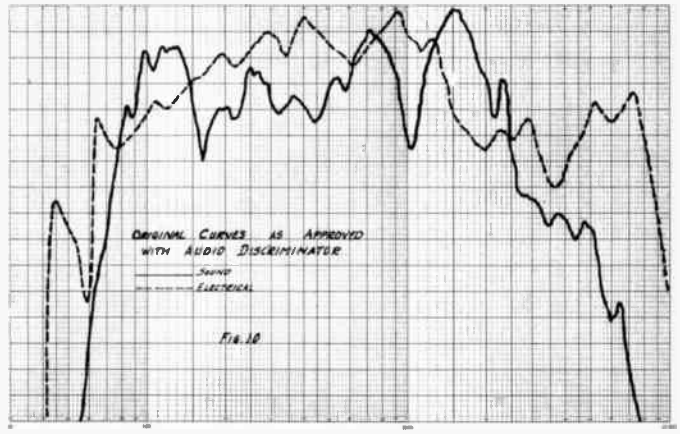
In the absence of technical publications on the above material, reference has been made to patents for convenient reference by those interested in obtaining further details.

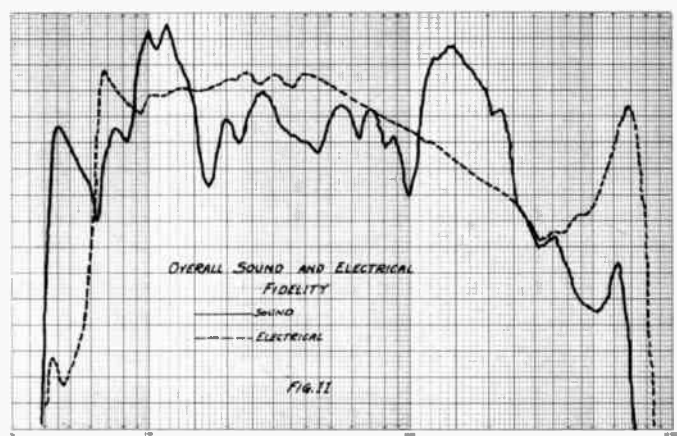
The next portion of the paper will deal with a piece of laboratory equipment which has been very helpful in determining the desired audio frequency characteristics for a given amplifier or input system to obtain pleasing sound output. It is called the audio frequency discriminator. Fundamentally, this device is a compound filter and amplifier provided with a system of controls which permit its frequency characteristics to be altered to almost any desired extent. This control of the frequency characteristic is effected by a division of the audio range of the amplifier, namely 20-10,000 cycles, into eleven filter bands the gain in each band being individually under control. The bands overlap at the sides and are so phased at these points that the combined overall response may be made substantially flat if so desired. The individual bands are slightly less than one octave in width at the overlap point and have a range of amplitude control averaging 12 d.b. up and down from the flat characteristic. A switch is provided which permits the operator to quickly transfer from the normal audio system to that which incorporates the discriminator. In this way it is very easy to compare an audio system which is being worked on with one having the desired characteristic and in this way determine the changes that are necessary to correct the former.

In addition to the eleven filter stages, the discriminator is provided with a continously variable high frequency cutoff filter. This filter is in no way connected with the band filters and allows a much finer control of the upper limit of the tonal range than does the band filter. It has essentially a vertical cutoff over its entire range from 3500 to 10,000 cycles. This cutoff filter may be switched in and out of the circuit as desired.

Figure #9 shows the overall electrical and sound curves of the phonograph being demonstrated with the discriminator adjusted for essentially flat response from 60 to 7,000 cycles. Listening tests on this instrument with this setting indicated that surface noise from a commercial standpoint was highly objectionable. As a result the discriminator controls were readjusted to give the curves shown in Figure #10. It will be noted that while the range remains essentially the same there is a definite tendency toward a trailing off characteristic above 4,000 cycles. The present recording system employed in Victor records has a slightly rising characteristic in this range and the overall result is one which is very pleasing from a musical standpoint; yet the surface noise is not objectionable. Having determined the desired characteristics, it then was necessary to provide the proper equalizing network to obtain similar performance without the discriminator. Figure #11 shows how closely it was possible to duplicate results. A brief demonstration of the use of the discriminator will be given at this point.







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HIGH FIDELITY RADIO RECEPTION

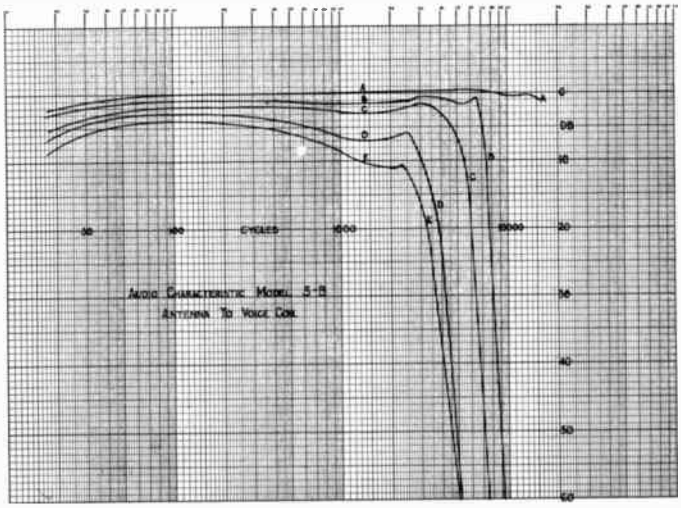
BY

LINCOLN WALSH*

Delivered before the Radio Club of America
December 10, 1936

A high fidelity system might be defined as a system of picking up sounds, transmitting them, and reproducing them so that they sound to the ear precisely like the original sounds. It can also be defined as a system which picks up the sounds, transmits and reproduces them, with all the original sinusoidal components present in their original proportion and phase relation but with the introduction of no new components. If the system does this, the reproduced sound duplicates the original and the system is an ideal high fidelity system. There remains, then the question of how closely a practical system must approach this ideal in order to be satisfactory and this can be answered only by the ear.

Perhaps more than any other sense, the sense of hearing is influenced by suggestion. If the listener has in his mind the conviction that his radio gives him an exact reproduction of the sound in the studio, his ear tells him that there is nothing wrong with the fidelity of his radio. But while the ear of such a listener tells him that all is well with the sound to which he is listening, even when he listens critically to judge the quality of reproduction, an inquiring observer would notice that the extent to which he listens casually to his radio bears a direct relation to its quality of reproduction. Actually, a person will tire quickly of listening to a receiver of poor tone quality, and yet not be conscious of its poor quality. Conversely the same person listening to a receiver of good quality may not be conscious that the quality is better, yet he will listen to it for much longer periods without tiring. It is a matter of common observation that there are many homes having midget receivers of obviously poor quality, whose owners are very proud of the quality and performance of their receivers, and yet those receivers are turned on only for special programs while in homes having receivers of better quality they operate almost continuously. Thus, while the ear is a highly uncritical instrument, it quickly tires of listening to voice and music which is distorted or lacking in important frequency ranges.



The musical instruments of today have evolved thru.centuries of listening, and the test which determined their survival was whether or not the tone was suitable as judged by the ear. These centuries of experience have shown that sounds of all frequencies thruout the wide frequency band of audibility are requisite for musical expression and so, when we as radio engineers design radio broadcasting receivers so that they cut off or seriously attenuate frequencies below about 100 cycles and frequencies above 4500 cycles, we are not only undertaking to give a new interpretation to music but we fly in the very face of man's centuries of musical experience.

The human voice is one sound to which we are always listening, and this experience therefore provides an excellent basis for the ready evaluation of the fidelity of reproduction. Even a receiver of very poor fidelity gives a high degree of understandability, because, as is well known, a range of 300 to 3000 cycles is all that is necessary to give understandability. But to give the naturalness that is necessary if the reproduction is to be untiring to the listener, it is necessary to reproduce the.entire audible range.

A particular instance of the importance of this is the reproduction of soprano voice. We all know how often the re-

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production of the voice of a soprano as heard over the radio is lacking in the qualities that make it a pleasure to listen to the singer in person. This results from the fact that the fundamental frequency is usually far higher than that of average speech, and the harmonics which give the voice its "color" are not reproduced by the receiver. Yet on a high fidelity system all the overtones are present and the reproduction is natural and pleasant.

It is a fact, however, that sometimes a program sounds better if the highs are reduced by lowering the cutoff at the high end. This is invariably due to the presence of distortion or to a high background noise level. In the absence of distortion and noise, any normal ear will choose the highest available cutoff.

The ear like most human senses is subject to habit, and if a person is accustomed by habit to listening to a receiver with a low cutoff, he may not immediately react favorably to a high fidelity system. A receiver having a medium cutoff and a peak near that cutoff, may at first sound to such a person as if it has more highs than a truly high fidelity system, but, again, he will be found to tire quickly of listening to the receiver with a peak. But he will listen indefinitely to the high fidelity system without fatigue.

A listener in a comparison test between a high fidelity system, and a system having a lower cutoff, will very often choose the lower cutoff at first but if he takes time, sometimes as much as an hour or two of listening, he will inevitably choose the high fidelity system. Some listeners when first hearing a high fidelity system by itself think it tinny, some think it very bass, some think it has too much bass and too much treble, and lacking in middle register, because they are hearing tones they are not accustomed to hear in radio reproduction. But after listening for an extended period to good program material they like it. And that is the final, and the only reliable test.

These observations are the result of a systematic study of listener reaction started by the writer some nine years ago. They report the conclusions arrived at after a study of the reactions of something over one hundred listeners only a relatively few of which were radio engineers or serious students of music. On the basis of these observations the writer long ago undertook to develop receiver design details to supply the latent and all too little recognized desire for a much closer approach to complete fidelity in radio reproduction. The results of that work have been incorporated in a receiver typical of the especial arrangements which have been found necessary to meet this need and the performance of that receiver will be demonstrated. Before proceeding to the demonstration, however, it will doubtless be of interest to describe something of the design details and their specific purposes and functions.

THE HIGH FIDELITY RADIO RECEIVER

From the standpoint of present broadcast receiver design practise, high fidelity means extending the audio frequency range at both its ends. It has long been known that to have the tone balanced and most pleasant, the audio range of a system must be centered somewhere between 400 and 1000 cycles. If we extend one end of the range, the other must also be extended for best effect.

THE LOUD SPEAKER

At the low end, it has been found desirable to extend the range of the amplifier to below 30 cycles, not withstanding the limited effectiveness of commonly available speakers in that range. To assist the speaker in this range the largest possible baffle must be used. Olney's work on accoustic labyrinths has pointed the way to improve low frequency speaker response where only limited baffle area is available. An electrical resonance in the circuits, or a mechanical resonance in the speaker - more commonly the latter - is sometimes used to provide a peak at about 100 to 130 cycles, which results in the "boom" of false bass response. In any program, there is always enough energy in the region of this peak to give the low pitched background which many consider to show good bass response. But this causes listener fatigue to develop very quickly and it is thus extremely undesirable to improve bass response by means of any such resonance, either electrical or mechanical. If speaker resonance must be countenanced it should occur at the lowest possible frequency, certainly below 50 cycles; the amplifier must be flat within 2 db down to 30 cycles; the baffle must be of corresponding size, or a suitable labyrinth must be employed; and any bass compensation that is employed must be entirely of resistance-capacity type, to avoid any bass resonance.

THE "TWEETER"

For the high frequency portion of the audio range an especially built "tweeter", which is quite similar to a standard 6 inch cone speaker has been found to supply the best practical solution to the problem presented by the need for efficient translation into acoustic energy of the high audio frequencies. Such a "tweeter" has been included in the demonstration receiver. It has a good sound pressure response curve up to 14,000 cycles, and it is not seriously down at 16,000 cycles. This is secured thru the use of an extremely light voice coil, a short tube connecting the voice coil to the cone, and a light paper cone, of rather low damping.

The "tweeter" is connected thru a small condenser directly across the plates of the push-pull output tubes, so that it cuts in gradually above 2000 cycles. Experience indicates that it is better to have a gradual transition from the low frequency speaker to the tweeter, than any abrupt transition as results from the use of sharp cutoff filters.

While speakers can be built that respond well up to 9000 cycles, the combination of a low frequency speaker and a "tweeter" shows itself to be far superior to any single speaker. One reason for this doubtless resides in the fact that the motion of a large cone diaphragm at high frequencies is made up of two sets of waves radiating out from the voice coil. One wave is longitudinal with respect to the paper of the cone while the other is a lateral wave in the paper, the former travelling at considerably higher velocity that the latter. Their propagation in the paper and their reflection at the edge of the cone determines the high frequency response, and the control of all of these factors is a far more complex and difficult problem than making two speakers of distinctly different proportions each with definite and supplementary characteristics. Additionally where efficient response up to 16,000 cycles is desired no single practical speaker has been found to serve at all satisfactorily.

ELIMINATION OF DISTORTION

Distortion is a very important factor about which volumes could be written. The problem starts at the RF amplifier. The signal voltages at the grid of this tube must be held low to prevent overload and harmonics that will modulate the carrier.

The converter in the demonstration receiver is a special circuit which has a very low noise level. The detector delivers an audio voltage only, and has no relation to the AVC system which is separate. The automatic volume control system holds the detector input voltage at 10 volts for all normal signal inputs, in order to avoid the distortion which occurs in a diode when operated at low voltage levels - within the "parabolic" range - and to avoid distortion in the last IF tube due to overload, which would occur if the diode had to be driven to high voltages.

Additionally, of course, the AVC system holds signal voltages thruout the receiver at such values that will avoid overloading any of the tubes such as the converter and the IF tubes.

It is not, perhaps, commonly appreciated to what degree the use of silicon steel in the interstage audio transformers introduces harmonics, particularly at low levels. But because of this fact the audio signal is carried thru resistance capacity coupled circuits, having no iron core devices anywhere, up to the input of the push-pull output tubes. At this point a push-pull transformer of special design, including a core of high-permeability alloy, which does not generate harmonics is employed, and which contributes greatly to the clean tone quality of the receiver.

All the audio amplification is provided by the use of low-mu triodes, which are the only amplifiers sufficiently free of distortion for a high fidelity system except, perhaps, as the newly developed degenerative circuit arrangements may make the multi element tube less unsuited to this field.

THE I. F. AMPLIFIER

The problems presented by the need for so designing the selective high frequency amplifier stages as to provide for high fidelity reception are basically impossible of solution since, under the American scheme of broadcast frequency allocation and assignment in which adjacent assignments differ by only ten K.C. and practically all frequencies so assigned are in simultaneous use, the requirement that interference-free reception be possible on any assigned frequency at any time and place, unavoidably limits the audio band width of reception to something less than five thousand cycles. Under these limitations there is little that the radio designer can do other than to provide a relatively narrow band width and pray that it will be found not too unacceptable. And, indeed this is precisely the direction in which the receiver designs of recent years have gone.

It is patently absurd, however, to so limit the fidelity of receivers so that only such painfully low fidelity is available to the listener who is located relatively close to his local transmitter or who may be in the high field strength area of a high powered transmitter and thus largely free of adjacent channel interference and interference from noise sources. And since the system under which our radio receivers are distributed to the purchasing public requires so complete a universality of usefulness there is obviously no solution but to give the receiver such variable selectivity as to provide, on the one hand, so narrow a band width as will allow of distance reception in areas of noise and interference and, on the other hand, to provide so great a band width as will allow of the reproduction of the entire audio range being broadcast by the best transmitting system.

In the demonstration receiver this is accomplished in the intermediate frequency amplifier since, as is usual in the superheterodyne type of receiver here largely resides the selectivity of the system. It has been found best to use two IF stages, including three double tuned IF transformers, the coupling between the primary and secondary circuits being varied by moving the secondary coil relative to the primary. At the position of minimum coupling, the coupling is considerably below critical, and the selectivity is at its highest. At the position of maximum coupling, position A in the figure, the resonance curve of the first two transformers becomes double peaked, with the peaks separated by about 35 kilocycles. The third transformer which feeds the detector is broadened, but not enough to show double peaks. The single peaked resonance curve of this latter transformer fills in the valley of the combined curves of the other transformers so that the resultant overall curve of the IF system is flat over a band of about 32 KC, thus permitting the unattenuated passage of sidebands corresponding to all audio frequencies up to 16,000 cycles. A second step of coupling, position B, is provided in the receiver in which similar conditions exist, with, however the band width reduced to about 16 KC corresponding to an audio band of 8,000 cycles. Positions C and D have band widths of 12 and 8 KC respectively, and the fifth position, E, is the position of minimum coupling, and passes about 5 KC. The audio bands corresponding to these are, respectively, 6,000, 4,000, and 2,500 cycles.

It might be well to point out in passing that the use of variable inductive coupling as here employed has certain advantages over other possible types of coupling that might be employed, wholly aside from the obvious advantages of economy of production, ease of production adjustment etc. It will, of course, be remembered that, in general, the peaks in the transmission characteristic of a pair of tuned and over-coupled circuits can be equal only when there is no loss in the coupling element and since mutual inductance is the one coupling element that is loss free, it is thus especially suited to this purpose. Another factor of interest here is that by varying only the mutual by the motion of one of the coils, the band width is varied while maintaining the midfrequency fixed.

R. F. AMPLIFIER

The radio frequency system has as its primary function the elimination of the image frequency, which in the demonstration receiver is 940 KC higher than the signal frequency. It serves also to eliminate such other signal frequencies, as might be brought in thru beating with harmonics of the oscillator. The RF system may, therefore be made broad enough to suit the widest band passed by the I. F. amplifier and need not be of variable band width. The antenna circuit of the demonstration receiver is double tuned, and double peaked, with peaks separated about 35 KC. The RF amplifier is single tuned and is broadened by the use of very fine wire in the winding, so that it just about fills in the valley of the double peaked antenna system. At the higher end of the broadcast band, the valley of the antenna circuit is less deep, and the RF stage is less sharp, so that the RF system as a whole is flat within 1 db over a band of about 35 KC.

This gives a system which will pass all frequencies up to 16,000 without any attenuation. When this system is tried on the air, the tone quality is all that might be expected. But after sunset, when the distant stations on adjacent channels begin to come in, 10 KC whistles are heard. If a filter is put in circuit to cut out 10 KC, the whistle disappears, but then the so called "monkey chatter" is heard. This chatter is the high pitched unintelligible sound which is due to the side bands of the adjacent channel beating with the carrier of the desired signal, in contradistinction to the more commonly experienced interference, known as "cross talk" which is the result of the adjacent channel side bands beating with their own carrier. It is of interest to note that in the receivers which have been built on the basis here discussed, crosstalk from the adjacent channel has been far less serious in creating interference than the adjacent channel chatter. As a matter of experience, it has been found that chatter is the ultimate limitation on fidelity.

It has been found however that, when receiving moderately strong signals from local stations, practically all the chatter can be eliminated by the use of an extremely sharp low pass filter, having its cutoff at about 7500 cycles.

The reason for this is not hard to find if it is remembered that workers in the acoustics and telephone fields have shown that most of the energy in speech and music lies in the frequency range below 2000 cycles, and that the energy drops progressively as the frequency under inspection increases. Thus it is found that there is very little energy in the region of 8000 cycles and higher although that little energy is of great importance in giving naturalness to the sound of which it is a part. Similarly, then most of the energy in the sidebands of a signal is in those frequencies differing by less than two thousand cycles from their carrier and hence by more than 8000 cycles from the carrier of the adjacent channel transmission. Thus, the resulting "monkey chatter" carries most of its energy in the frequencies above 8000 cycles and is thus subject to effective elimination thru the suppression of all frequencies of that order or higher.

THE AUDIO FILTER

To eliminate both the chatter and the whistle in the demonstration receiver, there is used a 4 section Campbell type filter, having 11 elements; 4 series inductances, 5 shunt capacities, and 2 mutual inductances. This filter is flat up to 7200 cycles, and is down approximately 30 db at 8000, 50 db at 9000, and 70 db at 10,000 cycles. These values of attenuation have been found necessary to receive local signals without chatter in the 7500 cycle setting of the fidelity control.

By means of the fidelity control in the demonstration receiver, the band width of the I.F. is varied and thru a simple mechanical linkage with the audio filter, the cutoff of the audio system is simultaneously changed to provide a cut off just below the cutoff of the IF system. This gives the highest possible audio band width for a given selectivity.

In practical use it has been found possible to receive signals as low as 10 millivolts without chatter on the 7500 cycle setting, but with all the superior fidelity implied by that high cutoff as compared with conventional receivers. Weaker signals may require that the cutoff be set at 5500 cycles, which provides better than usual fidelity. When very high selectivity is desired as in cases of high noise levels and in the reception of distant stations, the band is narrowed to 4000 or even to 2500 cycles. This latter band width appears pointlessly narrow but it must be remembered that many listeners want high selectivity and the ability to get distant signals for a short period after they buy their receiver and these circuit arrangements make that possible. Happily, however, after they experience the pleasure of good programs they want little more than the nearby stations, and these they may receive on the high fidelity ranges free of chatter or noise and with highest fidelity consistent with the conditions of transmission and reception.

Experience indicates that in daylight the local stations can usually be heard without whistles, chatter, or any noise on the 16,000 cycle setting, and indeed many stations are transmitting programs of a quality which shows a definite improvement when the cutoff is raised from the 7500 to the 16,000 cycle position. Especially well does this type of receiver operate when receiving the high fidelity stations which have 20 KC channel separation, and good lines and amplifiers, and which are therefore best listened to on the widest band.

FIDELITY OF BROADCAST TRANSMITTERS

It is often argued that the broadcasting stations themselves do not supply programs with such frequency band widths as to justify the use of high fidelity receivers. With this in mind the writer sometime ago made casual investigation as to the upper cut-off of the higher powered transmitters in the New York area and he was happily surprised to note the care which has been taken in the design of much of the broadcasting equipment to maintain the upper frequency limit of the equipment at such a high value as to provide for truly high fidelity transmission. It was found on inquiry, for example, that the studio and transmitter equipment of WJZ and WEAF is good up to 16,000 cycles per second as is that of WABC and WOR. The telephone lines connecting the studios and the transmitters of these several stations are reportedly not as good as the equipment. Thus, the line to the WEAF transmitter is reported to cut off at 7000, the WABC line, at 8500, the WJZ line at 10,000 and the WOR line at 11,000. These latter data on the cutoff frequency of the telephone lines are not to be viewed as inherently limiting the broadcasting system since, they doubtless result largely from economic and not technical factors, and can and doubtless will be raised as the demand for such improvement developes. It must, however, be admitted that there is little incentive for any broadcaster to assume the added burdens of cost and maintainance required by any further expansion of the band width of his transmission until and unless the receivers in the audience to his transmissions are capable of taking full advantage of that transmission.

Thus, once again we have arrived at the stage in the development of broadcasting where the transmitter leads the receiver and now awaits being overtaken. Throuot the history of broadcasting, first the receiver and then the transmitter has been the more thoroughly developed element of the system and in each step in the sixteen years of progress in radio broadcasting each improvement in the lagging element has placed it so markedly in advance of the other as to provide the incentive for major improvement in the other which in turn prompted further improvement in the first, and so on and so on around the widening spiral of progress.

It seems quite reasonable, therefore, to believe that not only will the next move in this continued progress be made by the receiver designer-in the direction here described in detail - and with the agreement of his commercially minded associates, of course - but that that next step will leave the broadcaster lagging once again only, however, to have him soon moving again in the direction of the ultimate perfection of broadcasting.

EDITOR'S NOTE

In the course of the delivery of Mr. Walsh's paper demonstrations were made in connection with a special program for the Club from radio station WQXR under the direction of Mr. John V. L. Hogan. This program included a variety of broadcasting material designed to show the superior effectiveness of the high fidelity reception through the transmission of a wide range of tones and special musical transmissions both with and without the use of cut off filters at the transmitting station. It can be reported that any cut off at the transmitter that tended to reduce its transmission band width below the maximum possible was quite easily evident in the reception as heard by those in attendance and it was of especial interest for most of the members of the club present to make comparisons between the reproduction of Mr. Hogan's voice and their memory of it as it so often when Mr. Hogan attends in person. The demonstration left no doubt as to the need for the transmission of the entire band width of the receiver and the transmitter where completely satisfactory fidelity is the aim. Additionally it should be pointed out that while this demonstration was made at Havemeter Hall, Columbia University at which previous experience has shown that the noise level is always objectionably high, the installation of a noise suppression antenna system, through the kindness of Messrs Amy and Aceves and King, left little noise interference to detract from the demonstration or from the enjoyment of the special musical program that was transmitted.

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Volume 14 | August, 1937 | No. 2

THE SURFACE WAVE IN RADIO PROPAGATION

BY

CHARLES R. BURROWS*

Delivered before the Radio Club of America
February 11, 1937

Mr. Burrows opened his paper with a discussion of the concept of the surface wave and its relation to radio propagation. He pointed out that by surface wave is meant a wave that is guided by a surface in generally the same manner that a wave is guided by a pair of wires or by a concentric tube transmission line or even by a hollow pipe. These were given as examples of what is termed one dimensional surface waves. A two dimensional surface wave can be thought of as that which would result from transmission between a pair of planes. In the one dimensional case, the energy is attenuated exponentially by absorption and thus results in the familiar expression of attenuation in decibels per mile. In the two dimensional case, however, in addition to the exponential attenuation factor, due to power absorption there is a decrease in the energy density in the wave as a result of the spreading of the wave over an ever increasing area as it advances. This reduces the energy density inversely with the distance and the field strength therefore varies inversely with the square root for the distance.

