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Number 130 MARCH — APRIL 1948

WELDING SMALL PARTS OF A MINIATURE VALVE AT THE RADIOTRON VALVE WORKS, ASHFIELD, N.S.W.

RADIOTRONICS

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IN THIS ISSUE

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Our cover portrays one of many operators carrying out the final assembly operation in the manufacture of Miniature Radio Valves. Various parts are assembled by the operator and fixed in position by means of a small electric spot welder. Some of these operations, such as inserting and anchoring the filament are very delicate and require the use of a magnifying glass to ensure true alignment with other elements of the valve.

"PRACTICAL ANALYSIS OF ULTRA HIGH FREQUENCY TRANSMISSION LINES. RESONANT SECTIONS. RESONANT CAVITIES. WAVE GUIDES."

by J. R. MEAGHER and H. J. MARKLEY (Part 2).

This is the second of two articles (Part 1 appeared in Radiotronics 129) dealing, in a simplified manner, with u-h-f transmission lines, etc., and is published through the courtesy of the R.C.A. Service Company, copyright proprietor. This article deals with wave guides and cavity resonators, and includes an appendix on the use of the slotted line.

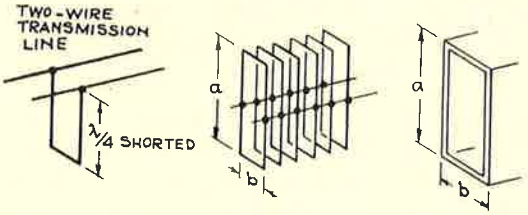
Wave Guides

A wave guide is a simple hollow metal tube having no central conductor. The losses are relatively low, since they will be produced mainly by the "inner skin" of the tube (which is of large perimeter and hence gives low loss). The *inner surface* should be clean and smooth. The *outer surface* can be grounded at any point since the r-f penetrates only a thin skin of the inner surface. *Sharp bends* are usually avoided, and all bends and twists are arranged to prevent a change in "mode" of propagation, or reflections. Instead of a hollow metal guide, a *solid dielectric* may be used as a wave guide. The action in this case is comparable to light waves traveling inside a *lucite rod.* In general, the loss in a solid dielectric wave guide is greater than in a hollow wave guide.

A wave guide cannot be conveniently treated like an ordinary transmission line. Wave guides must be approached from the viewpoint of an electromagnetic wave in a dielectric, using the same basis of treatment as that of radiation.

Wave guides may be rectangular, round, or oval. At the present time, rectangular wave guides are most simple and common; this discussion will refer to rectangular wave guides for the most part, but much of this information can be extended to guides of other shapes.

The following development of a simple type wave guide is intended to serve as a means of bridging the gap between transmission lines and wave guides, although they operate on different principles.

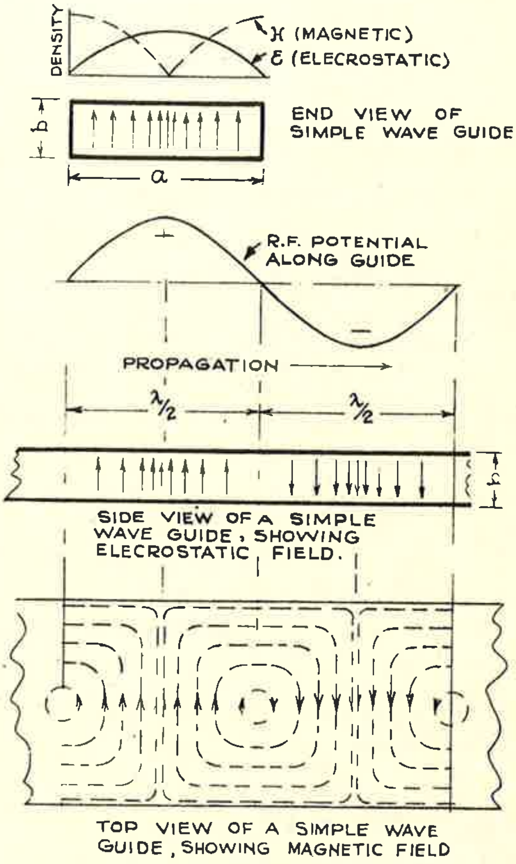


Above at left is shown a section of two wire transmission line. A quarter-wave resonant stub has been added across the line. The ends of the stub where connected to the transmission line, will represent a high impedance—many times higher than the impedance of any practical two wire transmission line. As a result the stub will have a negligible effect.

Suppose an infinite number of quarter-wave shorted stubs are added, resulting in a continuous pipe of rectangular cross section, or one type of wave guide. See illustration above.

"a"—Has a minimum dimension of half a wavelength (in order to propagate a signal) but it may be greater. The "cut-off" frequency depends on dimension "a".