The concept of the surface wave was introduced into studies of radio propagation over the surface of the earth in 1907 when Zenneck showed that the interface between earth and air could support a plane surface wave that was exponentially attenuated in the direction of propagation and decreased exponentially with increase in distance from the surface both upwards and downwards. Zenneck did not show that an antenna could generate such a surface wave but because it offered a plausible explanation of radio transmission to great distances it was generally accepted.

Two years later Sommerfeld considered the problem of the spreading of electromagnetic waves from a short doublet antenna located in the interface between earth and air. He expressed his result as the sum of two components, one of which he identified as a cylindrical surface wave which at great distances was equal to Zenneck's surface wave. This theoretical work of Sommerfeld seemed to prove that the surface wave was an important component of the radiation from an antenna on the surface of the earth.

Ten years later Weyl reconsidered the problem and obtained an expression for the radiation from an antenna on the surface of the earth which did not explicitly contain the surface wave. From this he concluded that the separation of the field by Sommerfeld into two components, one of which was called the surface wave, was merely mathematical fiction and had no physical counterpart. He was, however, apparently of the opinion that his results agreed numerically with those of Sommerfeld.

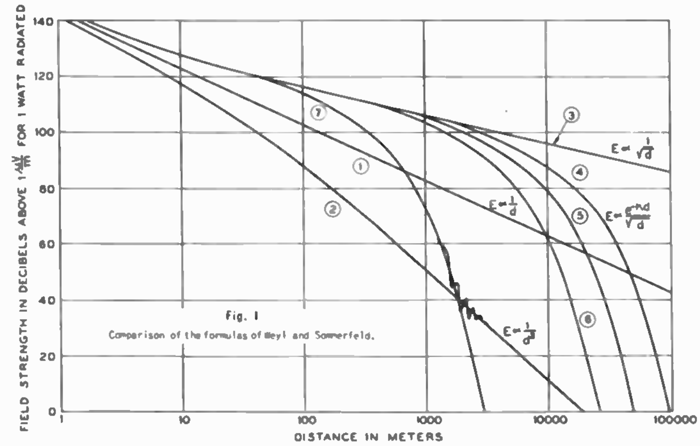

Fig. 1
Comparison of the formulas of Weyl and Sommerfeld.

Mr. Burrows here pointed out that a careful comparison of the results of the work of Weyl and of Sommerfeld shows that they differ by precisely the surface wave component in question which was the subject of his paper. The comparison of the formulas of Sommerfeld and Weyl are indicated by Figure 1 where the field strength is plotted as a function of distance both scales being on logarithmic. From this figure it was pointed out that for transmission over a perfectly conducting plane the field strength varies inversely with the distance as shown by curve 1. Curve 2 was de-

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scribed as a plot of Weyl's formula for transmission over a dielectric plane. This shows the field strength varying inversely with the square of the distance at the greater distance. Curve 3 is a plot of the surface wave for propagation over a perfect dielectric showing the field strength varying inversely with the square root of the distance at the greater distances. Where the dielectric is not perfect but has appreciable conductivity the surface wave decreases exponentially with distance as indicated by curve 4. Curves 5, 6, and 7 having been plotted for increasing values of conductivity show that as the conductivity is increased the marked influence of the exponential factor sets in at shorter and shorter distances. The Sommerfeld-Rolf curve results from adding the surface wave component to the Weyl curve. The interference of these two components produces oscillations in the curves where the two components are approximately equal. Under the condition where these two components are equal and out of phase at the same distance the theory of Sommerfeld predicts zero field strength at a finite distance as pointed out by Rolf.

Thus, Mr. Burrows stated, resort to experiment was indicated as being desirable in order to decide which of these two curves is correct. In making such an experimental investigation, however it is highly desirable to make transmission tests under conditions where the received field strength predicted by these two formulas differ greatly. This occurs for propagation over a perfect dielectric and since fresh water is the nearest approach to a perfect dielectric available in sufficient volume and area for a test of this kind the locale of making the tests was largely determined by this fact.

The departure of these two formulas from one another increases also with the frequency so that a most revealing test would comprise a determination of the variation of the field strength with distance over fresh water in ultra high frequency transmission. Thus a frequency of two meters was chosen as a convenient and useful frequency for the work reported by Mr. Burrows.

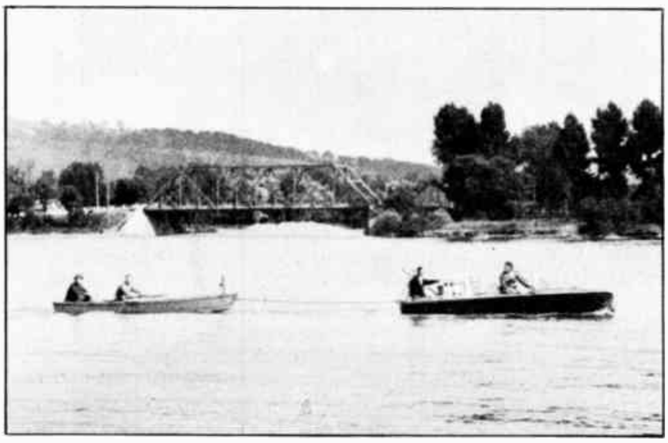

Fig. 2

Experimental arrangement for determining the variation of the received field strength with distance.

During the summer of 1936, Messers Burrows, Decino and Hunt took some two meter measuring equipment to Seneca Lake, New York State and there made these tests. Figure 2 shows a picture of the experimental arrangements. The transmitter was carried in a rowboat towed by a motor boat containing the receiver. The antennas were loaded quarter wave doublets whose midpoints were a quarter wave-length above the water's surface. The experimental procedure was to drive these boats along path 1 of the Figure 3 at a fixed distance apart for a sufficient length of time to make certain that there was no fading such as might be produced by reflections from the bottom of the lake. The distance between the transmitter and receiver was measured by a fishline under a fixed tension.

For distances between receiver and transmitter greater than 150 meters, it was necessary to alter the experimental procedure. For these distances the receiver was located at the end of a pier at Hector Falls and the

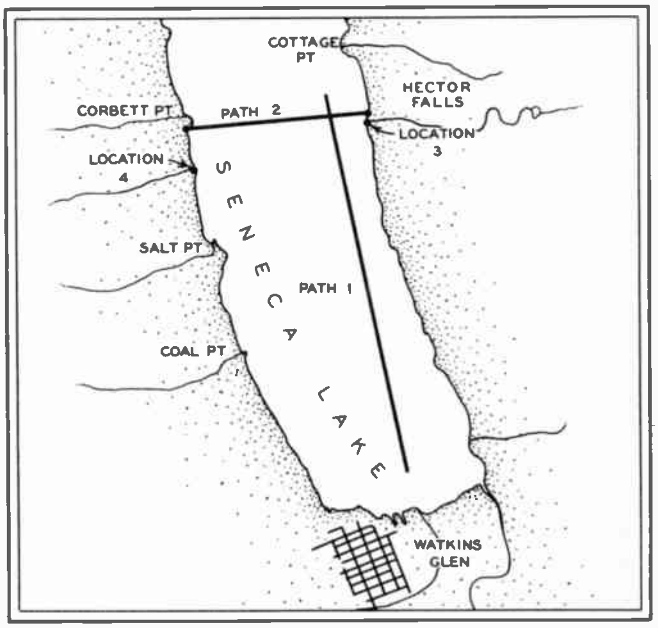

Fig. 3

Map of part of Seneca Lake showing the location of the experiment. Path 1 shows the location of the two-boat experiment. Path 2 the one-boat experiment. Locations 3 and 4 indicate the positions of the terminals for the variable height test.

transmitter was located in the motor boat which was driven along path 2 of Figure 3. This, of course, introduced considerable difficulty in measuring the distance between receiver and transmitter. To minimize the error in this measurement, it was reported by Mr. Burrows that three independent methods were used. First, the motor boat was driven at constant speed in a fixed direction between two points a known separation. Second, the distance between the transmitter and the receiver was measured by means of a transit located at the receiver and a stadia rod carried by the boat. Third, a sextant was used to measure the angle subtended at the boat by two poles located on the shore, one near and the other at the receiver. To complete the measurement the angle between the line joining the two poles and the direction of the boat was measured by the transit.

Figure 4 shows the experimental data so obtained. The solid circles represent data obtained when using the two boats and the open circles those obtained when the receiver was located on the end of the pier. Curve 1 shows the inverse distance variation that would result from transmission over a perfectly conducting plane. Curve 2 is a plot of the Weyl's formula for transmission calculated for water of the characteristics of that of Seneca Lake. Curve 3 is a plot of Sommerfeld's formula for transmission over Seneca Lake water. Curve 4 is a plot of Sommerfeld's formula for transmission over a perfect dielectric.

It will be noted that the experimental data is in good agreement with values calculated by Weyl's formula which as Mr. Burrows reiterated *does not* contain the surface wave component.

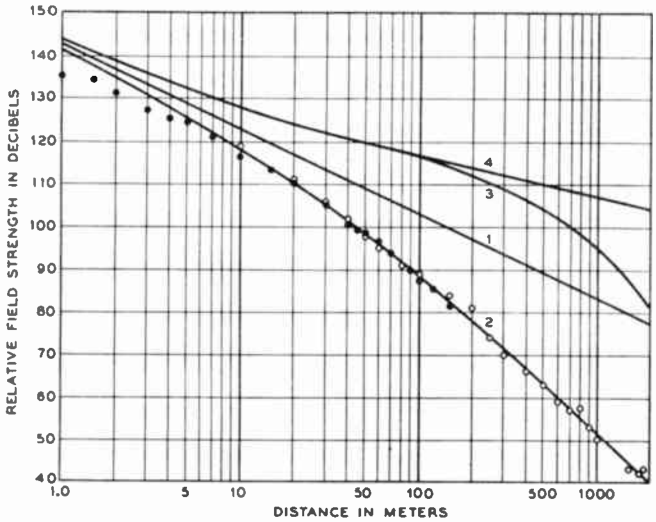

Fig. 4

Comparison of experiment and theory.

The values shown by the theoretical curves depend, of course, upon the distance, wavelength, dielectric constant and conductivity all of which, therefore required evaluation. The methods used in measuring the distance between transmitter and receiver have already been described. The measurement of the wavelength with the required precision introduced no difficulty and since the dielectric constant of water is a known function of its temperature the evaluation of the dielectric constant required, simply, the measuring of the water temperature. The conductivity was determined by laboratory measurements of samples of the water.

Then, leaving the matter of the variation of field strength with distance, Mr. Burrows pointed out that there is another property of a surface wave that might, with interest, be observed experimentally. This is the variation of the field strength with height above the earth's surface. That is, if the field strength is measured over a range of antenna elevations at distances where the surface wave, if any, would be large as compared with the remaining component, there is afforded additional experimental information indicating whether or not the surface wave exists. Accordingly, portable antenna masts were erected at opposite sides of Seneca Lake. The received field strength was determined as a function of the antenna height for two antenna heights at the other terminal as shown in Figure 6. The solid

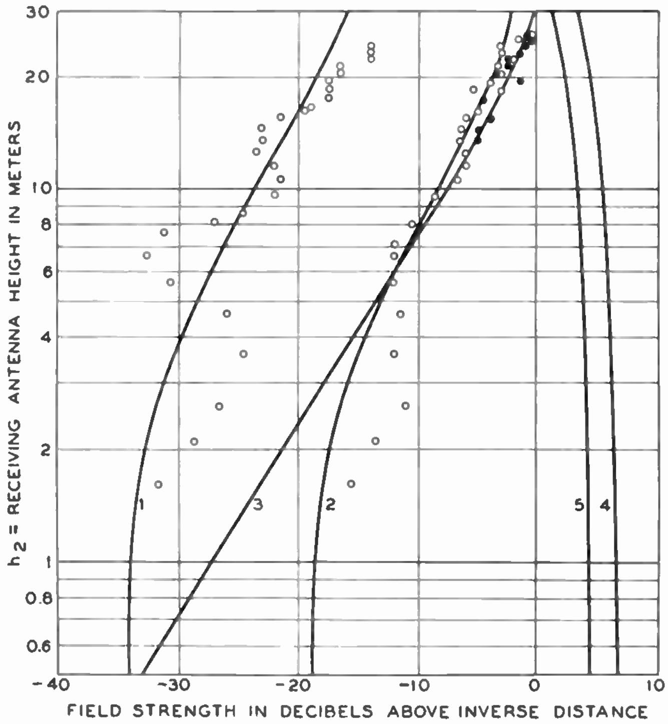

Fig. 6

Variation of received field strength with antenna height.

circles represent data taken on horizontal polarization with one antenna at 24.8 meters above the surface of the water. Since there is no uncertainty in the formula for the field strength with horizontal antennas these points may be used as a calibration of the equipment. Curve 3 is a plot of the formula for the received field strength with horizontal antennas. When this curve is made to fit the experimental data the locations of curves 1, 2, 4 and 5 are fixed. Curves 1 and 2 show the variation of received field strength with vertical antennas that would result if there were no surface wave for the two transmitting antenna heights. Curves 4 and 5 are plots of the surface wave for these two conditions. It will be noted that there is no semblance of agreement between that data and curves 4 and 5. The experimental data does, however, agree with that of curves 1 and 2. These latter are, of course valid only if there is *no surface wave component* either in absolute magnitude or in the variation of magnitude with antenna heights.

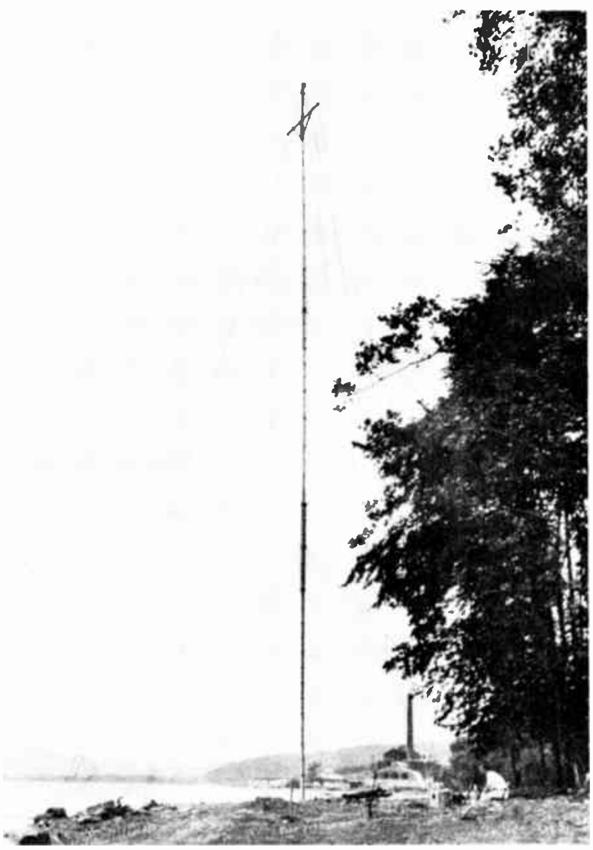

Fig. 5

Picture of transmitting site at "4" of Figure 3, showing the portable 25-meter mast and transmitting antenna.

Mr. Burrows called attention to the fact that the oscillations in the experimental data are presumably due to reflections from the hills and cliffs which line the

Lake and from trees near the receiving antenna. He pointed further that this experimental evidence proves conclusively that simple antennas *do not generate a surface wave* and therefore the Sommerfeld-Rolf formulae and curves require revision for all conditions where the dielectric constant cannot be neglected.

To further indicate the departure between the two sets of concepts discussed by him, Mr. Burrows showed figure 7 which compares the Sommerfeld-Rolf curves with

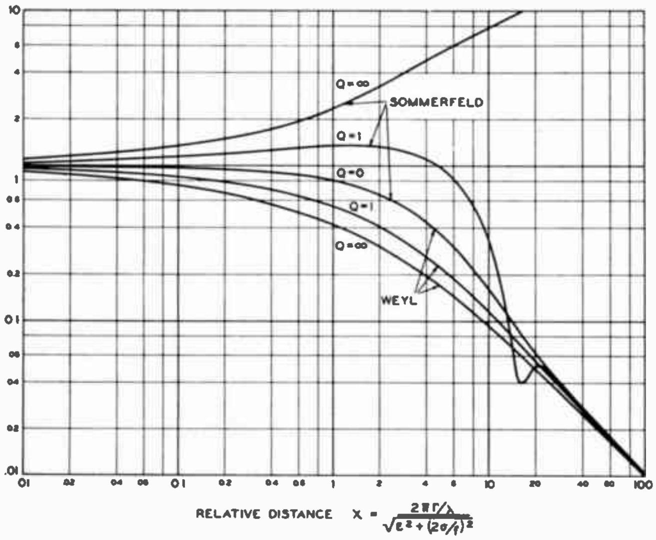

Fig. 7

Comparison of Sommerfeld-Rolf curves with the Weyl-Norton curves.

the new Weyl-Norton curves the validity of which were established by the experimental work which he described. The attenuation factor is plotted against the relative distance. In this the attenuation factor is defined as the factor by which the field strength that would result from transmission over a perfectly conducting plane must be multiplied to give the field strength under the conditions of interest.

For transmission over an imperfect conductor in which the conduction current is large compared with the dielectric amount the two formulas agree as indicated by the curve marked Q = 0.

When the conduction current is equal to the dielectric current (Q = 1) the Sommerfeld formula indicates an attenuation factor greater than unity up to a certain distance, while according to the Weyl formula the attenuation factor is somewhat less than that for the conductivity case. For transmission over a perfect dielectric (Q = oo) the Sommerfeld formula indicates that the attenuation factor is always greater than unity and increases indefinitely with increase in distance, while the Weyl formula indicates the attenuation factor is only slightly less (up to about 10 db.) than that for the conductivity case.

In conclusion Mr. Burrows stated that the validity of the Weyl formula has been unquestionably established by the work reported in his paper as against the previously generally accepted Sommerfeld formula.

The discussion that followed the reading of the paper brought to light much that was of interest to the membership. It was the first reaction of those in attendance that it was to be assumed from Mr. Burrow's conclusions that the concept of the radio wave as being a wave moving forward with its "feet in the ground" would have to be abandoned and it was quite evident from the comments that the attendants were by no means completely willing to abandon this old and generally useful concept unless something more useful could be supplied in its stead.

No alternate concept was offered and, indeed, it was not insisted that this simple concept need be abandoned even in view of the apparent need for abandoning the conclusions usually drawn from the Sommerfeld formulae. Instead it was repeated that the important aspect of the conclusions from the work reported by Mr. Burrows concerns itself with the magnitudes of the field strengths to be expected under practical operating conditions as compared with those to be expected under the Sommerfeld formula.

Of major importance in this is the fact that this abandonment of the Sommerfeld formula brings with it the realization that the field strength of any transmitting station is less effected by the Q of the ground than had heretofore been supposed and that, in fact, there is no great advantage in transmission over a perfect dielectric as has long been assumed to be fact.

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EXPERIMENTS IN GENERATION, DETECTION AND MEASUREMENT AT ONE METER WAVELENGTHS

BY

PAUL ZOTTU*

Delivered before the Radio Club of America
March 11, 1937

Mr. Zottu introduced the subject of his paper by pointing out that the problem of developing oscillators, amplifiers and the like for use at extremely high frequencies - below a wavelength of one meter - is commonly thought of as being largely one involving the development of circuit elements which will provide impedances of sufficiently high value. Thus, the first step in the development of ultra-uhf equipment is the development of high impedance coupling circuits. He disposed of the possible usefulness of conventional types of tuned circuits by pointing out that while they held considerable promise, at the moment the use of lines of various types recommends itself as being the obviously and immediately useful type of arrangement for oscillator, stabilization, etc. He limited his further discussion to the closed quarter wave line as being the obviously useful type for providing high impedances and high Q for tube couplings. In this he defined the Q of a quarter wave resonator as F. E. Terman defines it in his article of July 1934 in Electrical Engineering, that is, by analogy with the Q of a lumped circuit. This gives an expression for the Q of such a line as

$$Q = \frac{2\pi Z_0 f}{RC}$$

where R is the resistance per unit length of the line, C is the velocity of light, Z_0 is the characteristic impedance of the line and f the frequency at which it is excited.

Mr. Zottu then proceeded to a discussion of the theoretical factors operating in the choice of dimensions of both open wire and concentric lines for maximum or optimum end impedance and Q. These provide certain optimum dimensional proportions, the optimum relationship between conductor spacing, diameter and wavelength in the case of open lines and the ratio of conductor diameters and wave length in the concentric lines being only approximately equal for maximum end impedance and for maximum Q. In this it was suggested that some difference of opinion exists between the speaker as the result of his experimental work and other workers in this field as the result of their analytical work, as to the influence of radiation resistance on these optimum relationships in the case of concentric lines. There was much discussion of this point after the paper, which might well be mentioned here: the point being that while it was obvious that in the case of open wire lines radiation must be taken into consideration since, as the lines are spaced further and further apart in order to get higher surge impedance and lower attenuation, the radiation resistance increases as a result of the increasing amount of power lost through radiation as the spacing is increased, so that one comes ultimately to a spacing between conductors beyond which the increase in power loss through radiation more than offsets the increase of surge impedance and reduction of assymetrical current distribution in the conductor, i.e. skin effect.

Figure 1 indicates the magnitude of both the purely resistive end-impedance and the Q of open wire lines of optimum dimensions. From this it appeared that Q's of about one thousand and end impedances of several hundreds of thousands are readily possible at 100 M. C. when using open wire lines.

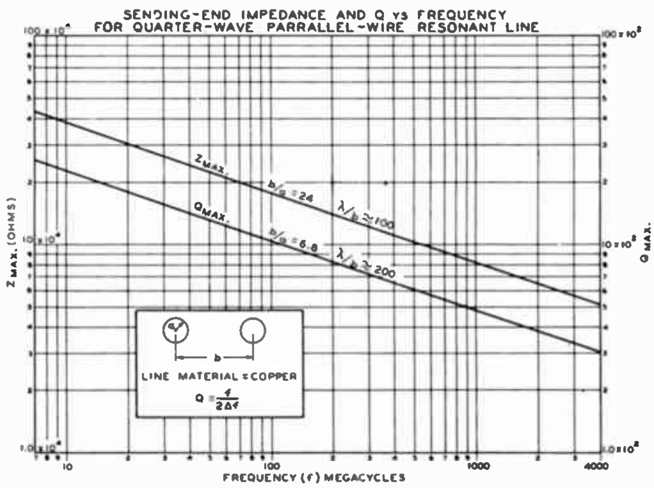

Fig. 1

Similar reasoning, including the consideration of radiation loss, if any, provided the basis for the determination of optimum proportions for concentric lines. Figure 2 shows the maximum end impedance and maximum Q over a wide range of frequencies - 1. to 400. M. C. - for optimum proportions. This indicates that end-impedances and Q's are about ten times as great as those of open lines. On the matter of radiation loss within a closed

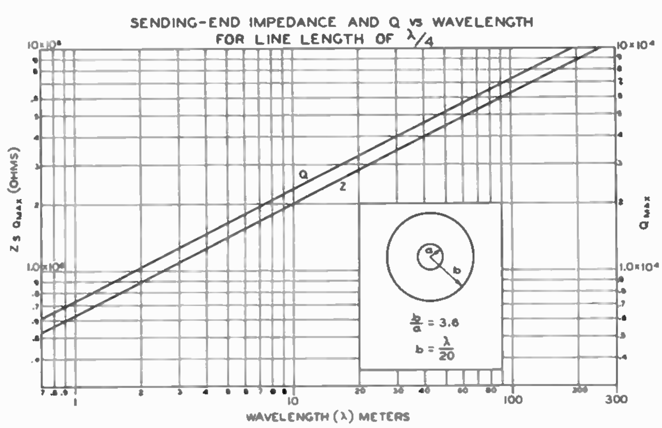

Fig. 2

*Engineer, R. C. A. Manufacturing Co.,
Harrison, N. J.

concentric line, the later discussion, as invited by Mr. Zottu's comment, waxed hot and heavy: the point being that Mr. Zottu had limited his analytical work to such dimensions of conductors that the radiation factor proposed by certain other unnamed workers in arriving at the optimum relationship was negligible notwithstanding the fact that he was unconvinced of the rationality of the assumptions basic to this point. The point at issue here appearing to reside in the fact that as the ratio of the diameters of the conductors of a concentric line gets great and as the difference between them gets to be an appreciable portion of the wavelength, the assumption of the instantaneity of the building up of the radial flux within the line introduces an increasing error and one which would doubtless limit the ratio of diameters to some finite value if valid correction is made for it. Mr. Zottu felt that such correction as proposed does not however meet the needs of the problem.

It might here be pointed out in passing that from the purely analytical viewpoint the lack of instantaneity of development of flux within a concentric line results in a departure from the precise in-phase or quadrature relations indicated by the analysis of the properties of loss-free lines and thus probably contributes to the characteristics of the line under consideration precisely as does the presence of losses in the line. Whether such a loss-like relationship in the properties of the lines discussed by Mr. Zottu actually means loss by radiation or otherwise appeared of only academic importance. The influence of this factor on the end-impedance and on the Q of the line is, however, of major practical importance from the view point of Mr. Zottu's paper but, unfortunately, was not completely included in his presentation of the problem.

It was brought out in a later section of Mr. Zottu's paper that in the case of an oscillator including a concentric line closed at both ends and including within itself all circuit elements including the tube, he was able to find no evidence of radiation or other fields external to the line.

Mr. Zottu next proceeded to a discussion of the application of the relations shown by his graphs. He indicated that a quarter wave line at any but the highest frequencies requires far more space than is likely to be available under practical conditions. Thus, as he pointed out, in the broadcast band for nearly optimum dimensions something like a pair of smoke stacks, something over three hundred feet high, would be required for the stabilizing of a broadcast transmitter by means of an "open wire line" while a pipe, sixty feet in diameter and three hundred feet long, would be required to serve as the outer conductor of a suitable concentric line. This pointed the obvious limitations of simple quarter wave lines and introduced the compromises that he has found he could make and still get the desired high impedances and high Q. In general, this was accomplished by two different means. The first and most useful appears to have been the use of a line of approximately optimum cross sectional dimensions but of only a small fraction of a quarter wave length terminating in a capacity. The capacity required for this purpose is secured by putting a cap over the end of the line and terminating the inner conductor in a large flange and making the cap movable relative to the flange for purposes of "tuning". It was not pointed out by Mr. Zottu, but it is in fact a moot question whether such a structure consisting of an extremely short line and terminating in a capacity is really a "line" or whether it is merely an extremely low loss and conveniently constructible inductor and a capacitor.

At any rate, the short line with capacity termination was one form that Mr. Zottu's circuit arrangements took in order to secure high impedance for the tube couplings.

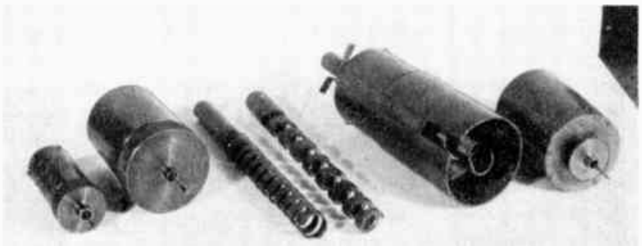

Fig. 3

The second form which was shown, Figure 3, is a concentric line in which the inner conductor consisted, not of a simple cylindrical member, but of a spiralled conductor of rectangular section; this, in the interest of raising the surge impedance, decreasing the axial velocity of propagation and hence, making a quarter wave line short enough to be useful. Because of its greater complexity this type of line was not much used.

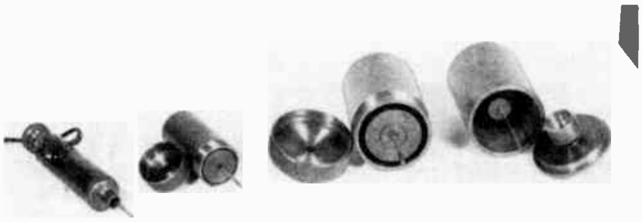

Fig. 4

Samples of a number of different sizes and types of these several forms of shortened lines as shown in Figure 4 were available for inspection after delivery of the paper.

In these forms of circuit construction provision has to be made for the coupling to the tubes or associated circuits. Such coupling was necessarily made variable so that "matching" would be readily possible. In general, this was accomplished by providing slots in the outer conductor near the low potential end through which conductors terminating on the inner conductor could pass for connection to the tube or other circuits. Where used with tubes this provided either for "back-coupling" of grid and plate circuits in the case of oscillators or plate-circuit-to-grid circuit coupling in the case of amplifier arrangements. In these practically useful forms of short lines provision for tuning was in each case made by an adjustment knob at the low potential end either for shorting portions of the spiralled inner conductor or varying the capacity at the remote end of the line, or for varying the effective length of the line itself as is shown in the figures.

After an extended discussion of these details, Mr. Zottu showed an oscillator consisting of eight tubes arranged radially on a circular metal plate, each tube carrying its own plate and grid circuits each consisting of a single loop of wire and coupled into a rather unusual form of short concentric line. This line consisted of an inner conductor of what was termed a concentric line, lacking however, the outer concentric conductor: the latter being replaced by three cylindrical standards of small diameter supporting the end plates of the line, one of the end plates comprising one of the plates of the line terminating condenser. The connections between the tube circuits and the line which acted as a single tank circuit for all eight tube circuits being provided by taps on the inner conductor connected through coupling condensers - either adjustable or fixed - to the single loop plate or grid circuits of each tube.

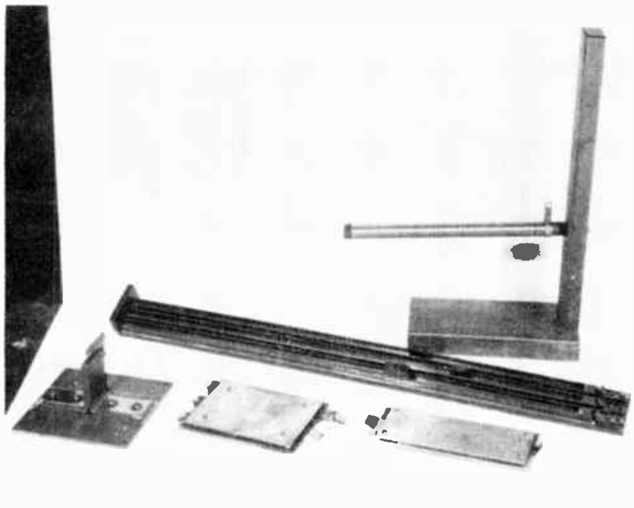

Fig. 5A

This was reported to give a power output of the order of 80 watts at about one meter.

Mr. Zottu explained at some length how this radially symmetrical structure might be extended by the use of a line tank circuit not so greatly foreshortened so as to provide for its being fed by a number of levels of radial tube generators and suggested further extensions of this scheme of symmetrical layout in the interest of making it possible to use tubes as oscillators at frequencies near the upper limit imposed on their operation by their individual impedances and thus, avoiding the paralleling of their capacities and impedance and the consequent limitation on the upper frequency of their operation. It was pointed out that one could, by the arrangements shown, operate an almost unlimited number of tubes all feeding into a single tank and thence to a single dissipator at the same high frequency as that at which the tubes would operate individually.

Mr. Zottu then reviewed the acorn tube oscillators, shown in the B. J. Thompson papers of sometime ago. In this he brought out a distinction between the original Thompson tubes and the now commercially available acorn tubes - the 955, etc. - by referring to the original and still experimental tubes, as "shoe button" tubes.

This portion of the paper stressed some of the details of the design of circuit elements such as the inclusion within the tube mounting of the required bypassing capacities by the simple expedient of mounting the socket elements on separate metal plates which were separated from the datum metal plate base of the assembly by thin dielectric in the interest of getting high bypassing capacity. Additionally, various combination current and voltage supply leads were shown in which thin sheets of dielectric were used between the strip or thin plate leads in the interest of getting by-passing.

In addition to bypasses, Mr. Zottu showed an interesting

METHOD OF MOUNTING AND BY-PASSING ACORN TUBES

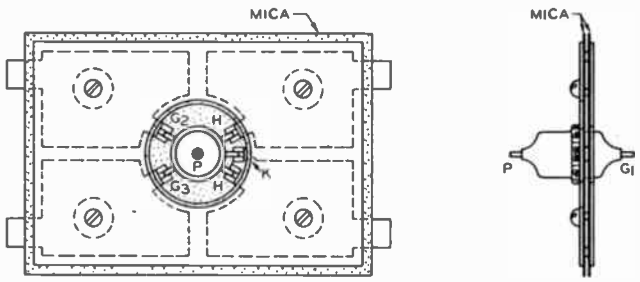

Fig. 5B

"universal" choke coil the operating range of which was from about two meters down to one half a meter. It consisted of a single layer solenoidal coil of small diameter - about half an inch in diameter - and about four to six inches long wound with a single layer of wire, of either copper or higher resistance material. The theory of its usefulness being that a coil of a sufficient number of wavelengths long and of sufficient attenuation will provide so little reflection from the receiving end back to the transmitting end that, as viewed from the transmitting end, it would offer only its own surge impedance at all frequencies. In addition, other forms of lines of use as chokes were shown, amongst them simple adjustable lines comprised of a small channel section as the outer conductor and having a concentric wire as the inner conductor and a slider to provide H. F. short-circuiting of the line to the channel section and thus to provide the adjustability necessary for making the end impedance of the line high and largely resistive. These as shown in Figures 5A and 5B and other similar devices were found necessary in oscillator and amplifier designs to maintain filaments and other elements requiring D.C. excitation at potentials determined by the H. F. requirements of the circuit and independent of the potential of the current and voltage sources.

Several other forms of oscillator were shown: amongst them one, shown in Figures 6A and 6B in which an acorn tube was mounted within the concentric line and provision made for changing the coupling of the line to the tube through the shifting of the line as a whole relative to the tube and for tuning the line by means of a break in the inner conductor which was closed through a dielectric adjustable as to capacity from outside the closed line.

Mr. Zottu then proceeded to a discussion of such measuring methods and devices as had been devised for work in this field. It was pointed out that with the means for measuring voltage and power available, most desired characteristic data could be secured. There were, therefore, two general types of measuring instrumentalities devised. The first to be described was the thermocouple type of watt meter and the second was the diode rectifier type of peak voltmeter. Of these the first took the form of a number of different types of vacuum

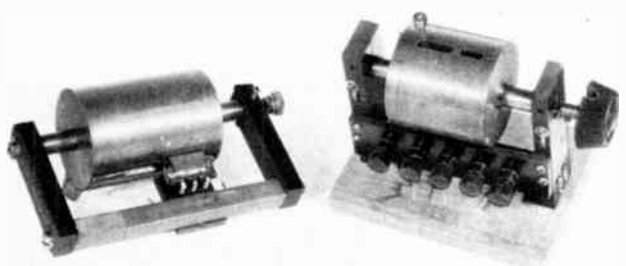

Fig. 6A

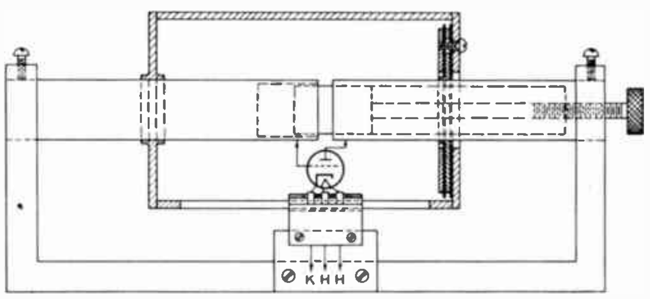

Fig. 6B

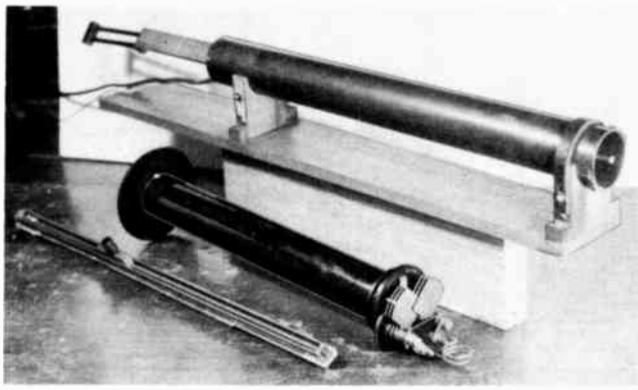

Fig. 7A

type thermocouples in which all the available power was dissipated in a heater and the resulting thermoelectric voltage read externally on a conventional indicating instrument, the arrangements of the heater and couple elements and terminal arrangements being such as to minimize the coupling between the heater and the couple circuits. In order to provide for rather unusually high heater temperatures the heater itself was made of extremely fine tungsten wire, while the nature of the couple conductors was undescribed. Calibration curves of several typical thermocouples were shown indicating that for the most sensitive thermocouple watt meter the system gave full scale deflection at about 40 milliwatts while for the least sensitive couple maximum power indication was gotten at about 40 watts.

Photographs of a number of diodes devised for use as diode peak voltmeters were shown. These were all characterized by extremely small anodes and filaments with microscopic clearances between them. Thus, the smallest of the diodes shown included a cylindrical anode six mils in outer diameter enclosing a filament of unspecified diameter. The circuits used with these diodes are quite conventional in that the diode output circuit comprises a condenser which is charged by the unidirectional electron current to such a potential as to reduce the electron current to a negligible amount. It was pointed out that the minute clearances were made essential by the influence of the electron transit time upon utility of the diode for this purpose. Even the smallest of the diodes shown required correction for frequency at the extremely high frequencies at which they were used. Thus, a calibration curve for this diode was shown in which the 60 cycle calibration departed little from the calibration at one meter.

A brief discussion of the relation between electron transit time - or the phase shift of the diode current and voltage due to transit time - and the current flowing in the diode served to indicate the procedure followed in making the frequency corrections to the diode voltmeter characteristics.

A third type of measuring instrument was shown in the slides: this, of the wave meter type. Two specific types were indicated and are here shown in Figures 7A and 7B. One of them consisted of a few turns of wire connected to a variable condenser of the Remler type presumably in order to secure complete symmetry to ground in the tuned circuit and to maintain a low minimum capacity, the LC combination was mounted at the end of a long dielectric rod or tube with the control and scale mounted at the end remote from the inductive elements. No resonance indicator was included in this type of wavemeter; thus, it was useful only for "absorption" measurements.

The second type consisted of a concentric line in which the effective length of the conductors was made vari-

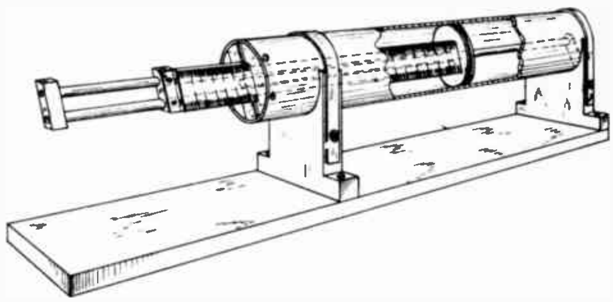

Fig. 7B

able through the provision of a sliding member within the line contacting both the inner and the outer conductors with a moving scale carried along with the sliding member and itself sliding under an indicator external to one end of the line, thus indicating wave length directly on the linear scale; the entire assembly being provided with a removable end so that it might act as an open or shorted line, thus providing for two wave length ranges.

Mr. Zottu then proceeded to a discussion of the adaptation of these types of structure to selective receiver circuits in conjunction with acorn tubes. He showed illustrations of several types of receivers already shown on previous occasions, notably in connection with the Thompson IRE papers in which conventional tuning arrangements of small size were used and pointed out that the gains per stage gotten by these means were small even at the lowest of the high frequencies at which the receiver operated; and, indeed, were never in excess of four per stage. This low gain appears to have provided the impetus required for the attempts to adapt the "line" type of tuned circuit to receiver uses and resulted in a three line receiver which was shown both in slides and "in the flesh" after the reading of the paper and is here shown in Figure 8. The problem to be

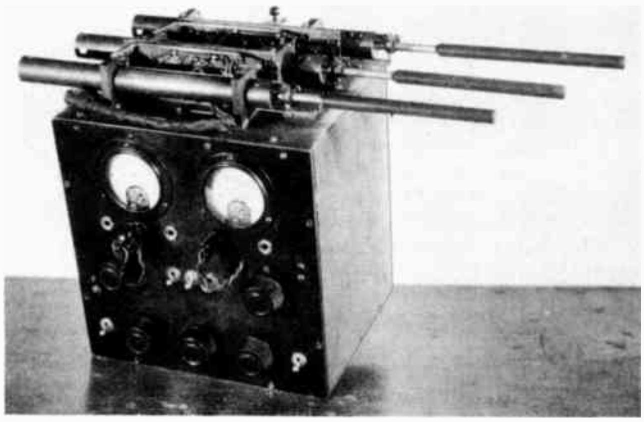

Fig. 8

faced in the design of the receiver was reported to have been one of providing such a wide range of adjustability as to accommodate a wide and apriori unknown range of tube impedances and such a wide frequency range as would determine the ultimate limits of the receiver's effectiveness. Thus, not only were the lines made adjustable as to length for a wide range of frequency variation but adjustable with respect to the point of connection along their lengths to the tube elements. Each of the three assemblies consisted of a yoke carrying a sliding line the tube being mounted on the former while a sliding member within each line provided for the adjustment of the effective length of each of the three lines to suit the desired frequency of operation. Thus, the tube in each one of the three assemblies comprising an amplifier stage could be connected to its own line on the plate side at the proper point along the length of the line by the sliding of the line within its yoke and to the adjacent line at the proper point along the length of that line by the sliding of the adjacent assembly, including the yoke and all and, in addition, the frequency of operation of each of the three stages was adjusted by sliding the short circuiting member within the line housing.

Mr. Zottu pointed out that. as was to be expected, when coupling the tube plate circuit to such a high impedance as was provided by the line, it was found necessary to make the plate-to-line connection relatively close to the low potential end of the line in order to secure low enough coupling to give the desired selectivity of operation but, as was not so definitely expected, it was found necessary to make the grid-to-line connection also near the low potential end and, even less expectedly, found necessary to make this connection even nearer the low potential end than the plate connection if anything approximating a desirable degree of selectivity was to be obtained. Thus the tuned R. F. receiver comprising the tubes and lines consisted of three sharply tuned circuits coupled loosely out of the tube plate circuits and even more loosely coupled into the tube grid circuits.

It was pointed out that the unhappily low input impedance was not due, as is sometimes suggested, to the capacitive reactance of the tube input but in large measure to the low value of the input resistance - or high conductance - which, in turn is due to the relatively high ratio of the electron transit time to the period of the circuits. In fact, it was shown that at a point in the range of the receiver *not* close to the upper limit of its frequency range, the input resistance of the tubes became so low that the gain of the receiver was reduced to unity.

Mr. Zottu's conclusion to his most interesting and instructive paper indicated that the solution to the problems that so seriously limit the development of suitable radio receivers, oscillators and other radio devices employing vacuum tubes for use at extremely high frequencies has been shown by the work reported in this paper *not* to reside in the design or construction of the circuit elements but rather in the design and production of vacuum tubes. Indeed, one understood from the paper, that further progress in the field reported on by Mr. Zottu was unquestionably and completely dependent on the development of new and suitable types of vacuum tubes.

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THE SURFACE WAVE IN RADIO PROPAGATION

BY

CHARLES R. BURROWS*

Delivered before the Radio Club of America
February 11, 1937

Mr. Burrows opened his paper with a discussion of the concept of the surface wave and its relation to radio propagation. He pointed out that by surface wave is meant a wave that is guided by a surface in generally the same manner that a wave is guided by a pair of wires or by a concentric tube transmission line or even by a hollow pipe. These were given as examples of what is termed one dimensional surface waves. A two dimensional surface wave can be thought of as that which would result from transmission between a pair of planes. In the one dimensional case, the energy is attenuated exponentially by absorption and thus results in the familiar expression of attenuation in decibels per mile. In the two dimensional case, however, in addition to the exponential attenuation factor, due to power absorption there is a decrease in the energy density in the wave as a result of the spreading of the wave over an ever increasing area as it advances. This reduces the energy density inversely with the distance and the field strength therefore varies inversely with the square root for the distance.

The concept of the surface wave was introduced into studies of radio propagation over the surface of the earth in 1907 when Zenneck showed that the interface between earth and air could support a plane surface wave that was exponentially attenuated in the direction of propagation and decreased exponentially with increase in distance from the surface both upwards and downwards. Zenneck did not show that an antenna could generate such a surface wave but because it offered a plausible explanation of radio transmission to great distances it was generally accepted.

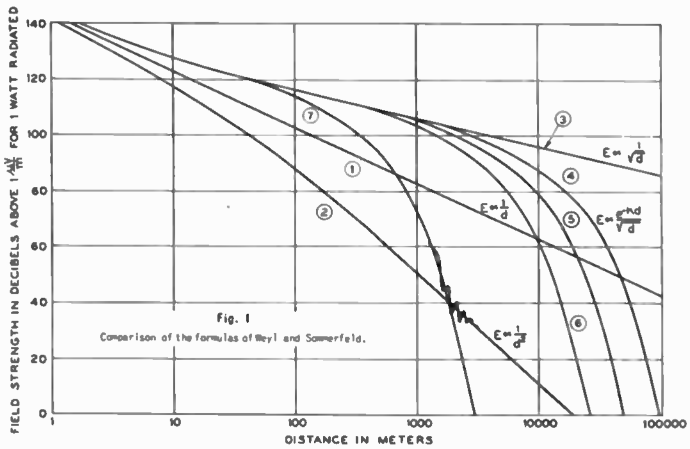

Fig. 1
Comparison of the formulas of Weyl and Sommerfeld.

Two years later Sommerfeld considered the problem of the spreading of electromagnetic waves from a short doublet antenna located in the interface between earth and air. He expressed his result as the sum of two components, one of which he identified as a cylindrical surface wave which at great distances was equal to Zenneck's surface wave. This theoretical work of Sommerfeld seemed to prove that the surface wave was an important component of the radiation from an antenna on the surface of the earth.

Ten years later Weyl reconsidered the problem and obtained an expression for the radiation from an antenna on the surface of the earth which did not explicitly contain the surface wave. From this he concluded that the separation of the field by Sommerfeld into two components, one of which was called the surface wave, was merely mathematical fiction and had no physical counterpart. He was, however, apparently of the opinion that his results agreed numerically with those of Sommerfeld.

Mr. Burrows here pointed out that a careful comparison of the results of the work of Weyl and of Sommerfeld shows that they differ by precisely the surface wave component in question which was the subject of his paper. The comparison of the formulas of Sommerfeld and Weyl are indicated by Figure 1 where the field strength is plotted as a function of distance both scales being on logarithmic. From this figure it was pointed out that for transmission over a perfectly conducting plane the field strength varies inversely with the distance as shown by curve 1. Curve 2 was de-

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scribed as a plot of Weyl's formula for transmission over a dielectric plane. This shows the field strength varying inversely with the square of the distance at the greater distance. Curve 3 is a plot of the surface wave for propagation over a perfect dielectric showing the field strength varying inversely with the square root of the distance at the greater distances. Where the dielectric is not perfect but has appreciable conductivity the surface wave decreases exponentially with distance as indicated by curve 4. Curves 5, 6, and 7 having been plotted for increasing values of conductivity show that as the conductivity is increased the marked influence of the exponential factor sets in at shorter and shorter distances. The Sommerfeld-Rolf curve results from adding the surface wave component to the Weyl curve. The interference of these two components produces oscillations in the curves where the two components are approximately equal. Under the condition where these two components are equal and out of phase at the same distance the theory of Sommerfeld predicts zero field strength at a finite distance as pointed out by Rolf.

Thus, Mr. Burrows stated, resort to experiment was indicated as being desirable in order to decide which of these two curves is correct. In making such an experimental investigation, however it is highly desirable to make transmission tests under conditions where the received field strength predicted by these two formulas differ greatly. This occurs for propagation over a perfect dielectric and since fresh water is the nearest approach to a perfect dielectric available in sufficient volume and area for a test of this kind the locale of making the tests was largely determined by this fact.

The departure of these two formulas from one another increases also with the frequency so that a most revealing test would comprise a determination of the variation of the field strength with distance over fresh water in ultra high frequency transmission. Thus a frequency of two meters was chosen as a convenient and useful frequency for the work reported by Mr. Burrows.

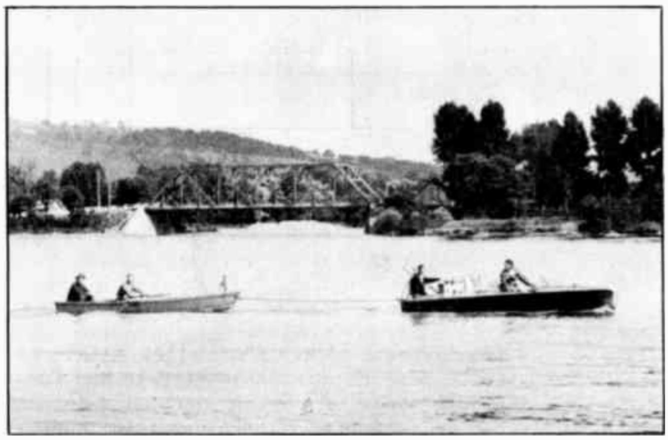

Fig. 2

Experimental arrangement for determining the variation of the received field strength with distance.

During the summer of 1936, Messers Burrows, Decino and Hunt took some two meter measuring equipment to Seneca Lake, New York State and there made these tests. Figure 2 shows a picture of the experimental arrangements. The transmitter was carried in a rowboat towed by a motor boat containing the receiver. The antennas were loaded quarter wave doublets whose midpoints were a quarter wave-length above the water's surface. The experimental procedure was to drive these boats along path 1 of the Figure 3 at a fixed distance apart for a sufficient length of time to make certain that there was no fading such as might be produced by reflections from the bottom of the lake. The distance between the transmitter and receiver was measured by a fishline under a fixed tension.