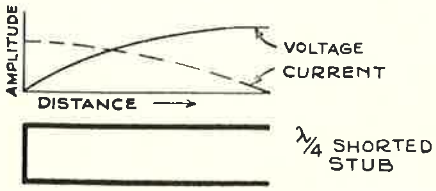
"b"—Is not critical except for voltage breakdown or the possibility of operation in a wrong "mode".

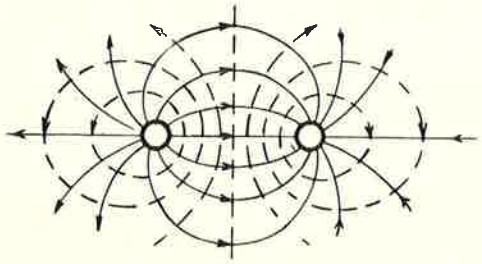


WAVE-GUIDE FIELDS

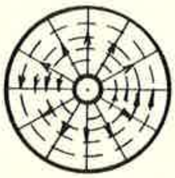
For a simple rectangular wave-guide, the electrostatic lines of force and density of distribution of the E & H fields are shown in the illustration. The electro-magnetic lines of force may be thought of as "whirlpools" in a plane, perpendicular to the electro-static lines of force, traveling down the tube in the direction of propagation. A rectangular wave guide will transmit satisfactorily if the component of the electric field tangent to the side surface is zero at every point on the surface.

A two-wire and a co-axial transmission line are shown with the magnetic and electrostatic fields indicated. A transmission line may be thought of as a guide for magnetic and electrostatic fields.





FIELDS ABOUT TWO-WIRE LINE



FIELDS IN A CO-AXIAL LINE

– – – – H OR MAGNETIC FIELD DUE TO CURRENT IN CONDUCTOR

——— E OR ELECTROSTATIC FIELD DUE TO VOLTAGE BETWEEN CONDUCTORS

TRANSMISSION LINE FIELDS

TYPES OF PROPAGATION

"TM" propagation refers to *transverse-magnetic,* and is applied to the fact there is no magnetic field (H) in the direction of propagation down the tube. The electric field (E) does have a longitudinal component in the direction of propagation. There are numerous modes of operation (or oscillation) depending upon the cross section of the hollow metal tube and how it is excited from the r-f source. Different modes of operation are identified by sub-numerals after the letters "TM."

"TE" propagation refers to *transverse-electrostatic,* and is applied to the fact that the magnetic (H) field is in the direction of propagation and the electric (E) field is transverse. There are numerous modes of operation; identified as mentioned in the paragraph above. It may be of interest to note that "TEo" mode has the lowest "cut off" frequency of any that can be transmitted through a given tube. The fields shown in this booklet are TEo-1.

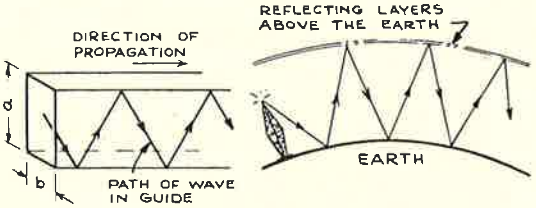
TRANSMISSION OF R-F ENERGY IN WAVE GUIDES

Electro-magnetic fields may be propagated down a hollow metal tube provided:

1. The frequency is high enough.
2. The fields have certain definite distributions.

Wave Guides are essentially ultra-high frequency devices since the frequency must be high before a field can be transmitted (from a practical point of view) through a wave guide. A wave guide has a definite cut-off frequency, as determined by the cross section of the hollow tube. and will not operate at a frequency lower than the cut-off frequency. When the guide is half a wavelength in width (dimension "a") the wave reflects back and forth across the guide making no progress at all (hence cut-off frequency). It may be considered that the wavelength of the r-f energy must be short enough to fit into the cross-sectional dimensions of the wave guide.

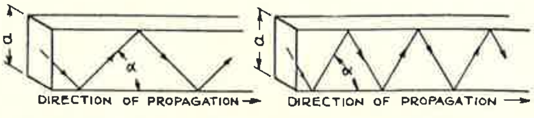
The velocity of propagation (group velocity) of r-f energy in a wave guide is *slower* than the speed in air. This is due to the fact that the wave does not travel straight through the guide, but is reflected from wall to wall. The length of the path that the wave travels is longer than the actual length of the guide. This action is comparable to the reflection of radio waves by ionized layers above the earth as illustrated.



The group velocity is dependent on the frequency and the tube dimensions (dimension "a" in a rectangular guide). For a given size of rectangular guide, the group velocity—

(1) Increases as the wavelength becomes shorter since there are fewer reflections from the wave-guide walls for a given amount of forward travel (see illustration "1") but is always less than the speed of light.