For distances between receiver and transmitter greater than 150 meters, it was necessary to alter the experimental procedure. For these distances the receiver was located at the end of a pier at Hector Falls and the

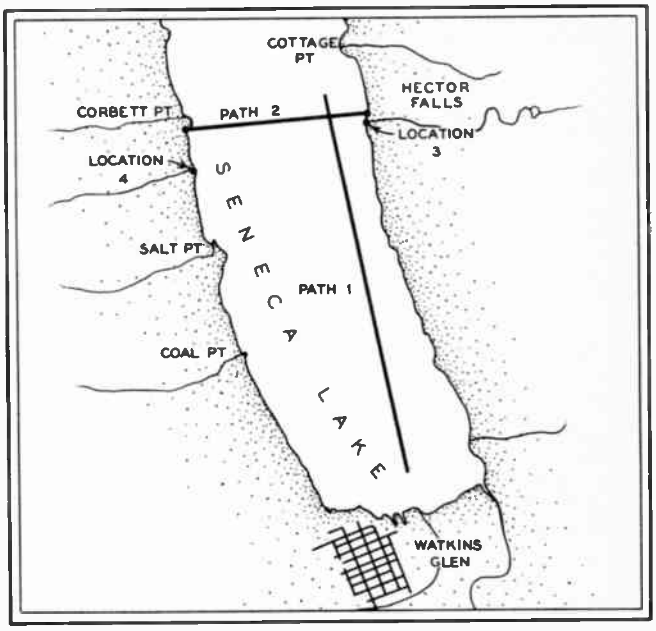

Fig. 3

Map of part of Seneca Lake showing the location of the experiment. Path 1 shows the location of the two-boat experiment. Path 2 the one-boat experiment. Locations 3 and 4 indicate the positions of the terminals for the variable height test.

transmitter was located in the motor boat which was driven along path 2 of Figure 3. This, of course, introduced considerable difficulty in measuring the distance between receiver and transmitter. To minimize the error in this measurement, it was reported by Mr. Burrows that three independent methods were used. First, the motor boat was driven at constant speed in a fixed direction between two points a known separation. Second, the distance between the transmitter and the receiver was measured by means of a transit located at the receiver and a stadia rod carried by the boat. Third, a sextant was used to measure the angle subtended at the boat by two poles located on the shore, one near and the other at the receiver. To complete the measurement the angle between the line joining the two poles and the direction of the boat was measured by the transit.

Figure 4 shows the experimental data so obtained. The solid circles represent data obtained when using the two boats and the open circles those obtained when the receiver was located on the end of the pier. Curve 1 shows the inverse distance variation that would result from transmission over a perfectly conducting plane. Curve 2 is a plot of the Weyl's formula for transmission calculated for water of the characteristics of that of Seneca Lake. Curve 3 is a plot of Sommerfeld's formula for transmission over Seneca Lake water. Curve 4 is a plot of Sommerfeld's formula for transmission over a perfect dielectric.

It will be noted that the experimental data is in good agreement with values calculated by Weyl's formula which as Mr. Burrows reiterated *does not* contain the surface wave component.

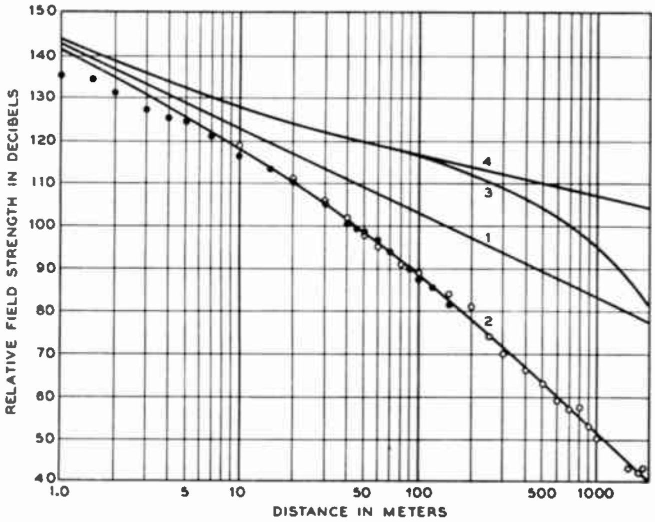

Fig. 4

Comparison of experiment and theory.

The values shown by the theoretical curves depend, of course, upon the distance, wavelength, dielectric constant and conductivity all of which, therefore required evaluation. The methods used in measuring the distance between transmitter and receiver have already been described. The measurement of the wavelength with the required precision introduced no difficulty and since the dielectric constant of water is a known function of its temperature the evaluation of the dielectric constant required, simply, the measuring of the water

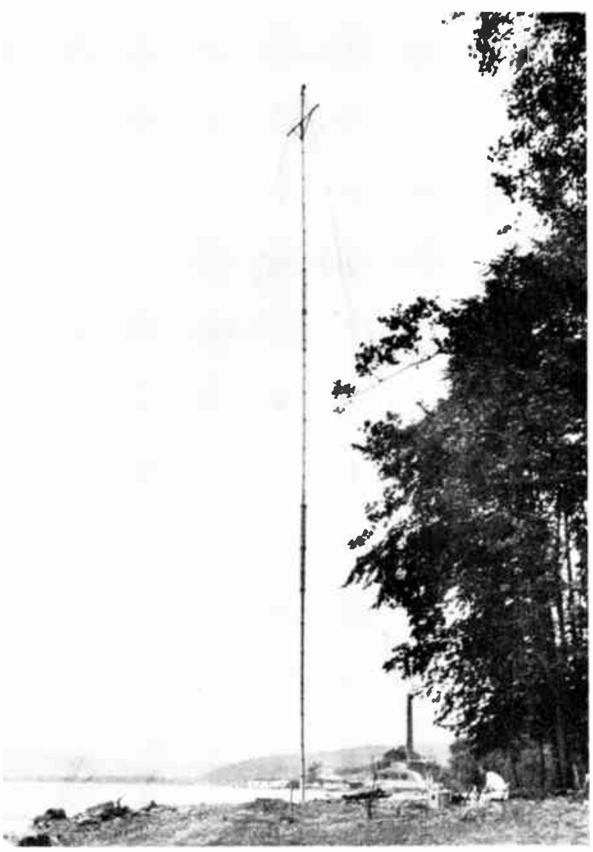

Fig. 5

Picture of transmitting site at "4" of Figure 3, showing the portable 25-meter mast and transmitting antenna.

temperature. The conductivity was determined by laboratory measurements of samples of the water.

Then, leaving the matter of the variation of field strength with distance, Mr. Burrows pointed out that there is another property of a surface wave that might, with interest, be observed experimentally. This is the variation of the field strength with height above the earth's surface. That is, if the field strength is measured over a range of antenna elevations at distances where the surface wave, if any, would be large as compared with the remaining component, there is afforded additional experimental information indicating whether or not the surface wave exists. Accordingly, portable antenna masts were erected at opposite sides of Seneca Lake. The received field strength was determined as a function of the antenna height for two antenna heights at the other terminal as shown in Figure 6. The solid

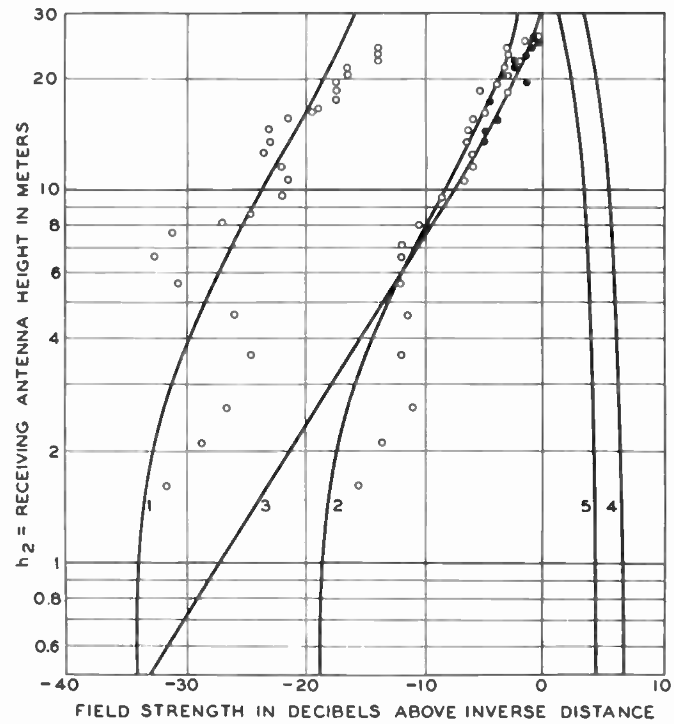

Fig. 6

Variation of received field strength with antenna height.

circles represent data taken on horizontal polarization with one antenna at 24.8 meters above the surface of the water. Since there is no uncertainty in the formula for the field strength with horizontal antennas these points may be used as a calibration of the equipment. Curve 3 is a plot of the formula for the received field strength with horizontal antennas. When this curve is made to fit the experimental data the locations of curves 1, 2, 4 and 5 are fixed. Curves 1 and 2 show the variation of received field strength with vertical antennas that would result if there were no surface wave for the two transmitting antenna heights. Curves 4 and 5 are plots of the surface wave for these two conditions. It will be noted that there is no semblance of agreement between that data and curves 4 and 5. The experimental data does, however, agree with that of curves 1 and 2. These latter are, of course valid only if there is *no surface wave component* either in absolute magnitude or in the variation of magnitude with antenna heights.

Mr. Burrows called attention to the fact that the oscillations in the experimental data are presumably due to reflections from the hills and cliffs which line the

Lake and from trees near the receiving antenna. He pointed further that this experimental evidence proves conclusively that simple antennas *do not generate a surface wave* and therefore the Sommerfeld-Rolf formulae and curves require revision for all conditions where the dielectric constant cannot be neglected.

To further indicate the departure between the two sets of concepts discussed by him, Mr. Burrows showed figure 7 which compares the Sommerfeld-Rolf curves with

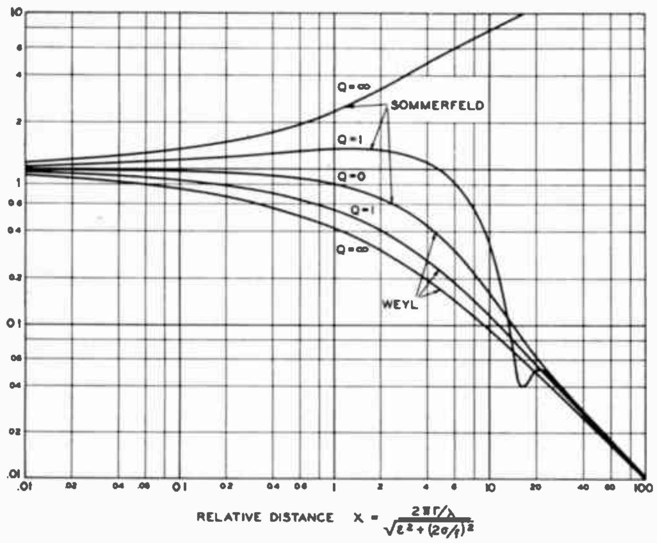

Fig. 7

Comparison of Sommerfeld-Rolf curves with the Weyl-Norton curves.

the new Weyl-Norton curves the validity of which were established by the experimental work which he described. The attenuation factor is plotted against the relative distance. In this the attenuation factor is defined as the factor by which the field strength that would result from transmission over a perfectly conducting plane must be multiplied to give the field strength under the conditions of interest.

For transmission over an imperfect conductor in which the conduction current is large compared with the dielectric amount the two formulas agree as indicated by the curve marked Q = 0.

When the conduction current is equal to the dielectric current (Q = 1) the Sommerfeld formula indicates an attenuation factor greater than unity up to a certain distance, while according to the Weyl formula the attenuation factor is somewhat less than that for the conductivity case. For transmission over a perfect dielectric (Q = oo) the Sommerfeld formula indicates that the attenuation factor is always greater than unity and increases indefinitely with increase in distance, while the Weyl formula indicates the attenuation factor is only slightly less (up to about 10 db.) than that for the conductivity case.

In conclusion Mr. Burrows stated that the validity of the Weyl formula has been unquestionably established by the work reported in his paper as against the previously generally accepted Sommerfeld formula.

The discussion that followed the reading of the paper brought to light much that was of interest to the membership. It was the first reaction of those in attendance that it was to be assumed from Mr. Burrow's conclusions that the concept of the radio wave as being a wave moving forward with its "feet in the ground" would have to be abandoned and it was quite evident from the comments that the attendants were by no means completely willing to abandon this old and generally useful concept unless something more useful could be supplied in its stead.

No alternate concept was offered and, indeed, it was not insisted that this simple concept need be abandoned even in view of the apparent need for abandoning the conclusions usually drawn from the Sommerfeld formulae. Instead it was repeated that the important aspect of the conclusions from the work reported by Mr. Burrows concerns itself with the magnitudes of the field strengths to be expected under practical operating conditions as compared with those to be expected under the Sommerfeld formula.

Of major importance in this is the fact that this abandonment of the Sommerfeld formula brings with it the realization that the field strength of any transmitting station is less effected by the Q of the ground than had heretofore been supposed and that, in fact, there is no great advantage in transmission over a perfect dielectric as has long been assumed to be fact.

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EXPERIMENTS IN GENERATION, DETECTION AND MEASUREMENT AT ONE METER WAVELENGTHS

BY

PAUL ZOTTU*

Delivered before the Radio Club of America
March 11, 1937

Mr. Zottu introduced the subject of his paper by pointing out that the problem of developing oscillators, amplifiers and the like for use at extremely high frequencies - below a wavelength of one meter - is commonly thought of as being largely one involving the development of circuit elements which will provide impedances of sufficiently high value. Thus, the first step in the development of ultra-uhf equipment is the development of high impedance coupling circuits. He disposed of the possible usefulness of conventional types of tuned circuits by pointing out that while they held considerable promise, at the moment the use of lines of various types recommends itself as being the obviously and immediately useful type of arrangement for oscillator stabilization, etc. He limited his further discussion to the closed quarter wave line as being the obviously useful type for providing high impedances and high Q for tube couplings. In this he defined the Q of a quarter wave resonator as F. E. Terman defines it in his article of July 1934 in Electrical Engineering, that is, by analogy with the Q of a lumped circuit. This gives an expression for the Q of such a line as

$$Q = \frac{2\pi Z_0 f}{RC}$$

where R is the resistance per unit length of the line, C is the velocity of light, Z_0 is the characteristic impedance of the line and f the frequency at which it is excited.

Mr. Zottu then proceeded to a discussion of the theoretical factors operating in the choice of dimensions of both open wire and concentric lines for maximum or optimum end impedance and Q. These provide certain optimum dimensional proportions, the optimum relationship between conductor spacing, diameter and wavelength in the case of open lines and the ratio of conductor diameters and wave length in the concentric lines being only approximately equal for maximum end impedance and for maximum Q. In this it was suggested that some difference of opinion exists between the speaker as the result of his experimental work and other workers in this field as the result of their analytical work, as to the influence of radiation resistance on these optimum relationships in the case of concentric lines. There was much discussion of this point after the paper, which might well be mentioned here: the point being that while it was obvious that in the case of open wire lines radiation must be taken into consideration since, as the lines are spaced further and further apart in order to get higher surge impedance and lower attenuation, the radiation resistance increases as a result of the increasing amount of power lost through radiation as the spacing is increased, so that one comes ultimately to a spacing between conductors beyond which the increase in power loss through radiation more than offsets the increase of surge impedance and reduction of assymetrical current distribution in the conductor, i.e. skin effect.

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Figure 1 indicates the magnitude of both the purely resistive end-impedance and the Q of open wire lines of optimum dimensions. From this it appeared that Q's of about one thousand and end impedances of several hundreds of thousands are readily possible at 100 M. C. when using open wire lines.

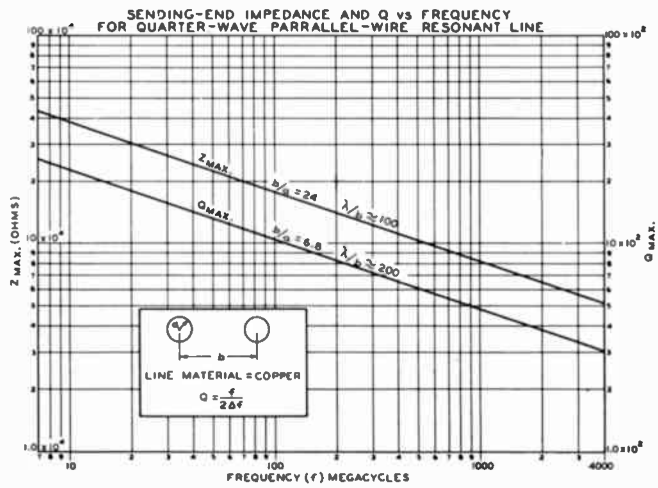

Fig. 1

Similar reasoning, including the consideration of radiation loss, if any, provided the basis for the determination of optimum proportions for concentric lines. Figure 2 shows the maximum end impedance and maximum Q over a wide range of frequencies - 1. to 400. M. C. - for optimum proportions. This indicates that end-impedances and Q's are about ten times as great as those of open lines. On the matter of radiation loss within a closed

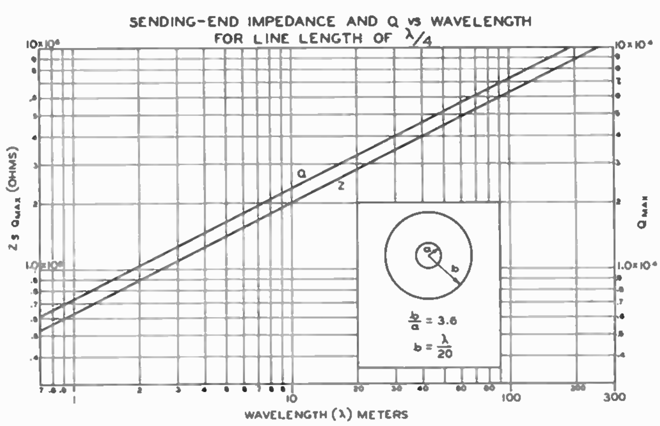

Fig. 2

concentric line, the later discussion, as invited by Mr. Zottu's comment, waxed hot and heavy: the point being that Mr. Zottu had limited his analytical work to such dimensions of conductors that the radiation factor proposed by certain other unnamed workers in arriving at the optimum relationship was negligible notwithstanding the fact that he was unconvinced of the rationality of the assumptions basic to this point. The point at issue here appearing to reside in the fact that as the ratio of the diameters of the conductors of a concentric line gets great and as the difference between them gets to be an appreciable portion of the wavelength, the assumption of the instantaneity of the building up of the radial flux within the line introduces an increasing error and one which would doubtless limit the ratio of diameters to some finite value if valid correction is made for it. Mr. Zottu felt that such correction as proposed does not however meet the needs of the problem.

It might here be pointed out in passing that from the purely analytical viewpoint the lack of instantaneity of development of flux within a concentric line results in a departure from the precise in-phase or quadrature relations indicated by the analysis of the properties of loss-free lines and thus probably contributes to the characteristics of the line under consideration precisely as does the presence of losses in the line. Whether such a loss-like relationship in the properties of the lines discussed by Mr. Zottu actually means loss by radiation or otherwise appeared of only academic importance. The influence of this factor on the end-impedance and on the Q of the line is, however, of major practical importance from the view point of Mr. Zottu's paper but, unfortunately, was not completely included in his presentation of the problem.

It was brought out in a later section of Mr. Zottu's paper that in the case of an oscillator including a concentric line closed at both ends and including within itself all circuit elements including the tube, he was able to find no evidence of radiation or other fields external to the line.

Mr. Zottu next proceeded to a discussion of the application of the relations shown by his graphs. He indicated that a quarter wave line at any but the highest frequencies requires far more space than is likely to be available under practical conditions. Thus, as he pointed out, in the broadcast band for nearly optimum dimensions something like a pair of smoke stacks, something over three hundred feet high, would be required for the stabilizing of a broadcast transmitter by means of an "open wire line" while a pipe, sixty feet in diameter and three hundred feet long, would be required to serve as the outer conductor of a suitable concentric line. This pointed the obvious limitations of simple quarter wave lines and introduced the compromises that he has found he could make and still get the desired high impedances and high Q. In general, this was accomplished by two different means. The first and most useful appears to have been the use of a line of approximately optimum cross sectional dimensions but of only a small fraction of a quarter wave length terminating in a capacity. The capacity required for this purpose is secured by putting a cap over the end of the line and terminating the inner conductor in a large flange and making the cap movable relative to the flange for purposes of "tuning". It was not pointed out by Mr. Zottu, but it is in fact a moot question whether such a structure consisting of an extremely short line and terminating in a capacity is really a "line" or whether it is merely an extremely low loss and conveniently constructible inductor and a capacitor.

At any rate, the short line with capacity termination was one form that Mr. Zottu's circuit arrangements took in order to secure high impedance for the tube couplings.

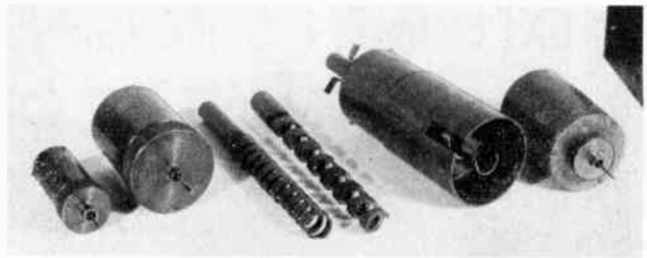

Fig. 3

The second form which was shown, Figure 3, is a concentric line in which the inner conductor consisted, not of a simple cylindrical member, but of a spiralled conductor of rectangular section; this, in the interest of raising the surge impedance, decreasing the axial velocity of propagation and hence, making a quarter wave line short enough to be useful. Because of its greater complexity this type of line was not much used.

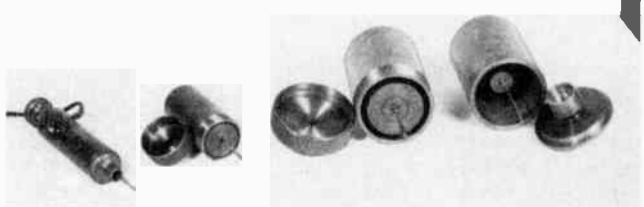

Fig. 4

Samples of a number of different sizes and types of these several forms of shortened lines as shown in Figure 4 were available for inspection after delivery of the paper.

In these forms of circuit construction provision has to be made for the coupling to the tubes or associated circuits. Such coupling was necessarily made variable so that "matching" would be readily possible. In general, this was accomplished by providing slots in the outer conductor near the low potential end through which conductors terminating on the inner conductor could pass for connection to the tube or other circuits. Where used with tubes this provided either for "back-coupling" of grid and plate circuits in the case of oscillators or plate-circuit-to-grid circuit coupling in the case of amplifier arrangements. In these practically useful forms of short lines provision for tuning was in each case made by an adjustment knob at the low potential end either for shorting portions of the spiralled inner conductor or varying the capacity at the remote end of the line, or for varying the effective length of the line itself as is shown in the figures.

After an extended discussion of these details, Mr. Zottu showed an oscillator consisting of eight tubes arranged radially on a circular metal plate, each tube carrying its own plate and grid circuits each consisting of a single loop of wire and coupled into a rather unusual form of short concentric line. This line consisted of an inner conductor of what was termed a concentric line, lacking however, the outer concentric conductor: the latter being replaced by three cylindrical standards of small diameter supporting the end plates of the line, one of the end plates comprising one of the plates of the line terminating condenser. The connections between the tube circuits and the line which acted as a single tank circuit for all eight tube circuits being provided by taps on the inner conductor connected through coupling condensers - either adjustable or fixed - to the single loop plate or grid circuits of each tube.

Fig. 5A

METHOD OF MOUNTING AND BY-PASSING ACORN TUBES

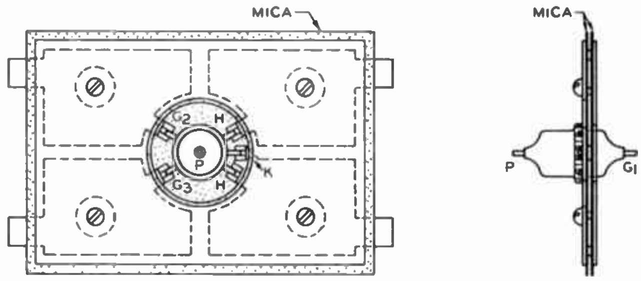

Fig. 5B

This was reported to give a power output of the order of 80 watts at about one meter.

Mr. Zottu explained at some length how this radially symmetrical structure might be extended by the use of a line tank circuit not so greatly foreshortened so as to provide for its being fed by a number of levels of radial tube generators and suggested further extensions of this scheme of symmetrical layout in the interest of making it possible to use tubes as oscillators at frequencies near the upper limit imposed on their operation by their individual impedances and thus, avoiding the paralleling of their capacities and impedance and the consequent limitation on the upper frequency of their operation. It was pointed out that one could, by the arrangements shown, operate an almost unlimited number of tubes all feeding into a single tank and thence to a single dissipator at the same high frequency as that at which the tubes would operate individually.

Mr. Zottu then reviewed the acorn tube oscillators, shown in the B. J. Thompson papers of sometime ago. In this he brought out a distinction between the original Thompson tubes and the now commercially available acorn tubes - the 955, etc. - by referring to the original and still experimental tubes, as "shoe button" tubes.

This portion of the paper stressed some of the details of the design of circuit elements such as the inclusion within the tube mounting of the required bypassing capacities by the simple expedient of mounting the socket elements on separate metal plates which were separated from the datum metal plate base of the assembly by thin dielectric in the interest of getting high bypassing capacity. Additionally, various combination current and voltage supply leads were shown in which thin sheets of dielectric were used between the strip or thin plate leads in the interest of getting by-passing.