(2) Decreases as the wavelength becomes longer (up to the cut-off frequency) since there are more reflections from the wave-guide walls for a given amount of forward travel (see illustration "2").

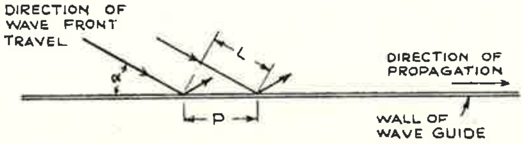


1-SHORTER WAVELENGTHS 2-LONGER WAVELENGTHS

Phase Velocity (apparent speed) is greater than the speed of propagation; or the speed of travel in an unrestricted medium.

$$\text{Apparent speed} = \frac{\text{(true speed)}}{\text{(cos } \alpha \text{ between wall and direction of travel.)}}$$

Thus the apparent speed is greater since the waves are striking the wall at an angle.



In the time required for the wave front to move the distance "L", the point of reflection has moved the greater distance "P". Thus the apparent speed (phase velocity) is greater than the true velocity of propagation.

The wavelength in a hollow wave guide (as measured on a slotted wave guide) is always greater than, or at the limit equal to, the wavelength in air.

Attenuation in a rectangular wave guide of given size varies as follows:

(1) As the wavelength of the signal increases toward the cut-off value, the attenuation increases since there are more reflections resulting in higher losses due to r-f currents in the "skin" of the inside wall.

(2) At a much shorter wavelength, attenuation again increases due to increased "skin effect" at higher frequencies (except in a special case where the electrostatic field does not terminate on the wall of the guide).

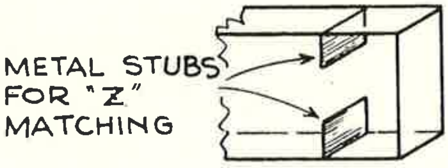
(3) There is an *optimum* frequency of least attenuation; however, this is usually not used because of economy reasons.

The characteristic impedance is different for every mode of *operation.* In a round wave guide the lowest characteristic impedance is about 350 ohms. In a rectangular wave guide the characteristic impedance may be any value as both dimensions are varied. The impedance is directly proportional to the narrow dimension "b" and if the other dimension "a" and the frequency are fixed, the impedance may be any value between approximately 0 and 465 ohms. The impedance approaches zero as the narrow dimension "b" is reduced.

The "Q" of a wave guide is a function of frequency. It is also directly proportional to the ratio of volume to inside area of the guide. "Q" may be of the order of 25,000.

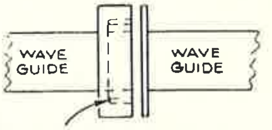
Practical Facts About Wave Guides

1. An electrical quarter-wave length or odd multiples of a quarter-wave length, wave guide *shorted,* reflects an "*opening*" where it joins another wave guide.
2. An electrical half-wave or multiples of a half-wave *shorted* reflects a "*solid wall*" where it joins another wave guide.
3. A quarter-wave section of wave guide inverts the impedance as with ordinary forms of transmission line.
4. A wave guide has a "characteristic impedance" which is usually rather high compared with the impedance of co-axial lines.

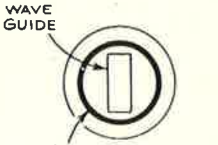


5. Impedance matching may be accomplished by using "shorting stubs." For instance, in matching to a dipole radiator. The stubs constrict the opening; may be used to "filter" out certain modes of operation.

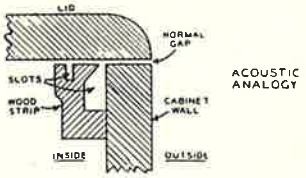
6. Usually where two wave-guide sections are joined in such a fashion, as to be quickly removable, an r-f choke is used.



$\lambda/4$ DEEP SLOT (CLOSED AT BOTTOM) ACTS AS R.F. CHOKE FOR R.F. SIGNAL WHICH MIGHT TRY TO "GET OUT" IN A DIRECTION PERPENDICULAR TO THE AXIS OF WAVE GUIDE.



$\lambda/4$ DEEP SLOT MILLED IN CIRCULAR PIECE OF METAL BRAZED TO WAVE GUIDE.

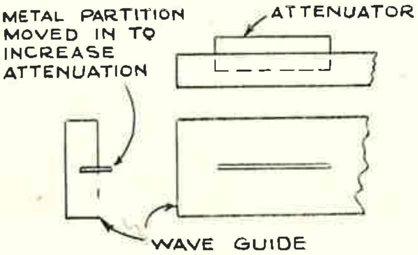


An acoustic analogy (RCA "Tone Guard") is shown above. The slots around the inside of the cabinet act as chokes to prevent undesired high-frequency sound (that is radiated by the vibrating parts of the phonograph pickup) from "getting out" through the normal gap between the lid and the cabinet.