In addition to bypasses, Mr. Zottu showed an interesting "universal" choke coil the operating range of which was from about two meters down to one half a meter. It consisted of a single layer solenoidal coil of small diameter - about half an inch in diameter - and about four to six inches long wound with a single layer of wire, of either copper or higher resistance material. The theory of its usefulness being that a coil of a sufficient number of wavelengths long and of sufficient attenuation will provide so little reflection from the receiving end back to the transmitting end that, as viewed from the transmitting end, it would offer only its own surge impedance at all frequencies. In addition, other forms of lines of use as chokes were shown, amongst them simple adjustable lines comprised of a small channel section as the outer conductor and having a concentric wire as the inner conductor and a slider to provide H. F. short-circuiting of the line to the channel section and thus to provide the adjustability necessary for making the end impedance of the line high and largely resistive. These as shown in Figures 5A and 5B and other similar devices were found necessary in oscillator and amplifier designs to maintain filaments and other elements requiring D.C. excitation at potentials determined by the H. F. requirements of the circuit and independent of the potential of the current and voltage sources.

Several other forms of oscillator were shown: amongst them one, shown in Figures 6A and 6B in which an acorn tube was mounted within the concentric line and provision made for changing the coupling of the line to the tube through the shifting of the line as a whole relative to the tube and for tuning the line by means of a break in the inner conductor which was closed through a dielectric adjustable as to capacity from outside the closed line.

Mr. Zottu then proceeded to a discussion of such measuring methods and devices as had been devised for work in this field. It was pointed out that with the means for measuring voltage and power available, most desired characteristic data could be secured. There were, therefore, two general types of measuring instrumentalities devised. The first to be described was the thermocouple type of watt meter and the second was the diode rectifier type of peak voltmeter. Of these the first took the form of a number of different types of vacuum

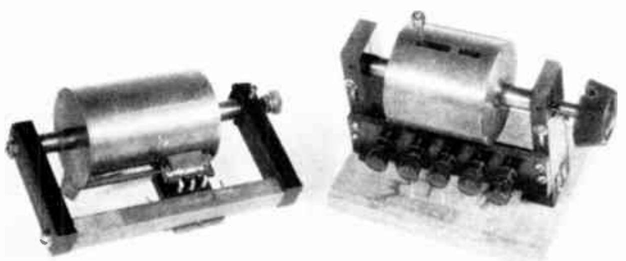

Fig. 6A

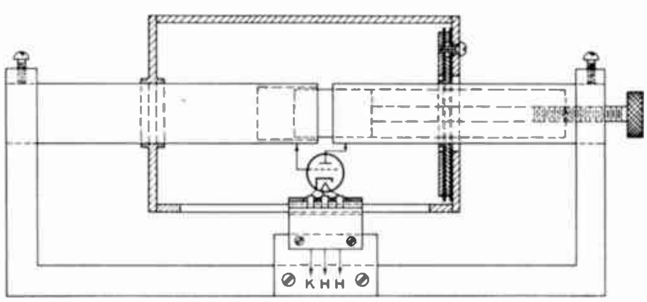

Fig. 6B

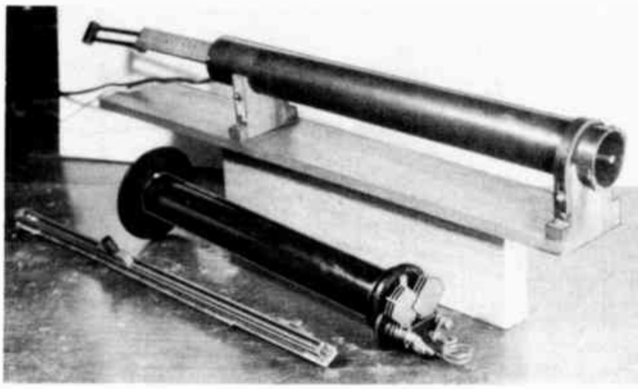

Fig. 7A

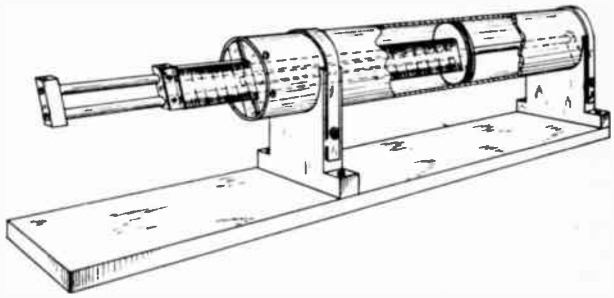

Fig. 7B

type thermocouples in which all the available power was dissipated in a heater and the resulting thermoelectric voltage read externally on a conventional indicating instrument, the arrangements of the heater and couple elements and terminal arrangements being such as to minimize the coupling between the heater and the couple circuits. In order to provide for rather unusually high heater temperatures the heater itself was made of extremely fine tungsten wire, while the nature of the couple conductors was undescribed. Calibration curves of several typical thermocouples were shown indicating that for the most sensitive thermocouple watt meter the system gave full scale deflection at about 40 milliwatts while for the least sensitive couple maximum power indication was gotten at about 40 watts.

Photographs of a number of diodes devised for use as diode peak voltmeters were shown. These were all characterized by extremely small anodes and filaments with microscopic clearances between them. Thus, the smallest of the diodes shown included a cylindrical anode six mils in outer diameter enclosing a filament of unspecified diameter. The circuits used with these diodes are quite conventional in that the diode output circuit comprises a condenser which is charged by the unidirectional electron current to such a potential as to reduce the electron current to a negligible amount. It was pointed out that the minute clearances were made essential by the influence of the electron transit time upon utility of the diode for this purpose. Even the smallest of the diodes shown required correction for frequency at the extremely high frequencies at which they were used. Thus, a calibration curve for this diode was shown in which the 60 cycle calibration departed little from the calibration at one meter.

A brief discussion of the relation between electron transit time - or the phase shift of the diode current and voltage due to transit time - and the current flowing in the diode served to indicate the procedure followed in making the frequency corrections to the diode voltmeter characteristics.

A third type of measuring instrument was shown in the slides: this, of the wave meter type. Two specific types were indicated and are here shown in Figures 7A and 7B. One of them consisted of a few turns of wire connected to a variable condenser of the Remler type presumably in order to secure complete symmetry to ground in the tuned circuit and to maintain a low minimum capacity, the LC combination was mounted at the end of a long dielectric rod or tube with the control and scale mounted at the end remote from the inductive elements. No resonance indicator was included in this type of wavemeter; thus, it was useful only for "absorption" measurements.

The second type consisted of a concentric line in which the effective length of the conductors was made variable through the provision of a sliding member within the line contacting both the inner and the outer conductors with a moving scale carried along with the sliding member and itself sliding under an indicator external to one end of the line, thus indicating wave length directly on the linear scale; the entire assembly being provided with a removable end so that it might act as an open or shorted line, thus providing for two wave length ranges.

Mr. Zottu then proceeded to a discussion of the adaptation of these types of structure to selective receiver circuits in conjunction with acorn tubes. He showed illustrations of several types of receivers already shown on previous occasions, notably in connection with the Thompson IRE papers in which conventional tuning arrangements of small size were used and pointed out that the gains per stage gotten by these means were small even at the lowest of the high frequencies at which the receiver operated; and, indeed, were never in excess of four per stage. This low gain appears to have provided the impetus required for the attempts to adapt the "line" type of tuned circuit to receiver uses and resulted in a three line receiver which was shown both in slides and "in the flesh" after the reading of the paper and is here shown in Figure 8. The problem to be

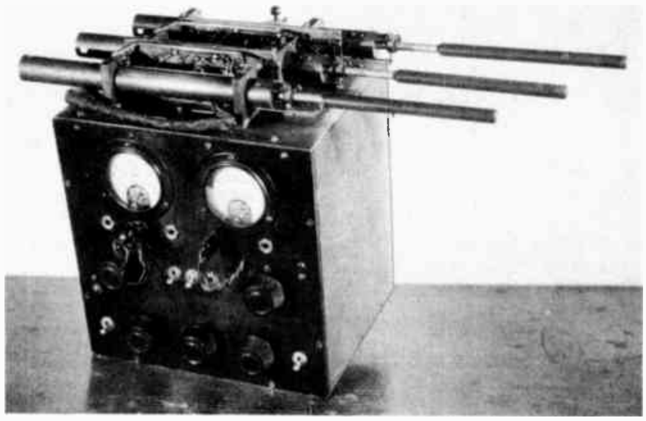

Fig. 8

faced in the design of the receiver was reported to have been one of providing such a wide range of adjustability as to accommodate a wide and apriori unknown range of tube impedances and such a wide frequency range as would determine the ultimate limits of the receiver's effectiveness. Thus, not only were the lines made adjustable as to length for a wide range of frequency variation but adjustable with respect to the point of connection along their lengths to the tube elements. Each of the three assemblies consisted of a yoke carrying a sliding line the tube being mounted on the former while a sliding member within each line provided for the adjustment of the effective length of each of the three lines to suit the desired frequency of operation. Thus, the tube in each one of the three assemblies comprising an amplifier stage could be connected to its own line on the plate side at the proper point along the length of the line by the sliding of the line within its yoke and to the adjacent line at the proper point along the length of that line by the sliding of the adjacent assembly, including the yoke and all and, in addition, the frequency of operation of each of the three stages was adjusted by sliding the short circuiting member within the line housing.

Mr. Zottu pointed out that, as was to be expected, when coupling the tube plate circuit to such a high impedance as was provided by the line, it was found necessary to make the plate-to-line connection relatively close to the low potential end of the line in order to secure low enough coupling to give the desired selectivity of operation but, as was not so definitely expected, it was found necessary to make the grid-to-line connection also near the low potential end and, even less expectedly, found necessary to make this connection even nearer the low potential end than the plate connection if anything approximating a desirable degree of selectivity was to be obtained. Thus the tuned R. F. receiver comprising the tubes and lines consisted of three sharply tuned circuits coupled loosely out of the tube plate circuits and even more loosely coupled into the tube grid circuits.

It was pointed out that the unhappily low input impedance was not due, as is sometimes suggested, to the capacitive reactance of the tube input but in large measure to the low value of the input resistance - or high conductance - which, in turn is due to the relatively high ratio of the electron transit time to the period of the circuits. In fact, it was shown that at a point in the range of the receiver *not* close to the upper limit of its frequency range, the input resistance of the tubes became so low that the gain of the receiver was reduced to unity.

Mr. Zottu's conclusion to his most interesting and instructive paper indicated that the solution to the problems that so seriously limit the development of suitable radio receivers, oscillators and other radio devices employing vacuum tubes for use at extremely high frequencies has been shown by the work reported in this paper *not* to reside in the design or construction of the circuit elements but rather in the design and production of vacuum tubes. Indeed, one understood from the paper, that further progress in the field reported on by Mr. Zottu was unquestionably and completely dependent on the development of new and suitable types of vacuum tubes.

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THE APPLICATION OF THE BROAD BAND CRYSTAL FILTER TO BROADCAST RECEIVERS

BY

ALEXIS GUERBILSKY*

Delivered before the Radio Club of America
December 9, 1937

The piezo-electric crystals are known to have a very sharp resonance and are used either to control one single frequency with a sharp definition as, for instance, in radio transmitters or to resonate for a very narrow band as used for telegraphic, so called single frequency receptions, where they function as filters. There are however exceptions, for instance, in ultra-sonic sea signaling where the crystal is dampted by the radiated energy. In such applications the selectivity of the crystal is of no importance and the energy producing the damping is precisely the energy which is used. But, if we try to damp the crystal artificially in most applications its resonance curve will be less sharp, but it will still provide a characteristic not unlike that of a simple tuned electrical tuned circuit. For some other applications, as for instance, for optical relays, such artificially dampted crystals require too much energy for their operation and thus lose most of their valuable qualities.

It has long been obvious that crystals which would resonate over a broad band of frequencies would be extremely useful and, indeed, attempts have been made to obtain such crystals. The first thought was to artificially damp the crystal, but as explained above this method is in many cases unsuitable because it destroys the very characteristics of the crystal which make it especially useful.

The second thought was to use several crystals connected in parallel, the frequency of the several crystals differing from one another so as to cover the desired band. The difficulty in this expedient is that between the two frequencies corresponding to two successive crystals there is always one at which the two crystals vibrate in opposite phase, one crystal being operated below resonance and the other one above. This results in the fact that the electrical resonance curve has a sharp minimum-approaching zero - which makes the arrangement generally unsatisfactory.

Let us now discuss a solution of the problem making it possible to obtain a single crystal with a practically uniform response for any desired band of frequencies, while, at the same time providing a response abruptly decreasing at the limits of the band. This solution is, primarily, the use of a crystal of non-uniform thickness.

CRYSTAL OF NON UNIFORM THICKNESS. EARLY YEARS OF TELEVISION.

When about 1926 the first steps had been taken in practical television, one of the greatest difficulties had been that of synchronization, this difficulty increasing with the number of the picture elements per image, i.e., definition. The idea of operating the scanning at the receiving station by the scanning system at the transmitting station seems to have been first suggested by Prof. Rosing in the early years of the century. This method makes the synchronization independent from the speed of scanning. To make this method practical it is possible to perform the television by using two parameters of a radio wave, its amplitude and its frequency; the amplitude corresponding to the luminosity of the transmitted picture element and the frequency to its sequential position.

Let us, in order to make it simpler, consider the transmission of only one line of the image supposing that the second scanning, at relatively low speed, can be made by other already known methods.

At the transmitting station a rotating optical scanning device is mechanically coupled to a rotatable, variable condenser which determines the frequency of the transmitted wave.

To every position of the scanning device there corresponds a frequency determined by the variable condenser and while the amplitude as determined conventionally by a photo-electric cell, corresponds to the luminosity of the transmitted picture element. At the receiving station we have then only to make the position of light spot correspond to the frequency of the received signal and its luminosity to the signal amplitude. The first idea was to use for this purpose a series of crystals of different frequencies corresponding to the received band width, each of these crystals constituting an electro-optical relay passing light when it is operated to the corresponding position. Its amplitude which is a function of the amplitude of the received signal determines the luminosity of the projected light. For this purpose the crystals have been disposed in a system of polarized light, their vibrations modifying the polarization and thus modulating the intensity of the light. Later this series of crystals were replaced by single crystal of a thickness varying from place to place to cover continuously the entire band of frequencies, every frequency determining a localized resonance in the portion of the corresponding thickness. As an example, a crystal of 10 cm. long and resonating over a band around 80 meters gave a localization of resonance of about 1 mm. The width of the band was about 300 kc.

SOUND RECORDING SYSTEM.

The next use of the crystals has been as a light valve for photographic sound recordings.

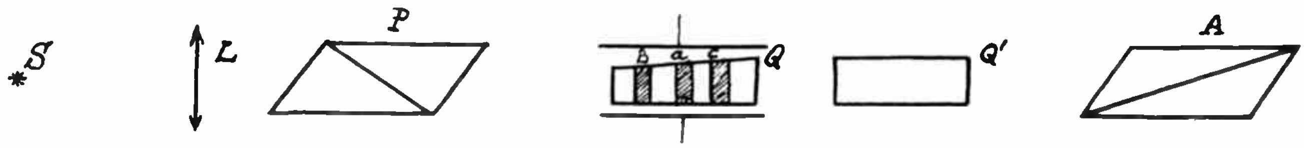

Figure 1

Figure 1 illustrates this application. The crystal, Q, is traversed by a beam of polarized light produced by source, S, a lens, L, and the polarizer, P. On its way the light traverses the portions of crystal of the different thicknesses. In the absence of any vibration of the crystal, the system is compensated by the compensator, Q', and the analyzer, A, at extinction. The crystal is excited by an oscillator of a frequency, f, corresponding to its middle portion, a. The generator is modulated by the microphonic current.

If the microphonic current is of a frequency F, there will be three portions of the crystal a, b, c, simultaneously set into vibrations corresponding to the frequencies f, f+F, f-F. The action on the polarization by the deformations of these three portions will be recomposed with conservation of phase and the luminosity of light correctly modulated for the whole band determined by the thicknesses of the crystal.

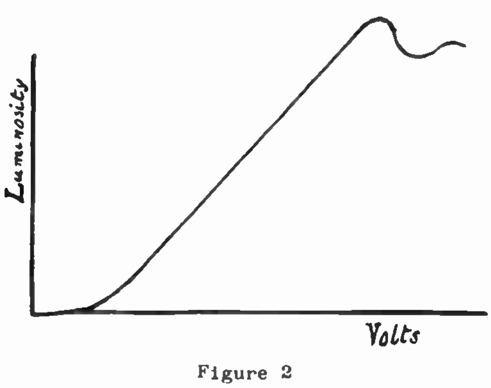

Figure 2

The curve giving the luminosity as function of the voltage across the crystal is shown in fig. 2. This curve has an important linear portion and the light can be modulated practically without distortion up to 90%. The luminosity is very good. In one set of equipment, when the Eastman sound negative is used, with a gamma of 0.5, the density is 0.5 for an A.C. voltage across the crystal of about 100 volts.

SOME OTHER APPLICATIONS OF THE NEW CRYSTALS

In all its applications these crystals are intended to vibrate either simultaneously or successively for a band of frequencies, as in sound recording and television.

Amongst other applications and before discussing the band-pass filters, which is the main object of the present paper, we will only mention two. First, the control of a transmitter, the frequency of which it may be desired to vary to avoid possible interferencies; and second the use as microphone.

To control a transmitter a wedge shaped crystal can be used, one of the electrodes being a narrow plate sliding along its surface, the portion of the crystal positioned between the electrodes determining the frequency.

In its use as a microphone, the crystal vibrates at a high frequency and its impedence which depends largely on the damping determines the useful microphonic current.

As a matter of fact the damping varies according to the density of the air around the crystal which density is modulated by the sound.

The same idea was used by Prof. Riabouchinsky and myself to make a dynamo-meter for aerodynamic measurements and recordings, the wedge shaped crystals making it possible to record rapid variations of air density or pressure.

The two last methods are schematically shown in fig. 3.

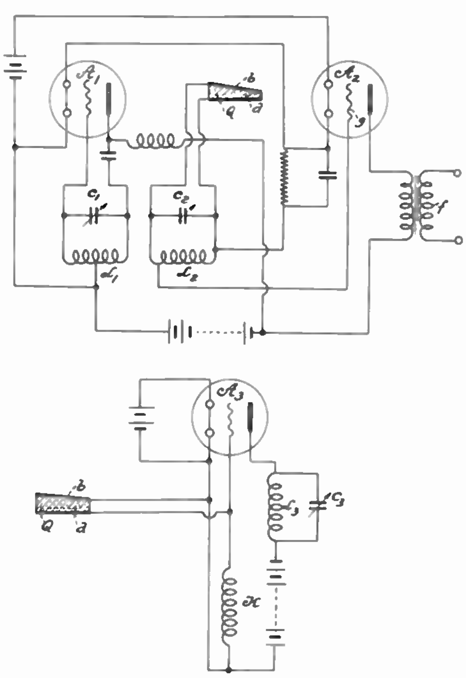

Figure 3

Figure 4

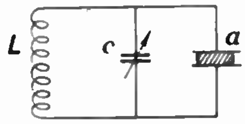

Figure 5

VIBRATION OF THE CRYSTAL.

In all the above discussion it has been assumed that the localized vibrations in the crystal travel continuously from one end to the other when the frequency varies. It is not however always so. If we record the resonance curve, this curve presents many irregularities. One of such curves can be seen on the fig. 4 which is an oscillogram of the voltage as function of frequency, of a circuit such as is shown in fig. 5, the "crevasse" of fig. 4 showing the resonance curve of the crystal.

To explain this curve and show how to correct the irregularities let us examine what occurs in the crystal when it is submitted to the action of an electric field of frequency varying between corresponding limits.

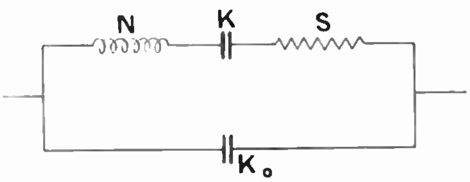

Figure 6

Let us consider first a usual crystal of uniform thickness. This crystal can be represented by the equivalent cell shown on the fig. 6, the inductance N, resistance S and capacities K and K_0 depending on the dimensions of the crystal. The corresponding resonance curve is shown in fig. 7. The sharp rise a corresponds to the resonance of the circuit NKS and the point b to the parallel resonance of the complete cell.

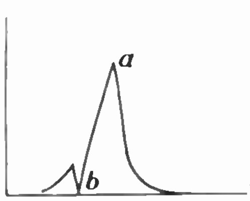

Figure 7

If now the crystal is of non-uniform thickness, it is represented by an infinity of small crystals of different thicknesses coupled together and the equivalent cell will be an infinity of elemental cells connected in parallel and coupled in such a way as to prevent any important shift of phase between any two adjacent cells. This condition corresponds to the mechanical realities of the crystal structure.

The resonance curve will be broad and will include all the frequencies of the band of the crystal, but to determine its shape we must not neglect all the phenomena which may occur.

First of all when the crystal is subjected to a frequency f a portion a, Fig. 8, is set in resonance vibration.

Figure 8

The corresponding elongation will also produce elongations in perpendicular direction the importance of which depends on the value of Poisson's modulus for the particular orientation of the crystal with respect to its axis. These latter elongations produce a wave propagated along the crystal in all directions. The edges of the crystal, such as the edge B, reflects a wave returning in a with a phase depending from the distance to B. When f varies the phase varies and the electrical effect will have as many maxima and minima as the varying phase. To avoid the reflections on the edges it is practically sufficient to dispose on these edges some absorbent material such as Canada balsam or any similarly viscous material. This will produce only useful damping, without markedly affecting the localized vibrations of the portion a.

Figure 9

Figure 10

The oscillogram of voltage as a function of frequency fig. 9 shows the characteristic of the crystal after the edge reflections have been attenuated as suggested above. There are still many maxima and minima due now to the coupling of the main mode of vibration with other resonators such as the harmonic resonances along the length, of flexion, torsion etc. These resonances are well understood and can be eliminated by localized damping. The curve resulting from such treatment as this is shown in fig. 10.

The curve of fig. 4 is of an X cut crystal as has been shown to illustrate better the different phenomenae. In the modern cut crystals, such as AT, AC, etc., the couplings and consequently the irregularities in the characteristic are greatly reduced and the corrections to be made are much easier and, when less uniformity is required, may be completely unnecessary.

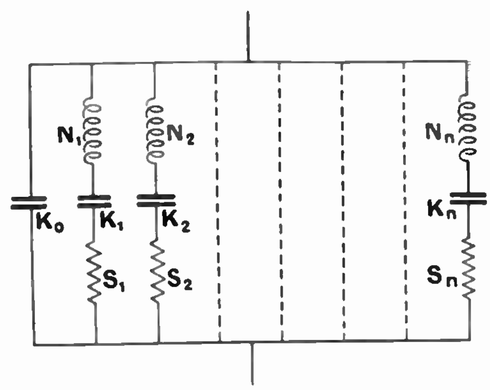

Figure 11

Now, assuming that all the spurious oscillations have been eliminated the crystal is fully represented by the equivalent cell of the fig. 11, (where the above discussed couplings between adjacent cells have been however omitted). The construction of the resonance curve is similar to the construction of the curve on the fig. 4. When the frequency reaches the value corresponding to the resonance of the first elemental cell, the current increases to the point A on the fig. 12 and then remains constant because of the identity of different elemental cells, until the antiresonance of the first cell is reached. At this point the current decreases to the point B and remains constant until at C the resonance of the last cell ceases. Then the anti-resonance of the

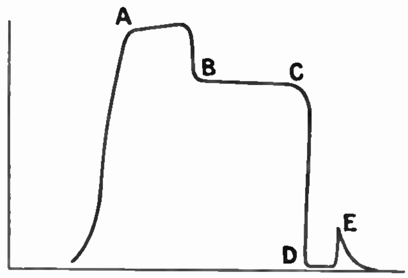

Figure 12

last cells produces a decrease of the current to the point D which corresponds to the point b on the fig. 4. The last anti-resonance ceases at the point E.

If we compensate the parallel capacity of the crystal by a bridge circuit, the current in the absence of resonance will be zero and any variation of the impedance will produce an increase of the current. The curve on fig. 12 will obviously become the curve on the fig. 13.

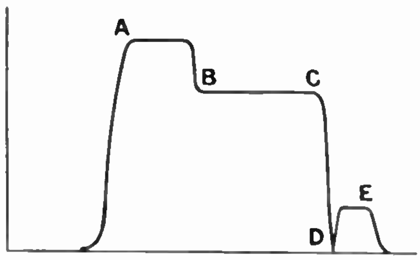

Figure 13

We see that this curve has three steps, but the importance of the height and width of the steps depends on the ratio of the portion of the crystal simultaneously vibrating to the whole of the crystal, or the number of elemental cells simultaneously resonating to the ensemble of cells. It means that by differently choosing the dimensions of the crystal, its length, thickness, slope, very different curves can be obtained which can be easily compared to the curve on the fig. 13. Some of such examples are shown by the oscillograms on the fig. 14, 15, 16.

Figure 14

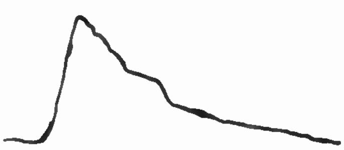

Figure 15

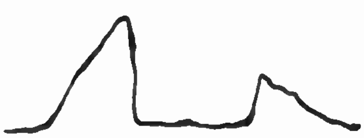

Figure 16

Furthermore, if the slope is too small the vibrating portion will be an important part of the crystal, as seen on the fig. 17 and for a small variation of frequency this portion will decrease and the corresponding curve will have no flat top and will resemble the usual curve of an electric tuned circuit. While if the slope of the crystal is too large the resulting damping may become seriously important.

Figure 17

When a crystal must be made for a predetermined frequency with a predetermined band it must be determined what is the value of the localization of the resonance, either theoretically, by considering the elastic constants of the crystal, usually quarz, or experimentally by analyzing the vibrating crystal in polarized light and by gradually changing the angle between faces.

I. F. TRANSFORMERS

The above described features of wedge shaped crystals make it possible to use the crystals to improve the response of pass-band filters and especially IF transformers. As a matter of fact, these crystals can be made for frequency bands as desired, the sharpness of the skirts of the resonance curve, being practically independent from the band width. Furthermore, the transformers using those crystals will have the selectivity practically independent from their frequency which can be, for example, as effective for 470 KC as for 1,500 KC or higher.

The circuit used for IF transformers is very much like the crystal transformers used in amateur so called single signal receivers, but I have tried to make them as simple as possible, because the price is in a radio receiver of a great importance. The circuit is represented on the

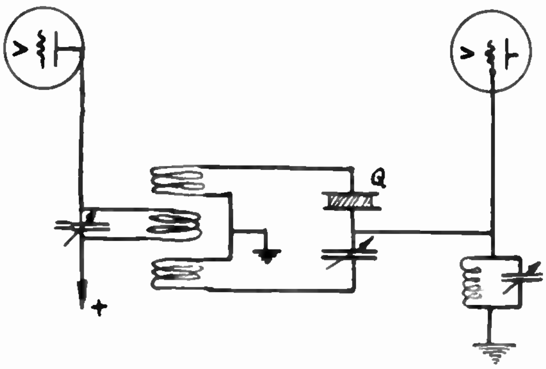

Figure 18

fig. 18. The anode circuit is tuned but the secondary circuit is not tuned. It comprises two coils connected in series and wound on the same bakelite tube as the primary coil, and arranged symetrically with respect to the primary. These two coils constitute two branches of the bridge, the connection between them being grounded. The two other branches of the bridge are the crystal with its electrodes and the neutralizing condenser, which may be the conventional trimmer. The junction of the two condensers is connected to the grid of the following tube. To fix the potential of the grid I use instead of the usual resistor, a tuned circuit. A simple resistor in this circuit arrangement would considerably reduce the gain of the stage because the effective input impedance of the tube must be large as compared with the coupling impedance which in this arrangement is practically the impedance of the vibrating crystal. The Q of this tuned grid circuit may be low and the coil may be made of solid wire because its resonant impedance must be of importance only as compared with the resonant impedance of the crystal which cannot be in any case more than a few thousand ohms. The important point, here, is to obtain high attenuation outside the crystal band. It means that the bridge must be perfectly balanced for a very broad frequency band, at least 40 to 50 KC. Theoretically, the balance of the bridge can be kept independent of the frequency only if the two coils are identical as well as the two condensers. In this case the ratio of the coil impedances as well as the ratio of the capacities is equal to 1 and independent of the frequency. Practically the two coils are never precisely equal,

especially because it is not only their self inductance which must consider, but also their mutual inductances with the primary coil.

If we call L the equivalent inductance of the first coil, R its resistance, $L + \Delta L$ and $R + \Delta R$ the corresponding values for the second coil, and $\Delta\omega$ its range of variation, V the potential across the two coils; then the voltage on the grid due to the variation of the frequency which as the result of lack of balance of the bridge will be given by

$$eg = \frac{\Delta\omega}{\omega}\left(\frac{\Delta R}{R} + \frac{\Delta L}{L}\right)\frac{V}{Q^2}$$

This value is obviously always negligible even if the coils are only approximately identical.

This is true, of course, only if we may assume that $\frac{1}{\omega C}$

is very large as compared to ωL, that is, if the coil system is operated at frequencies remote from the natural frequencies of the individual coils. If not, the expression for the coil impedance instead of being $\sqrt{\omega^2L^2+R^2}$

will become $$\frac{\frac{1}{\omega C}\sqrt{R^2 + \omega^2L^2}}{\sqrt{R^2 + (\omega L - \frac{1}{\omega C})^2}}$$

and R is then of importance as compared to $(\omega L - \frac{1}{\omega C})$

In this C is the capacity across the coil which is the distributed capacity plus the capacity due to the connections and the like.

Consequently it is suitable to be as far as possible from the resonance for every individual coil and have a Q reasonably high.

The balance of the bridge in the receiver which I am demonstrating is obtained by a variable balancing condenser and the two coils of the bridge are fixed. However it is possible to use a fixed balancing condenser, constituted by a plate of glass or mica, the balancing of the bridge being obtained either by varying the inductance of one of the coils or merely by changing its position relatively to the primary coil. The only element which must be adjusted with precision is the element determining the balance, because other circuits must not be sharp, the selectivity resulting largely from the crystal. Indeed their sharpness may spoil the curve. Consequently it is very important to choose all the elements on which the balance of the bridge depends so that they do not change with temperature or any other condition.

The crystal in the demonstration receiver is silver plated and the holder is a metal plate on which the crystal rests and which provides one of the contacts, the second contact being provided by a spring pressing upon the opposite face of the crystal.

The curve 1 on the fig. 19 represents the response of the demonstration receiver comprising a RF stage, a crystal transformer operating at 475 KC and a second and almost a periodic transformer. The curve 2 is the response of the same receiver with the same RF, but with two good iron core transformers. The gain in both cases is approximately the same. In this case the flat top is of about 7KC wide to obtain a very great selectivity, but it is of course possible to make it as broad as desirable. The same crystal without the RF gives a flat top of 8KC.

If broader band is desired the selectivity will of course decrease, but the attenuation at the ends of the band remains practically the same.

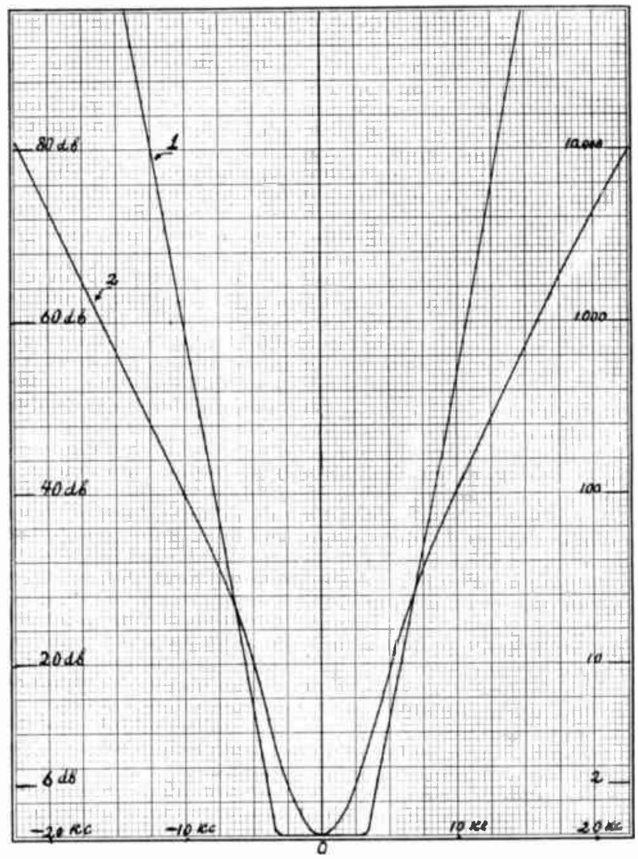

Figure 19

To obtain a variable selectivity it would be sufficient to provide a switch disconnecting the crystal. In this case the balance of the bridge will be destroyed and the resulting curve will be the curve of the electric circuit.

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11 West 42nd Street + + New York City

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PROCEEDINGS of the RADIO CLUB OF AMERICA

Volume 14 December, 1937 No. 3

THE APPLICATION OF THE BROAD BAND CRYSTAL FILTER TO BROADCAST RECEIVERS

BY

ALEXIS GUERBILSKY*

Delivered before the Radio Club of America
December 9, 1937

The piezo-electric crystals are known to have a very sharp resonance and are used either to control one single frequency with a sharp definition as, for instance, in radio transmitters or to resonate for a very narrow band as used for telegraphic, so called single frequency receptions, where they function as filters. There are however exceptions, for instance, in ultra-sonic sea signaling where the crystal is dampted by the radiated energy. In such applications the selectivity of the crystal is of no importance and the energy producing the damping is precisely the energy which is used. But, if we try to damp the crystal artificially in most applications its resonance curve will be less sharp, but it will still provide a characteristic not unlike that of a simple tuned electrical tuned circuit. For some other applications, as for instance, for optical relays, such artificially dampted crystals require too much energy for their operation and thus lose most of their valuable qualities.

It has long been obvious that crystals which would resonate over a broad band of frequencies would be extremely useful and, indeed, attempts have been made to obtain such crystals. The first thought was to artificially damp the crystal, but as explained above this method is in many cases unsuitable because it destroys the very characteristics of the crystal which make it especially useful.

The second thought was to use several crystals connected in parallel, the frequency of the several crystals differing from one another so as to cover the desired band. The difficulty in this expedient is that between the two frequencies corresponding to two successive crystals there is always one at which the two crystals vibrate in opposite phase, one crystal being operated below resonance and the other one above. This results in the fact that the electrical resonance curve has a sharp minimum-approaching zero - which makes the arrangement generally unsatisfactory.

Let us now discuss a solution of the problem making it possible to obtain a single crystal with a practically uniform response for any desired band of frequencies, while, at the same time providing a response abruptly decreasing at the limits of the band. This solution is, primarily, the use of a crystal of non-uniform thickness.

CRYSTAL OF NON UNIFORM THICKNESS.

EARLY YEARS OF TELEVISION.

When about 1926 the first steps had been taken in practical television, one of the greatest difficulties had been that of synchronization, this difficulty increasing with the number of the picture elements per image, i.e., definition. The idea of operating the scanning at the receiving station by the scanning system at the transmitting station seems to have been first suggested by Prof. Rosing in the early years of the century. This method makes the synchronization independent from the speed of scanning. To make this method practical it is possible to perform the television by using two parameters of a radio wave, its amplitude and its frequency; the amplitude corresponding to the luminosity of the transmitted picture element and the frequency to its sequential position.

Let us, in order to make it simpler, consider the transmission of only one line of the image supposing that the second scanning, at relatively low speed, can be made by other already known methods.

At the transmitting station a rotating optical scanning device is mechanically coupled to a rotatable, variable condenser which determines the frequency of the transmitted wave.

To every position of the scanning device there corresponds a frequency determined by the variable condenser and while the amplitude as determined conventionally by a photo-electric cell, corresponds to the luminosity of the transmitted picture element. At the receiving station we have then only to make the position of light spot correspond to the frequency of the received signal and its luminosity to the signal amplitude. The first idea was to use for this purpose a series of crystals of different frequencies corresponding to the received band width, each of these crystals constituting an electro-optical relay passing light when it is operated to the corresponding position. Its amplitude which is a function of the amplitude of the received signal determines the luminosity of the projected light. For this purpose the crystals have been disposed in a system of polarized light, their vibrations modifying the polarization and thus modulating the intensity of the light. Later this series of crystals were replaced by single crystal of a thickness varying from place to place to cover continuously the entire band of frequencies, every frequency determining a localized resonance in the portion of the corresponding thickness. As an example, a crystal of 10 cm. long and resonating over a band around 80 meters gave a localization of resonance of about 1 mm. The width of the band was about 300 kc.

SOUND RECORDING SYSTEM.

The next use of the crystals has been as a light valve for photographic sound recordings.

modulated practically without distortion up to 90%. The luminosity is very good. In one set of equipment, when the Eastman sound negative is used, with a gamma of 0.5, the density is 0.5 for an A.C. voltage across the crystal of about 100 volts.

SOME OTHER APPLICATIONS OF THE NEW CRYSTALS

In all its applications these crystals are intended to vibrate either simultaneously or successively for a band of frequencies, as in sound recording and television.

Amongst other applications and before discussing the band-pass filters, which is the main object of the present paper, we will only mention two. First, the control of a transmitter, the frequency of which it may be desired to vary to avoid possible interferencies; and second the use as microphone.

To control a transmitter a wedge shaped crystal can be used, one of the electrodes being a narrow plate sliding along its surface, the portion of the crystal positioned between the electrodes determining the frequency.

In its use as a microphone, the crystal vibrates at a high frequency and its impedence which depends largely on the damping determines the useful microphonic current.

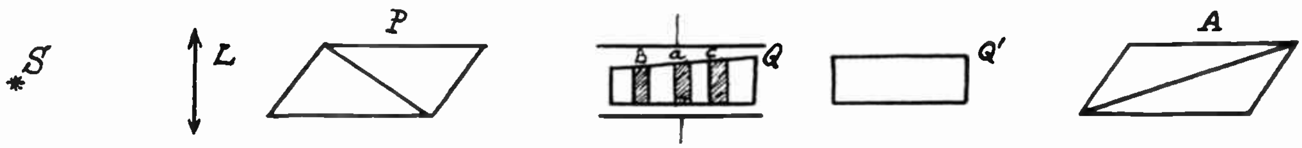


Figure 1

Figure 1 illustrates this application. The crystal, Q, is traversed by a beam of polarized light produced by source, S, a lens, L, and the polarizer, P. On its way the light traverses the portions of crystal of the different thicknesses. In the absence of any vibration of the crystal, the system is compensated by the compensator, Q', and the analyzer, A, at extinction. The crystal is excited by an oscillator of a frequency, f, corresponding to its middle portion, a. The generator is modulated by the microphonic current.

If the microphonic current is of a frequency F, there will be three portions of the crystal a, b, c, simultaneously set into vibrations corresponding to the frequencies f, f+F, f-F. The action on the polarization by the deformations of these three portions will be recomposed with conservation of phase and the luminosity of light correctly modulated for the whole band determined by the thicknesses of the crystal.

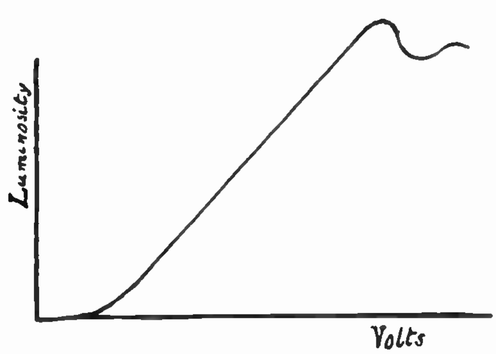


Figure 2

The curve giving the luminosity as function of the voltage across the crystal is shown in fig. 2. This curve has an important linear portion and the light can be

As a matter of fact the damping varies according to the density of the air around the crystal which density is modulated by the sound.

The same idea was used by Prof. Riabouchinsky and myself to make a dynamo-meter for aerodynamic measurements and recordings, the wedge shaped crystals making it possible to record rapid variations of air density or pressure.

The two last methods are schematically shown in fig. 3.

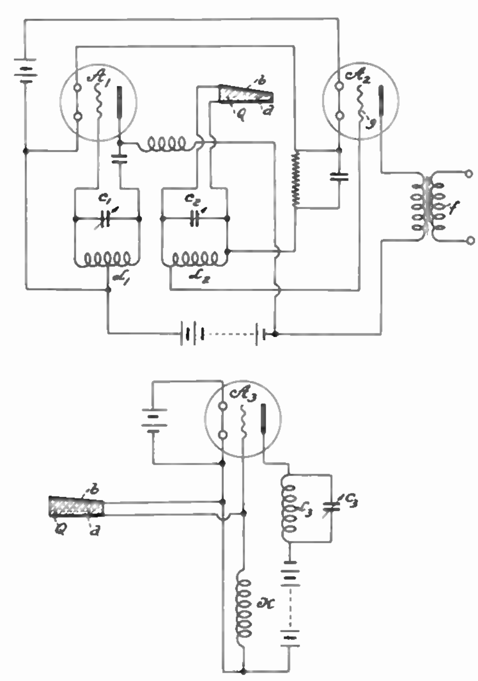


Figure 3

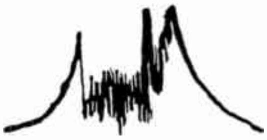


Figure 4

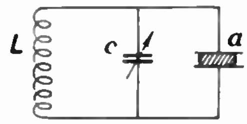


Figure 5

VIBRATION OF THE CRYSTAL.

In all the above discussion it has been assumed that the localized vibrations in the crystal travel continuously from one end to the other when the frequency varies. It is not however always so. If we record the resonance curve, this curve presents many irregularities. One of such curves can be seen on the fig. 4 which is an oscillogram of the voltage as function of frequency, of a circuit such as is shown in fig. 5, the "crevasse" of fig. 4 showing the resonance curve of the crystal.

To explain this curve and show how to correct the irregularities let us examine what occurs in the crystal when it is submitted to the action of an electric field of frequency varying between corresponding limits.

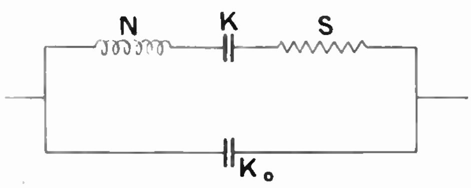


Figure 6

Let us consider first a usual crystal of uniform thickness. This crystal can be represented by the equivalent cell shown on the fig. 6, the inductance N, resistance S and capacities K and K_0 depending on the dimensions of the crystal. The corresponding resonance curve is shown in fig. 7. The sharp rise a corresponds to the resonance of the circuit NKS and the point b to the parallel resonance of the complete cell.

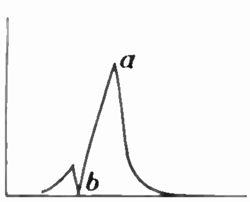


Figure 7

If now the crystal is of non-uniform thickness, it is represented by an infinity of small crystals of different thicknesses coupled together and the equivalent cell will be an infinity of elemental cells connected in parallel and coupled in such a way as to prevent any important shift of phase between any two adjacent cells. This condition corresponds to the mechanical realities of the crystal structure.

The resonance curve will be broad and will include all the frequencies of the band of the crystal, but to determine its shape we must not neglect all the phenomena which may occur.

First of all when the crystal is subjected to a frequency f a portion a, Fig. 8, is set in resonance vibration.

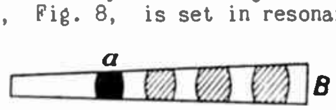


Figure 8

The corresponding elongation will also produce elongations in perpendicular direction the importance of which depends on the value of Poisson's modulus for the particular orientation of the crystal with respect to its axis. These latter elongations produce a wave propagated along the crystal in all directions. The edges of the crystal, such as the edge B, reflects a wave returning in a with a phase depending from the distance to B. When f varies the phase varies and the electrical effect will have as many maxima and minima as the varying phase. To avoid the reflections on the edges it is practically sufficient to dispose on these edges some absorbent material such as Canada balsam or any similarly viscous material. This will produce only useful damping, without markedly affecting the localized vibrations of the portion a.

Figure 9

Figure 10

The oscillogram of voltage as a function of frequency fig. 9 shows the characteristic of the crystal after the edge reflections have been attenuated as suggested above. There are still many maxima and minima due now to the coupling of the main mode of vibration with other resonators such as the harmonic resonances along the length, of flexion, torsion etc. These resonances are well understood and can be eliminated by localized damping. The curve resulting from such treatment as this is shown in fig. 10.

The curve of fig. 4 is of an X cut crystal as has been shown to illustrate better the different phenomenae. In the modern cut crystals, such as AT, AC, etc., the couplings and consequently the irregularities in the characteristic are greatly reduced and the corrections to be made are much easier and, when less uniformity is required, may be completely unnecessary.

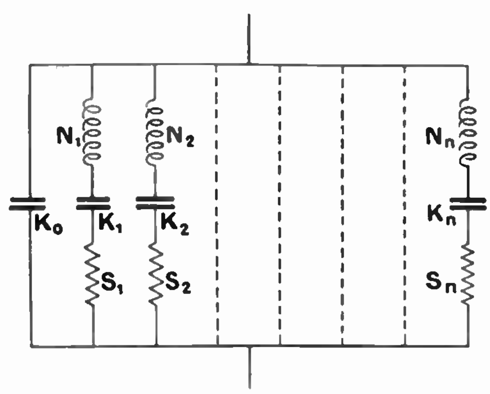


Figure 11

Now, assuming that all the spurious oscillations have been eliminated the crystal is fully represented by the equivalent cell of the fig. 11, (where the above discussed couplings between adjacent cells have been however omitted). The construction of the resonance curve is similar to the construction of the curve on the fig. 4. When the frequency reaches the value corresponding to the resonance of the first elemental cell, the current increases to the point A on the fig. 12 and then remains constant because of the identity of different elemental cells, until the antiresonance of the first cell is reached. At this point the current decreases to the point B and remains constant until at C the resonance of the last cell ceases. Then the anti-resonance of the

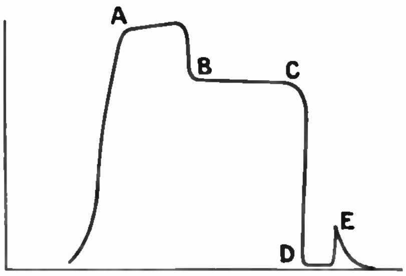


Figure 12

last cells produces a decrease of the current to the point D which corresponds to the point b on the fig. 4. The last anti-resonance ceases at the point E.

If we compensate the parallel capacity of the crystal by a bridge circuit, the current in the absence of resonance will be zero and any variation of the impedance will produce an increase of the current. The curve on fig. 12 will obviously become the curve on the fig. 13.

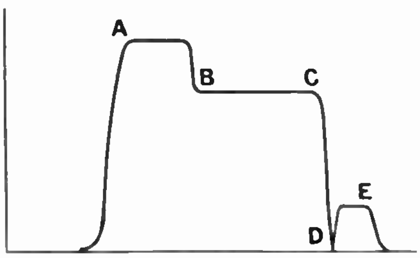


Figure 13

We see that this curve has three steps, but the importance of the height and width of the steps depends on the ratio of the portion of the crystal simultaneously vibrating to the whole of the crystal, or the number of elemental cells simultaneously resonating to the ensemble of cells. It means that by differently choosing the dimensions of the crystal, its length, thickness, slope, very different curves can be obtained which can be easily compared to the curve on the fig. 13. Some of such examples are shown by the oscillograms on the fig. 14, 15, 16.

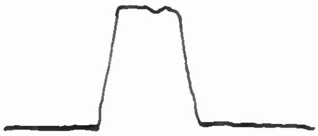


Figure 14

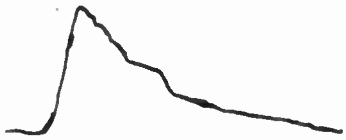


Figure 15

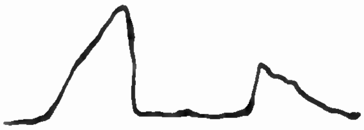


Figure 16

Furthermore, if the slope is too small the vibrating portion will be an important part of the crystal, as seen on the fig. 17 and for a small variation of frequency this portion will decrease and the corresponding curve will have no flat top and will resemble the usual curve of an electric tuned circuit. While if the slope of the crystal is too large the resulting damping may become seriously important.

Figure 17

When a crystal must be made for a predetermined frequency with a predetermined band it must be determined what is the value of the localization of the resonance, either theoretically, by considering the elastic constants of the crystal, usually quarz, or experimentally by analyzing the vibrating crystal in polarized light and by gradually changing the angle between faces.

I. F. TRANSFORMERS

The above described features of wedge shaped crystals make it possible to use the crystals to improve the response of pass-band filters and especially IF transformers. As a matter of fact, these crystals can be made for frequency bands as desired, the sharpness of the skirts of the resonance curve, being practically independent from the band width. Furthermore, the transformers using those crystals will have the selectivity practically independent from their frequency which can be, for example, as effective for 470 KC as for 1,500 KC or higher.

The circuit used for IF transformers is very much like the crystal transformers used in amateur so called single signal receivers, but I have tried to make them as simple as possible, because the price is in a radio receiver of a great importance. The circuit is represented on the

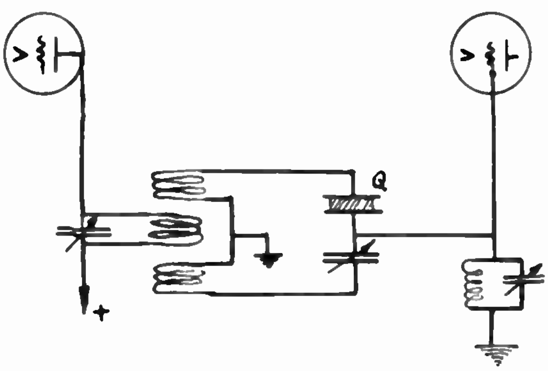


Figure 18

fig. 18. The anode circuit is tuned but the secondary circuit is not tuned. It comprises two coils connected in series and wound on the same bakelite tube as the primary coil, and arranged symetrically with respect to the primary. These two coils constitute two branches of the bridge, the connection between them being grounded. The two other branches of the bridge are the crystal with its electrodes and the neutralizing condenser, which may be the conventional trimmer. The junction of the two condensers is connected to the grid of the following tube. To fix the potential of the grid I use instead of the usual resistor, a tuned circuit. A simple resistor in this circuit arrangement would considerably reduce the gain of the stage because the effective input impedance of the tube must be large as compared with the coupling impedance which in this arrangement is practically the impedance of the vibrating crystal. The Q of this tuned grid circuit may be low and the coil may be made of solid wire because its resonant impedance must be of importance only as compared with the resonant impedance of the crystal which cannot be in any case more than a few thousand ohms. The important point, here, is to obtain high attenuation outside the crystal band. It means that the bridge must be perfectly balanced for a very broad frequency band, at least 40 to 50 KC. Theoretically, the balance of the bridge can be kept independent of the frequency only if the two coils are identical as well as the two condensers. In this case the ratio of the coil impedances as well as the ratio of the capacities is equal to 1 and independent of the frequency. Practically the two coils are never precisely equal,

especially because it is not only their self inductance which must consider, but also their mutual inductances with the primary coil.

If we call L the equivalent inductance of the first coil, R its resistance, $L + \Delta L$ and $R + \Delta R$ the corresponding values for the second coil, and $\Delta\omega$ its range of variation, V the potential across the two coils; then the voltage on the grid due to the variation of the frequency which as the result of lack of balance of the bridge will be given by

$$e_g = \frac{\Delta\omega}{\omega}\left(\frac{\Delta R}{R} + \frac{\Delta L}{L}\right)\frac{V}{Q^2}$$

This value is obviously always negligible even if the coils are only approximately identical.

This is true, of course, only if we may assume that $\frac{1}{\omega C}$ is very large as compared to ωL, that is, if the coil system is operated at frequencies remote from the natural frequencies of the individual coils. If not, the expression for the coil impedance instead of being $\sqrt{\omega^2 L^2 + R^2}$

will become
$$\frac{\frac{1}{\omega C}\sqrt{R^2 + \omega^2 L^2}}{\sqrt{R^2 + (\omega L - \frac{1}{\omega C})^2}}$$

and R is then of importance as compared to $(\omega L - \frac{1}{\omega C})$

In this C is the capacity across the coil which is the distributed capacity plus the capacity due to the connections and the like.

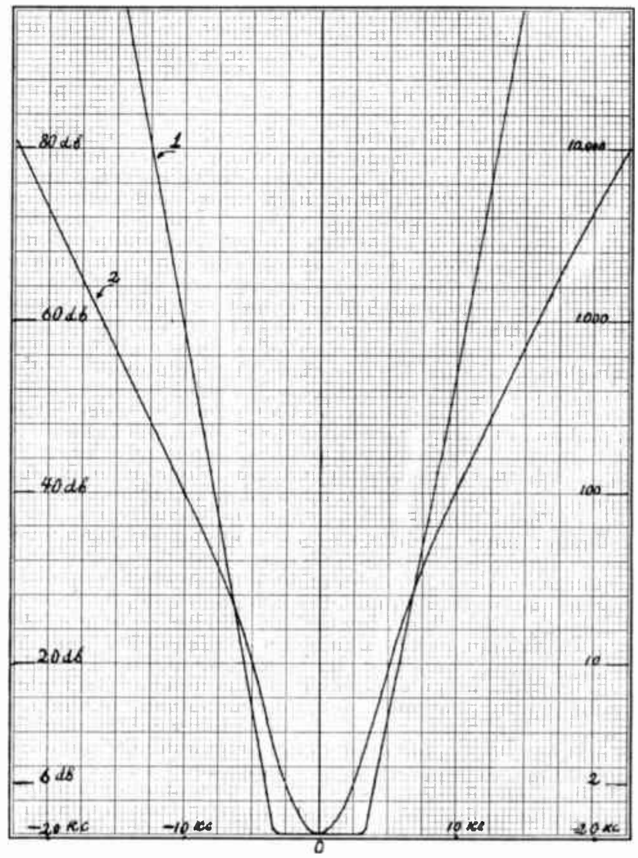
Consequently it is suitable to be as far as possible from the resonance for every individual coil and have a Q reasonably high.

The balance of the bridge in the receiver which I am demonstrating is obtained by a variable balancing condenser and the two coils of the bridge are fixed. However it is possible to use a fixed balancing condenser, constituted by a plate of glass or mica, the balancing of the bridge being obtained either by varying the inductance of one of the coils or merely by changing its position relatively to the primary coil. The only element which must be adjusted with precision is the element determining the balance, because other circuits must not be sharp, the selectivity resulting largely from the crystal. Indeed their sharpness may spoil the curve. Consequently it is very important to choose all the elements on which the balance of the bridge depends so that they do not change with temperature or any other condition.

The crystal in the demonstration receiver is silver plated and the holder is a metal plate on which the crystal rests and which provides one of the contacts, the second contact being provided by a spring pressing upon the opposite face of the crystal.

The curve 1 on the fig. 19 represents the response of the demonstration receiver comprising a RF stage, a crystal transformer operating at 475 KC and a second and almost a periodic transformer. The curve 2 is the response of the same receiver with the same RF, but with two good iron core transformers. The gain in both cases is approximately the same. In this case the flat top is of about 7KC wide to obtain a very great selectivity, but it is of course possible to make it as broad as desirable. The same crystal without the RF gives a flat top of 8KC.

If broader band is desired the selectivity will of course decrease, but the attenuation at the ends of the band remains practically the same.



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JANUARY, 1941

Volume 18, No. 1

EXTENDING THE RANGE OF ACOUSTIC REPRODUCERS

By Harry F. Olson

RADIO CLUB OF AMERICA, Inc.
11 West 42nd Street ◆ ◆ New York City

THE RADIO CLUB OF AMERICA, INC.

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Telephone -- Longacre 5-6622

MEETINGS

Technical meetings are held on the second Thursday evening each month from September through May at either Havemeyer or Pupin Hall, Columbia University, Broadway and 116th Street, New York. The public is invited.

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CLUB NEWS

NEW MEMBERS

We welcome the following new members elected at the January meeting of the Directors: Ray D. Hutchens, RCA Communications, 66 Broad St., N.Y.C.; F. A. Lidbury, Electro-Chemical Co., Niagara Falls, N.Y.; Harvey Sampson, 6 Bixby Drive, Baldwin, L.I., N.Y.; and Anton Schmitt, 516 West 136th Street, N.Y.C.

WILLIAM H. CAPEN

Members will be grieved to hear of the death of William H. Capen, which occurred on January 15th, following a few weeks of illness. He was Assistant General Technical Director of the International Telephone and Telegraph Company.

He was born in Newton, Mass., and educated at Harvard. For a number of years he engaged in telephone transmission work at the Bell Telephone Laboratories. As his responsibilities increased, he took greater interest in radio, and joined the Club in 1935, becoming a fellow in February 1940.

IMPORTANT WORK FROM PERCE COLLISON

Perce Collison is in the Navy now at the Naval Reserve Radio School, Noroton Heights, Conn., teaching rafts of fellows to be good ops. He writes particularly as follows: (1) A joyous welcome awaits Club members any Sunday afternoon from one till three (it is between Darien and Stamford--better plan a trip); (2) Club members likely to be drafted should go to Naval Reserve Headquarters, 90 Church Street, New York, and sign up for a year, asking for assignment to the Noroton Heights Radio School (he has this advice in red ink); and (3) He sends his best regards to all the gang and says that he really misses the meals and meetings together.

FEBRUARY PROCEEDINGS

The paper "Signal-Measuring Devices", presented before the Club last season by Jerry Minter of Measurements Corporation, Boonton, New Jersey, will appear in the February issue of the PROCEEDINGS. A general treatment is given, followed by practical discussion of stable tube voltmeters.

MARCH MEETING ON TUBES FOR U-H-F

On March 13th the Club will have two papers on the orbital-beam seconday-emission tube and the inductive-output tube, the authors being W. R. Ferris and H.M. Wagner of the RCA Manufacturing Company, and O.E. Dow of RCA Communications. The intense interest now felt in the ultra-high frequencies make this an important coming event.

****

A personal letter to each member from John L. Callahan, our new President, appears on the back cover of this issue.

PROCEEDINGS
OF THE
RADIO CLUB OF AMERICA

Volume 18 | January, 1941 | No. 1

EXTENDING THE RANGE OF ACOUSTIC REPRODUCERS

By Harry F. Olson*

Presented before the Club on November 19, 1940

Introduction

The almost universal use of the direct-radiator loud speaker is due to its simplicity of construction, small space requirements and the relatively uniform frequency characteristic. Uniform response over a moderate frequency band may be obtained with any simple direct-radiator loud speaker. However, reproduction over a wide frequency range is restricted by practical limitations. The portion of the speech range required for intelligibility falls in the mid-audio band. The range of the fundamental frequencies of most horn, reed, and string musical instruments also falls within this band. This is rather fortunate because it is a very simple task to build mechanical and acoustical vibrating systems to cover only this mid-frequency band. The two extreme ends of the audio-frequency band are the most difficult to reproduce with efficiency comparable to the mid-frequency range. Inefficiency at the low frequencies is primarily due to small radiation resistance. Inefficiency at the high frequencies is primarily due to large mass reactance.

The volume range is another factor involved in sound reproduction. The ear in the middle frequency band has a volume range of a million to one in pressure, or a trillion to one in energy. To build linear reproducing apparatus for this tremendous range is practically impossible today. As a matter of fact, it is not practical to reproduce the volume range of all musical instruments.

The practical difficulties involved in increasing the frequency and volume ranges are of course translated into increased cost. Economic considerations are involved in all apparatus designed for mass production and sold in large quantities. The relation between the volume and frequency ranges and the cost of radio receivers is illustrated by the data of Fig. 1. Graphs A, B, C, and D depict table-model and console receivers as sold today.

*RCA Manufacturing Company, Inc., Camden, N.J.

Fig. 1-E shows the average characteristics of various high-fidelity receivers which have been sold in the past few years — these sets have not had much success from the standpoint of sales volume, since the number sold is only a fraction of a percent of total sales. The volume and frequency ranges of speech and orchestral music are shown in G and H of Fig. 1. The upper volume range of certain instruments such as the organ is higher than the orchestra at the low frequencies.

The ambient noise in the room is another factor. The spectrum of room noise in the average living room is shown by the curve of small circles in Fig. 1-G. This curve gives the value of the spectrum level, $B = 10 \log_{10} (\Delta I \Delta f)$, where ΔI is the power per sq. cm. over the fre-

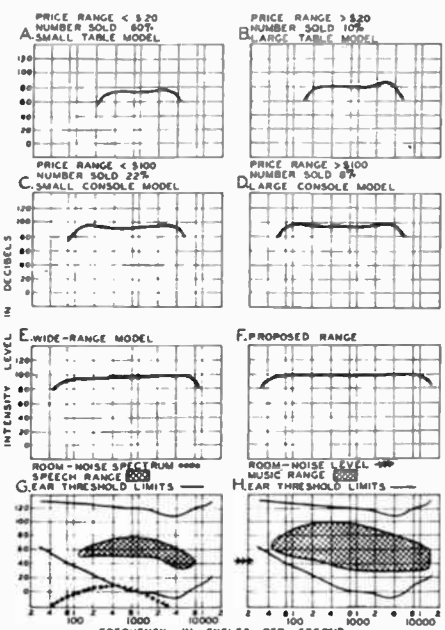

Fig. 1 - Frequency and volume ranges of radio receivers, and of music, speech, and residence room noise. 0 db = 0.0002 dynes per sq. cm.

quency interval from f to $f+\Delta f$. The curve is therefore a plot in decibels of the noise per cycle of the audio range. The intensity level for a finite band of frequencies is obtained by integration of $\Delta I/\Delta f$ between the desired frequency limits. The value in decibels is then obtainable as 10 times the common logarithm. In this way much of the spectrum-level curve of Fig. 1-G can be below the audibility curve, and yet represent noise that is easily heard under the usual conditions. The essential fact is that each point on the audibility curve represents a single tone which is the only sound present at the observation, while the noise curve is a power spectrum, of which a considerable frequency range is active simultaneously. An integration of $\Delta I/\Delta f$ over the entire audio range yields the intensity level shown by the triple arrow head in Fig. 1-H.

The frequency and volume ranges in Fig. 1 may not agree with those of other observers; however, the relative performance is the significant information shown in the figure, and this should agree with the conclusions of others. It is interesting to note that the volume and frequency range of a radio receiver is a function of the cost. Another interesting fact is the large sales in the lower price brackets in spite of the poor quality. The frequency and volume range of the small table model does not cover these ranges in speech. However, the performance is sufficient to give 90 percent syllable articulation, or almost perfect intelligibility. The reproduced sound quality is not good on speech and very poor on music. From these receivers the volume and frequency ranges increase steadily with increase in the cost. The frequency and volume ranges of a proposed receiver which will cover practically the full ranges of speech and orchestral music are shown in Fig. 1-F. It will be seen that the added frequency and volume ranges over those of the average console are considerable at both the ends of the frequency band.

An increase in the volume and frequency ranges of the loud speaker multiplies the problems connected with obtaining the proper directional pattern, low nonlinear distortion, and suitable transient response. The directional characteristics of the conventional direct-radiator loud speaker are quite adequate for the frequency range of the present-day broadcast receivers. However, when the high-frequency range is increased by one to two octaves, the directional pattern becomes quite narrow and some consideration must be given to this problem. The problem of nonlinear distortion is multiplied several times by the addition of one or two octaves. The additional volume range of course complicates the problem of nonlinear distortion. It has been found that poor transient response is not objectionable in the case of a receiver with a limited frequency range — in some cases it actually enhances the reproduction. However, a wide-range high-fidelity radio receiver should exhibit good transient response. From the above discussion it will be seen that the additional volume and frequency ranges increase the complexity of the technical problems manifold. It is the purpose of this paper to outline some of the factors involved in extending the range of loud speakers used in radio receivers.

Factors Involved In the High-Frequency Range

The simple dynamic loud speaker consists of a paper cone driven by a voice coil located in a magnetic field. The efficiency of this system in percent, when mounted in an infinite baffle, in terms of the resistivity and density of the voice coil, the mass of the cone, the air-mass reactance, the radiation resistance, and the flux density, is given by

$$\mu = \frac{100\, B^2\, r_{MA}\, m_C}{\rho K\,(x_{MA} - x_{MC} - x_{MD})^2 \times 10^9} \quad \text{---------} \quad (1)$$

where B = flux density in the air gap, in gausses;

r_{MA} = mechanical resistance of the air load upon the cone, in mechanical ohms;

m_C = mass of the voice coil, in grams;

ρ = density of the conductor, in grams per cc,

K = resistivity of the voice coil, in ohms per centimeter cube;

x_{MA} = mechanical reactance of the air load upon the cone, in mechanical ohms;

x_{MC} = mechanical reactance of the coil, in mechanical ohms; and

x_{MD} = mechanical reactance of the cone, in mechanical ohms.

This equation shows that from the standpoint of maximum efficiency it is desirable to make x_{MD} the mechanical reactance of the cone, and therefore the mass of the cone, as small as possible. For a particular cone the maximum efficiency occurs when the mass of the voice coil is equal to the air-load mass plus the cone mass. To fulfill this condition is not practical save at the high frequencies.

The impedance and efficiency characteristics[1] of loud speakers with cones of 16-inch, 4-inch and 1-inch diameter are shown in Fig. 2. The weights of the cones and voice coils of the 16-inch and 4-inch cones are typical of good loud speakers of these sizes in use today. The flux density is the upper limit in use today. The constants of the 1-inch cone were chosen to give approximately the same efficiency as the large cones at the low frequencies. A comparison of the characteristics shows that it is possible to obtain efficiency comparable to that of a large cone over a wide range by using a small cone and coil.

SYSTEM	A	B	C
DIAMETER, INCHES	16	4	1
MASS OF CONE, GRAMS	40	1	.015
MASS OF VOICE COIL, GRAMS	4	.35	.015
VOICE-COIL MATERIAL	COPPER	COPPER	ALUMINUM
AIR-GAP FLUX, GAUSSES	10000	10000	10000

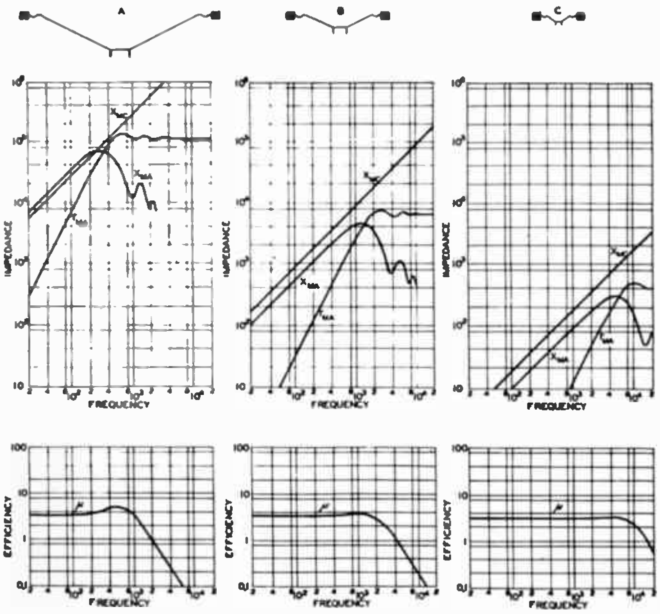

Fig. 2 - Impedance-frequency characteristics of three loud speakers having 1-inch, 4-inch and 16-inch cone diameters, and the efficiency-frequency characteristics of the three loud speakers: X_{MC} = mechanical reactance of the cone and coil; X_{MA} = mechanical reactance of the air load; r_{MA} = mechanical resistance due to the air load; and μ = efficiency.

The curves of Fig. 2 show that a mass-controlled system delivers constant output below the point of constant radiation resistance, r_{MA}. To deliver constant output in the range where the resistance is a constant, the impedance of the entire system must be independent of the frequency. By suitable processing of the cone it is possible to reduce the impedance at the higher frequencies. In any case there is some wave propagation in the diaphragm at the higher frequencies which, in effect, reduces the impedance of the vibrating system.

The inductance, skin effect, and magnetic hysteresis in the voice coil, in combination with the existing vacuum-tube driving system, are other factors which reduce the response in a dynamic loud speaker at the higher frequencies. Power amplifiers are generally designed so that the voltage across the loud speaker is independent of the frequency for constant voltage applied to the input of the power stage. Therefore, the current in the voice coil is inversely proportional to the impedance. The impedance characteristics[2] of several voice coils are shown in Fig. 3. In the case of a large, heavy voice coil the rapid increase of the impedance at the higher frequencies causes a corresponding reduction in the driving force. To maintain constant driving force at the higher frequencies requires a relatively low ratio of the inductance to the resistance which, for a constant value of the resistance, is equivalent to a reduction in the mass of the voice coil.

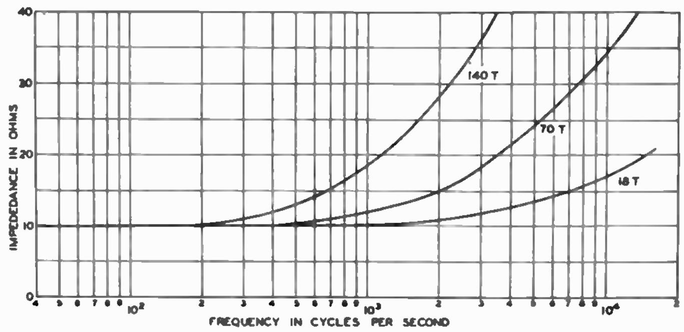

Fig. 3 - Impedance-frequency characteristics of voice coils of 1.5-inch diameter and 140, 70, and 18 turns; all have 10 ohms d-c resistance.

Equation 1 shows that the efficiency is proportional to the square of the flux density. To reduce the cost of the loud speaker, the copper in the average field structure has been decreased from several pounds to less than a pound in the last few years. To make up for the loss in acoustic power output, the power output of the amplifier has been correspondingly increased. To accomplish this, power output systems have been used which inherently possess high distortion. In some cases the power output is so large that the magnetomotive force produced by the voice coil is comparable to the magnetomotive force produced by the field coil. The result of the interaction between these two fields is nonlinear distortion. For a moderate frequency range the subjective effect of these nonlinear distortions does not appear to be of much consequence. However, for wide-range operation, such output systems cannot be tolerated.

The directional characteristics of a dynamic cone loud speaker are a function of the frequency. At the low frequencies, where the dimensions of the cone are small compared to the wavelength, the system is nondirectional. At the higher frequencies the directional pattern becomes progressively sharper. The directional characteristics of a cone 8 inches in diameter (10-inch loud speaker) are shown in Fig. 4. For the range up to 5000 cycles the directional pattern is satisfactory. However, above 5000 cycles the directional pattern becomes quite sharp. The characteristics can be broadened in this range by corrugating the cone, a procedure which reduces the effective diameter of the cone at the higher frequencies. The characteristics can also be broadened by using a distributor, as shown in Fig. 5. The effect of a distributor[3] may be obtained by comparing Figs. 4 and 6. Reducing the diameter of the cone also broadens the directional pattern as shown by the characteristics for a 2-inch cone in Fig. 7. As the frequency range is increased it is quite important to maintain a broad directional pattern.

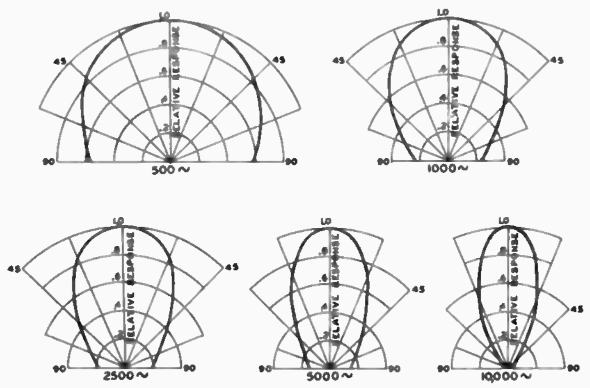

Fig. 4 - The directional characteristics of an 8-inch cone loud speaker at various frequencies.

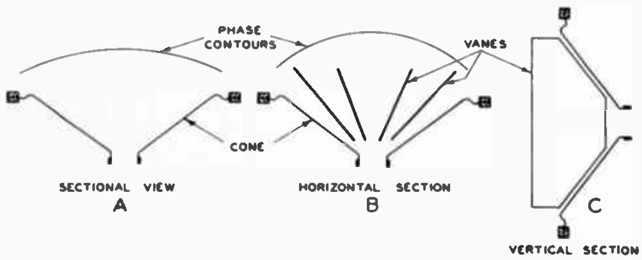

Fig. 5 - High-frequency sound distributors for direct-radiator loud speakers: A - Contour of equal phase for a plain cone; B - Horizontal cross-sectional view of a cone with vane distributor and the contour of equal phase; C - Vertical cross-sectional view of a cone with a vane distributor.

The nonlinear distortion at the high frequencies is usually quite low in well designed and carefully built loud speakers, because the amplitude of motion is quite small and the forces developed in the system are well within the linear characteristic of the materials.

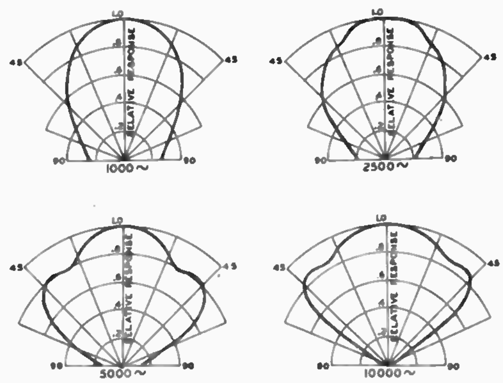

Fig. 6 - The directional characteristics of the cone loud speaker of Fig. 4 with distributor of Fig. 5.

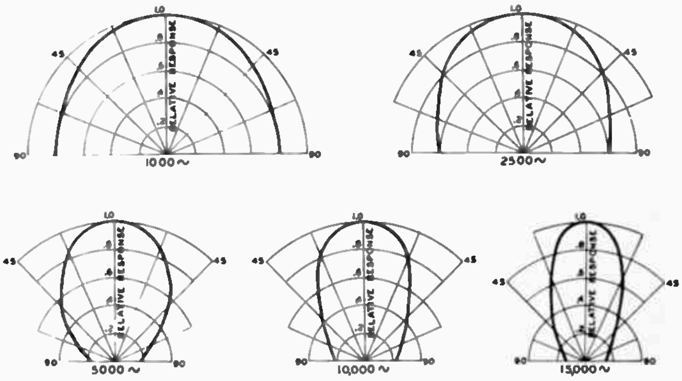

Fig. 7 - The directional characteristics of a 2-inch cone loud speaker at various frequencies.

The transient response at the high frequencies is usually quite good when the response characteristic is smooth. In the case of sharp peaks and dips, the transient response will usually be poor, a condition which is subjectively manifested as fuzzy or unclean reproduction.

Factors Involved In the Low-Frequency Range

The inefficiency of a direct-radiation loud speaker at the low frequencies is primarily due to the small radiation resistance. Referring to Fig. 2, it will be seen that up to a certain frequency the mechanical resistance is proportional to the square of the frequency. In this range the velocity must be inversely proportional to the frequency to obtain uniform acoustic output with respect to frequency. The output as a function of the frequency for the 16-inch cone of Fig. 2 when it is controlled by mass, resistance, and stiffness, is shown in Fig. 8. This shows that the system should be mass-controlled at

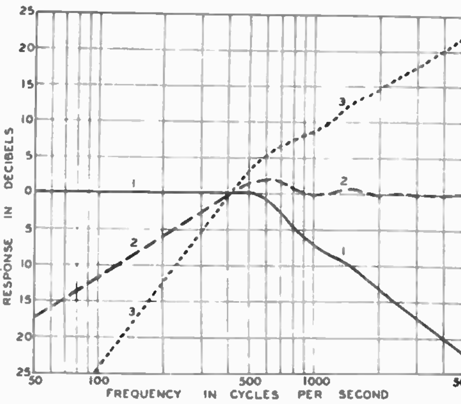

Fig. 8 - The response of a 16-inch cone for various types of control: 1 - Mass control; 2 - Resistance control; 3 - Stiffness control.

the low frequencies to obtain uniform response. Furthermore, the fundamental resonance must be placed at the lower limit of reproduction. The characteristics of Fig. 8 are for an infinite baffle. In the case of a finite baffle the system is an acoustic doublet at the low frequencies where the dimensions are small compared to the wavelength. At the high frequencies, where the dimensions of the baffle are large compared to the wavelength, the two sides of the cone act as separate sources of sound. In a finite baffle the radiation resistance changes from the fourth power of the frequency to the second power at the transition from doublet to singlet operation. The response of a mass-controlled system for various sizes of baffle is shown in Fig. 9. It will be seen that the output falls off 6 db per octave below the transition point from doublet to singlet operation. The performance of a mass-controlled unit in a cabinet is quite similar, as shown in Fig. 10, the only difference being due to cabinet resonance.

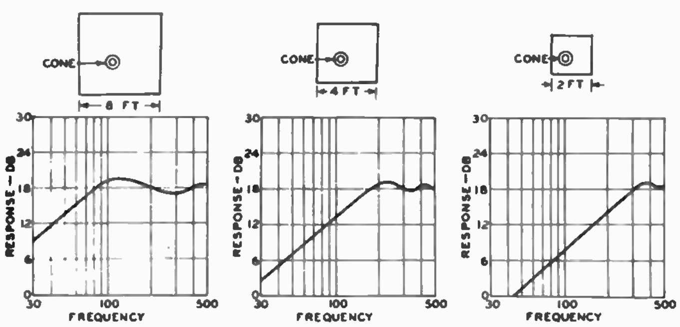

Fig. 9 - Response-frequency characteristics of a direct-radiator loud speaker mechanism with a resonant frequency of 20 cycles, mounted in baffles of three sizes. The loud speaker is mounted off center to reduce selective-interference effects due to interaction between the two sides of the cone.

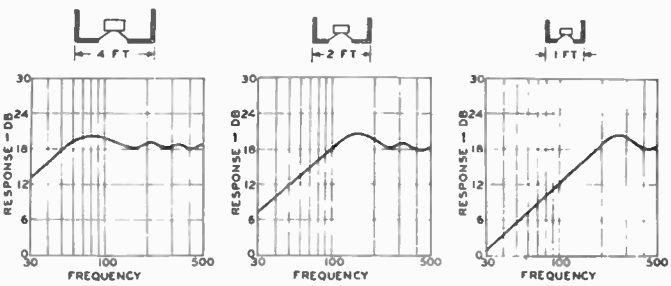

Fig. 10 - Response-frequency characteristics of a direct-radiator loud-speaker mechanism with a 20-cycle resonant frequency, mounted in open-back cabinets of three sizes.

The power output of a direct-radiator loud speaker at the lower frequencies is another factor of great importance in wide-range systems. The peak amplitude versus frequency[4] of vibrating pistons of 16-inch, 4-inch and 1-inch diameter, mounted in an infinite baffle, for one watt of sound output, is shown in Fig. 11. Since the excursion of the diaphragm is limited, these characteristics show that a relatively large cone is required to deliver adequate power at the lower frequencies. In addition a relatively heavy cone is required in order to prevent generation of harmonics due to spurious vibrations of the large surfaces. The construction must be rugged so that the forces will be kept within the elastic limits of the materials.

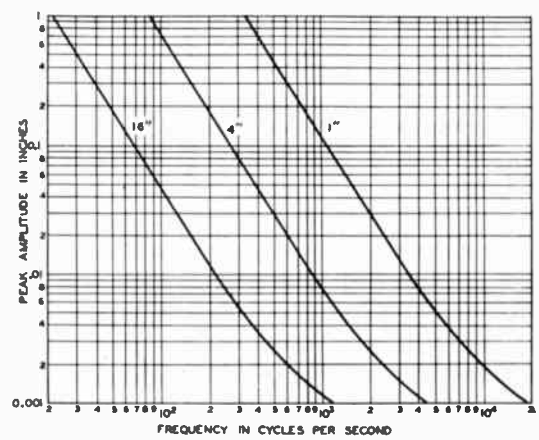

Fig. 11 - The amplitude-frequency characteristics of vibrating pistons of various diameters mounted in an infinite wall, for one watt output on one side.

Nonlinear distortion occurs when a nonlinear element is present in the vibrating system. The outside diaphragm suspension is an example of a nonlinear element. The stiffness is not a constant but is a function of the amplitude and, in general, increases for larger amplitudes. A conventional dynamic loud speaker and the equivalent electrical circuit are shown in Fig. 12. Above the fundamental-resonance frequency[5] the velocity is not appreciably affected by the suspension because the reactance of the compliance C_{MS} is small compared to the impedance of the remainder of the system. Below the resonant frequency the reactance of the compliance is the controlling impedance. In this range the nonlinear characteristics of the suspension will be most marked. The curves of Fig. 12 show the total distortion for 2, 5, and 10 watts input to the loud speaker. The distortion increases with input and with decrease of the frequency. The fundamental resonance of this loud speaker occurs at 80 cycles. It will be seen that the distortion is very small above the resonant frequency, where the influence of the compliance is small. The distortion due to the suspension may be obviated by placing the fundamental-resonance frequency of the loudspeaker at the lower limit of the reproduction range.

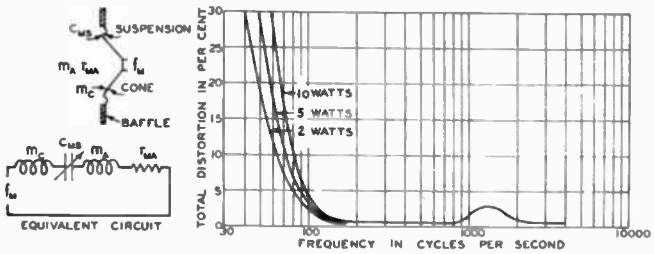

Fig. 12 - Cross-sectional view of the vibrating system of a 10-inch dynamic loud speaker mounted in a large baffle. The graph shows the distortion-frequency characteristics for inputs to the voice coil of 2, 5 and 10 watts, respectively.

If the stiffness of the conventional suspension system is reduced so that a low resonant frequency is obtained, the result is a very fragile vibrating system. To obviate this difficulty a new type of suspension combining ruggedness and small stiffness has been developed. A loud speaker employing this new folded suspension is shown in Fig. 13. This suspension reduces the radial constraining forces which arise in the conventional suspension, and thus reduces the stiffness. In addition, this suspension presents a constant stiffness over a greater amplitude range. The nonlinear distortion characteristics of two loud speakers having the same field structure, voice coil, and cone diameter, one with a conventional suspension and the other with a folded suspension, are shown in Fig. 14. These results show a large reduction in nonlinear distortion.

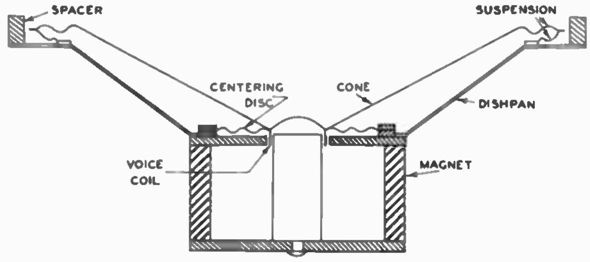

Fig. 13 - Cross-sectional view of a direct-radiator loudspeaker mechanism with a folded suspension system.

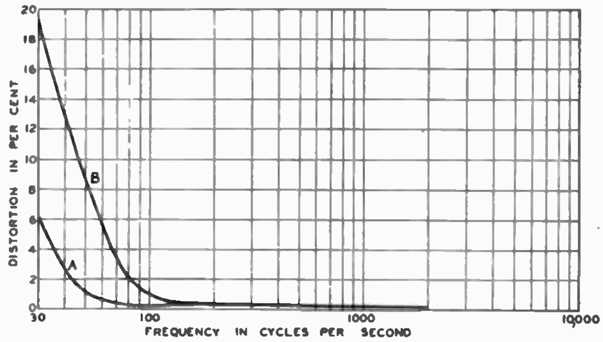

Fig. 14 - Distortion-frequency characteristics with (A) folded suspension system; and (B) standard suspension system.

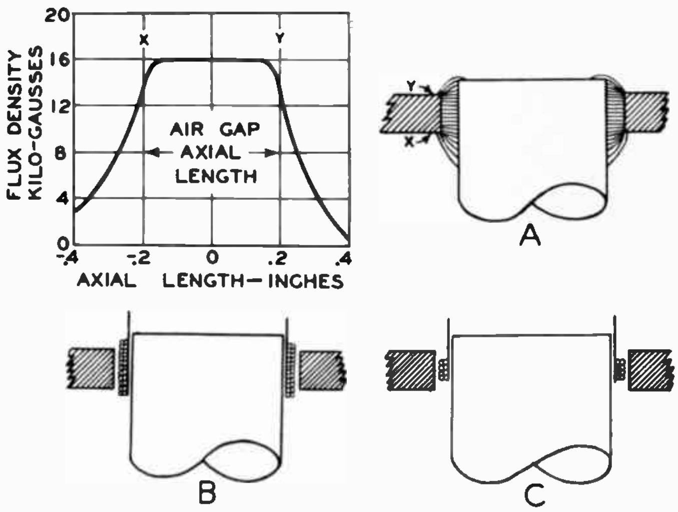

Fig. 15 - Flux and voice coils in the air gap. The plot shows the flux distribution in an air gap. The other sketches show: (A) - Typical distribution of the flux lines in the air gap: (B) - Arrangement of parts with voice coil longer than the air gap; and (C) - Arrangement with voice coil shorter than the air gap.

Inhomogeneity of the flux density[6], through which the voice coil moves, is another source of distortion. This may be seen from the plot and Sketch A of Fig. 15. The result of inhomogeneity is that the driving force does not correspond to the voltage developed by the generator in the electrical driving system. Furthermore the motional impedance is a function of the amplitude. This type of distortion can be eliminated by making an air gap of such dimensions that the voice coil remains at all times in a uniform field, as shown in Fig. 15-C. This type of distortion can also be eliminated by making the voice coil longer than the air gap so that the summation of the products of each turn and the flux density is independent of the amplitude; this construction is shown in Fig. 15-B.

Transient Performance

The sounds of speech and music are of a transient rather than of a steady-state character. Therefore practically all sounds which are reproduced in radio may be considered to be of a transient character, and the transient response of a loud speaker is an important factor in sound reproduction. The performance of a loud speaker in this respect may be determined by means of a suddenly applied voltage.[7] The transient performance of a loud speaker is poorest at the fundamental resonance of the loud speaker and cabinet. When a loud speaker operates from a high-impedance vacuum tube, the internal mechanical resistance of the loud speaker is the major factor in influencing the transient response. The response of a 12-inch loud speaker to a unit force for various values of mechanical resistance is shown in Fig. 16. To correlate with the response of other systems, the impedance-frequency characteristics for each system are included. These characteristics are for a loud speaker coupled to a generator with a very high internal impedance. For this type of operation it is customary to provide a large mechanical resistance r_{MS}, the middle and lower conditions of Fig. 16.

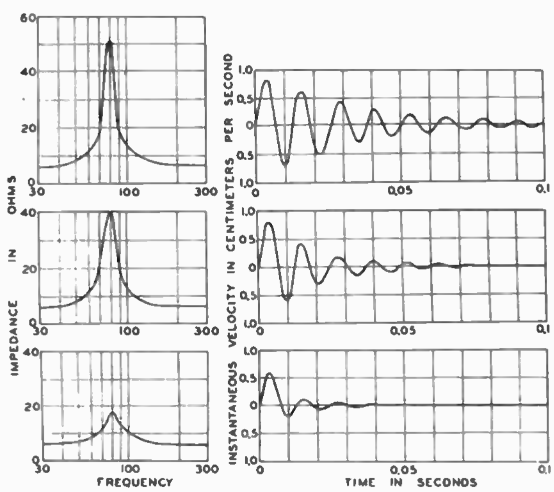

Fig. 16 - The unit-force response of a loud speaker coupled to a generator of very high resistance, for different values of the internal mechanical resistance. The impedance-frequency characteristics indicate the degree of internal damping.

In Fig. 17 there is shown the effect of the impedance of the vacuum tube upon the transient response of a loudspeaker. In this case the loud speaker is connected first to a very high resistance, then to a generator of one-half the resistance of the loud speaker, and finally to a generator of very low impedance. Fig. 17 shows that the damping exerted by the electrical system is of consequence.

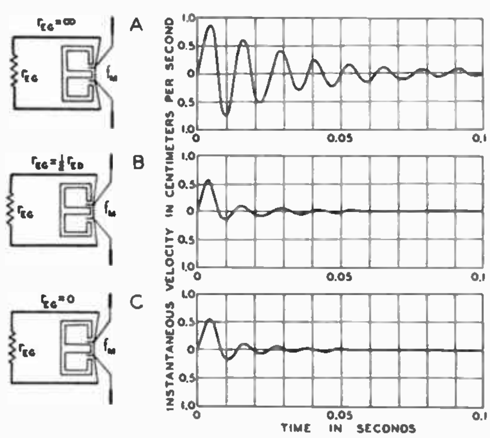

Fig. 17 - The transient response of a 12-inch loud speaker to a unit force for various generators.

Experimental results on the transient response of various systems are shown in Fig. 18. Here Fig. 18-A shows the transient reponse of a small table-model receiver; Fig. 18-B shows the transient response of a console receiver with a high-impedance output system; Fig. 18-C shows a console receiver with a low-impedance output system; Fig. 18-D shows the transient response of a combination horn and direct-radiator loud speaker. The effect of the high radiation resistance of the horn in providing damping is quite evident.

A number of investigators have stated that square waves cannot be obtained from acoustic systems. The response of a combination horn and direct-radiator loud speaker to square waves of various frequencies is shown in Fig. 19. The response of this loud speaker is essentially

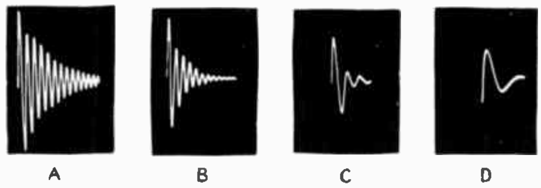

Fig. 18 - Oscillograms showing the transient response of a loud speaker to a unit voltage: (A) - Small table-model loud speaker; fundamental resonance 170 cycles; high-impedance driver. (B) - Console-model loud speaker; fundamental resonance 110 cycles; high-impedance driver. (C) - Console-model receiver; low-impedance driver; fundamental resonance 100 cycles. (D) - Combination horn and direct-radiator loud speaker; low impedance driver.

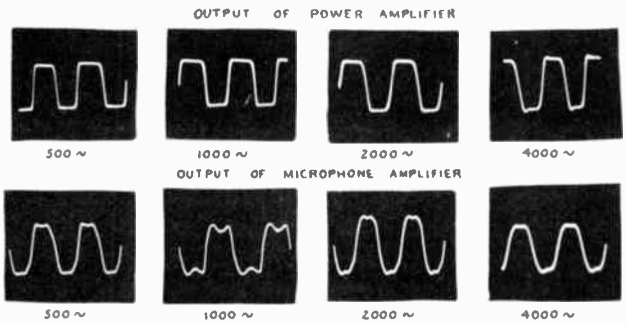

Fig. 19 - Oscillograms showing the shape of square waves after passing through a power amplifier feeding a loud speaker, and after passing through the loud speaker, microphone, and pre-amplifier, for frequencies of 500, 1000, 2000, and 4000 cycles.

uniform over the range from 60 to 12,000 cycles. The response of the amplifier feeding the loud speaker is also shown in Fig. 19. The transient response of this loud speaker is quite comparable to that of other audio elements covering this frequency range. Fig. 19 shows that square waves can be obtained from acoustic systems.

Systems for Wide-Range Reproduction

The preceding sections have been concerned with a discussion of the factors involved in the design of a loud speaker for radio reception which will exhibit the following characteristics: wide frequency range, large volume range, low nonlinear distortion, uniform directional characteristics, and good transient response. It is the purpose of this section to illustrate some of the systems suitable for wide-range radio reproduction.

The characteristics of Fig. 2 show that the low-frequency efficiency may be maintained to the high-frequency ranges by employing a small, relatively light cone and coil. On the other hand, to obtain adequate power-handling capacity at the lower frequencies require a cone of relatively large area. To insure operation below the elastic limits of the materials, a cone of large area must be of sturdy construction. A large cone of this type must be driven by a relatively large coil to obtain tolerable efficiency. It follows that the efficiency of such a system must be low in the high-frequency range.

Adequate power-handling capacity and response over a wide frequency range may be obtained by employing a single-coil single-cone loud speaker in two different arrangements. The first of these is an arrangement of a large-diameter cone driven by a heavy coil for the reproduction of the low-frequency range, a small-diameter cone driven by a light coil, and a filter for dividing the frequency range properly between the two speakers; this is shown in Fig. 20-A. The second arrangement consists of small-diameter cones in sufficient number to satisfy the low-frequency power requirements and of light enough construction to satisfy the high-frequency response requirements; this is shown in Fig. 20-B.

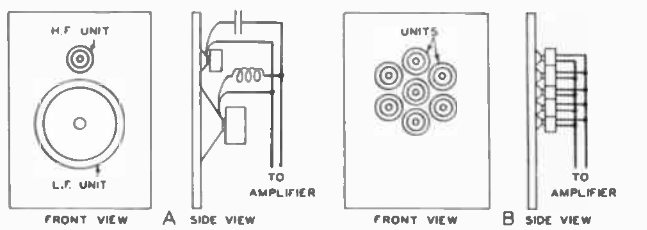

Fig. 20 - Multiple single-cone single-coil loud speaker: (A) - Large low-frequency unit, small high-frequency unit, and filter system; (B) - Seven small units connected in parallel.

The double-coil double-cone loud speaker[8] consists of a light coil coupled to a small cone, connected by a compliance to a heavy coil and large cone. This is shown in Fig. 21. In this system an increase in range is obtained by reducing the impedance of both the coil and the diaphragm at the high frequencies. Both parts of the voice coil are used at the low frequencies; no air-gap flux is wasted as would be the case if two separate units were used.

In the above systems the response is maintained to the higher frequencies by reducing the effective mass of the vibrating system with increase in frequency. A wide directional pattern is maintained as the high frequencies by reducing the effective diameter of the vibrating system with increase in frequency.

As indicated by Fig. 10, to maintain good response, efficiency, and power-handling capacity down to 30 cycles requires a relatively large cabinet. An improvement in low-frequency efficiency can be obtained by the use of the following: an acoustic phase inverter[9], as shown in Fig. 22; an acoustical labyrinth[10], as shown in Fig. 23; or a combination horn and direct radiator[11], as shown in Fig. 24.

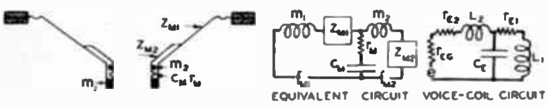

Fig. 21 - Cross-sectional view of the double-cone loud speaker, the equivalenycircuit of the mechanical system, and the voice-coil circuit.

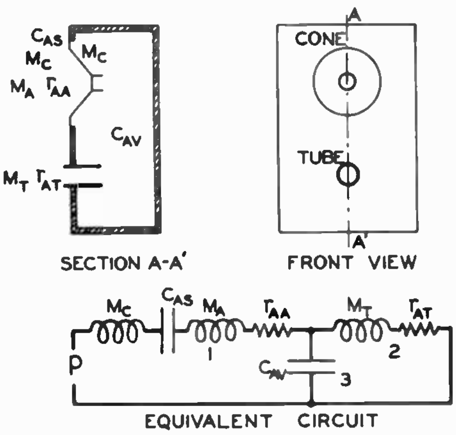

Fig. 22 - Phase-inverter loud speaker. After Dickey, Caulton and Perry.

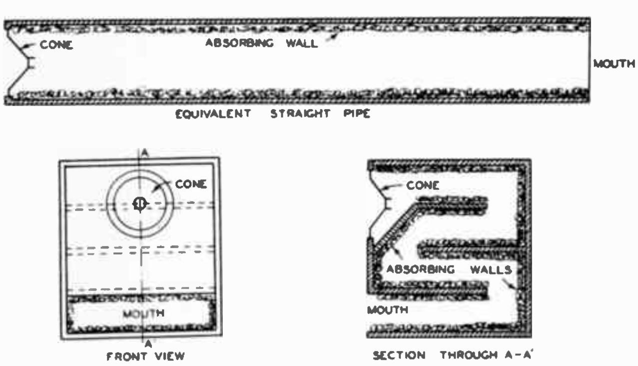

Fig. 23 - Acoustical-labyrinth loud speaker. After Olney.

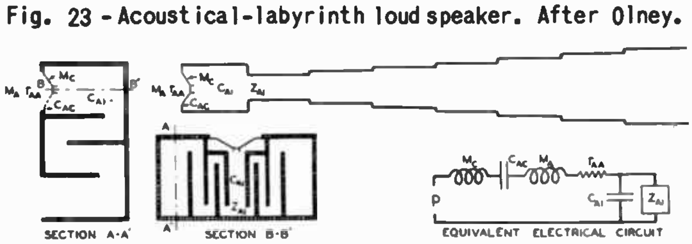

Fig. 24 - Combination horn and direct-radiator loud speaker. After Hackley and Olson.

Conclusion

An examination of the multiple vibrating systems and the low-frequency systems described above shows that their cost will be considerably greater than that of the convention single-unit single-coil single-cone loud speaker mounted in a conventional cabinet. Great care must be exercised in both the design and manufacture to hold the nonlinear distortion to a low level. Reproduction below 40 cycles requires very large cabinets regardless of the system. Good reproduction down to 20 cycles with any of the systems known today results in cabinets which would not be tolerated by anyone, save those interested in reproduction regardless of size. In this case good reproduction over this range does not imply a response down 20 decibels at the extreme ends of the frequency range.

At the present time it appears that the problem of extending the frequency and volume ranges is to a large extent economic rather than technical, because, as outlined in this paper, much technical information for improving the performance is known. As in the past, some improvements will eventually be made without added cost, or in some instances with a reduction in cost. However, progress along these lines is, in general, slow. To effect a major improvement in the performance of a radio receiver at the present time by an extension of the volume and frequency ranges will increase the cost. Of course, if wide-range radio receivers are accepted by the public and sold in considerable quantities, the differential in cost between wide-range and existing commercial reproducers can be reduced to the point where it will not be a detrimental sales factor.

[1,2,3,4,5,6,7] Olson, "Elements of Acoustical Engineering", D. Van Nostrand Company, New York City, 1940.

[8] Olson, JOUR. ACOUS. SOC. AMER., Vol. 10, No. 4, P. 305, 1939.

[9] Dickey, Caulton and Perry, RADIO ENGINEERING, Vol. 8, No. 2, P. 104, 1936.

[10] Olney, Benj., JOUR. ACOUS. SOC. AMER., Vol. 8, No. 2, P. 104, 1936.

[11] Olson and Hackley, PROC. INST. RAD. ENG., Vol. 24, No. 12, P. 1557, 1936.

The Radio Club of America, Inc.
11 West 42nd Street
New York City

RCA
FOUNDED 1909

Fellow Members of The Radio Club of America:

The Radio Club of America is the pioneer radio society in the United States. This fact is known to the "Old Timers" and we hope to those who have recently joined the Club. Don't keep it to yourself - tell others about it. Another point which should be known and practiced by all concerned is that the Club is "dedicated to the free interchange of ideas among all radio enthusiasts". We are rightfully proud of the papers presented at our monthly meetings and the free and open discussion which follows.

New members are essential if the Club is to continue through the coming years to render maximum service to those interested in the radio art. Our past record is an excellent one, requiring special effort to equal or better it. Every member should consider himself a membership committee of one and proceed to sell the Club to associates who are not members. Ask for several copies of the Club membership folder. Become acquainted with the contents, pass the word along, and bring in the applications.

Our meetings are too good to miss. Mark the second Thursday of each month on your calendar as a "Must Date" at Columbia University and follow through. If you become accustomed to attending, the effort will be repaid many times over. The dinner get-togethers in John Jay Hall prior to the meetings provide an excellent opportunity to meet "the gang" - don't miss it.

The questionnaire recently sent to the members has made available a wealth of suggestions regarding Club activities. We are endeavoring to reduce them to practice, and trust you will continue to keep the Club in mind and turn in suggestions as they occur to you.

The Officers and Directors appreciate the honor and responsibility you have conferred upon us, and with your help will endeavor to make 1941 a banner year for the Club.

J. L. Callahan

J. L. Callahan
President