7. For some purposes, usually test, a wave guide may be matched to air by increasing the internal cross section of the guide slightly at the "air" end; normally an r-f choke, as shown above, is also used.

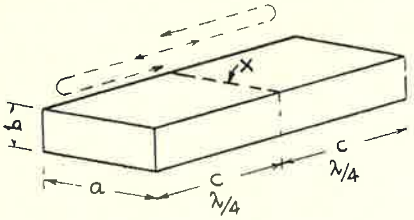
   The end of the wave guide may also be flared outward (increasing the cross section) and used for radiation.

8. A wave-guide attenuator may consist of a movable metal partition as shown below; the attenuation increasing as the partition is moved in.



9. Standing-wave ratios are checked in a manner similar to that used for a co-axial line. A section of wave guide with a narrow slot parallel to the axis of the wave guide (located in the maximum of the electro-static field) is used. A probe with a "crystal detector" or an instrument fuse (1/200A heated to almost the blowing point by d-c) is used to detect the presence of standing waves, as with a slotted co-axial measuring line. Due to the much higher frequency, measurements are considerably more delicate.

10. A section of wave guide may be used as a tuned circuit or as a transformer. In the accompanying illustration, if r-f energy is introduced at "X," reflection will occur at the closed ends. If dimension "C" is correct, the r-f voltage will be reinforced at "X." In this example "C" is a quarter of a (guide) wavelength, however other lengths may be used as long as the reflected voltage arrives back at "X" in phase with the r-f source.

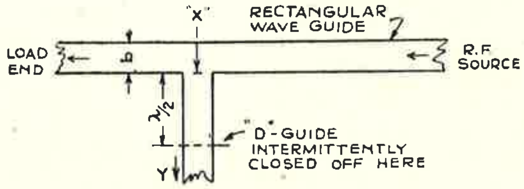


The resonant effect is due to reflections (from the closed ends) setting up standing waves in the guide. The action is similar to that of a "cavity resonator."

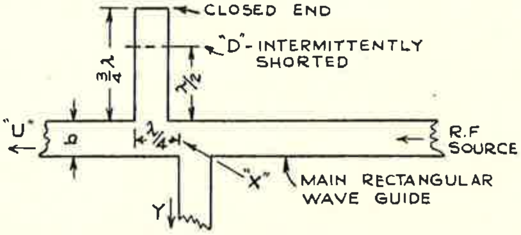
The r-f energy may be introduced inductively, capacitively, or by radiation. Output may be obtained in a similar manner.

11. Sections of open and closed wave guides may be used in switching circuits, etc. In the following examples where a quarter or three-quarters wavelength is indicated, any odd multiple of a quarter wavelength may be used. Where a half wavelength is indicated, any multiple of a half wavelength may be used. (Electrical wavelength in the guide is referred to, not physical length).

In the illustration below an intermittent short at "D" (mechanical or by means of a special tube) is used. A short at "D" will result in effectively a "solid wall" looking in at point "X" when "D" is shorted.



In the following illustration an intermittent short at "D" (mechanical or by means of a special tube) is used.



When "D" is shorted, effectively a "solid wall" will result at the junction of the closed stub to the main wave guide. Paths "Y" and "U" may receive energy.

When "D" is not shorted, effectively a "solid wall" will result looking in at point "X" which is two half waves (a full wave) from the closed end. There will be no transmission in the direction "U" but there will be transmission in direction "Y."

12. A typical wave guide in practical use now, which happened to be a standard size of rectangular tubing:

    Width—about 1.5% greater than a half wavelength (about as small as can be used).

    Thickness—about 44% of the width.

    Wavelength—about 40% longer than in air (standing-wave measurement).

    Loss—about 1/10 db per meter correctly matched).

    Standing Wave Ratio—of as high as 3 to 1 may be tolerated.

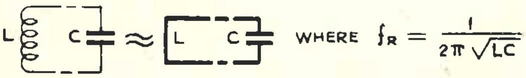
Cavity Resonators

Cavity resonators are tuned resonant circuits for extremely high frequencies where it becomes impossible or impractical to use tuned lines or lumped circuits.

No unique definition of L, C and R can be found in a cavity resonator. A cavity resonator is similar to a wave guide in that electro-magnetic lines of force oscillate back and forth within the cavity in some particular mode, depending upon the shape and method of excitation of the cavity.

UHF cavity resonators may be compared to conventional acoustic resonators. An example is the boomy sound in a room with smooth hard surfaces (good acoustic reflectors). Sound from a source will be reflected from wall to wall with only slight absorption of energy at each reflection. If the frequency of the sound is such as to produce standing waves between two surfaces, or combination of surfaces, the sound is reinforced. The resonant frequency depends on the room dimensions. The "Q" depends on the reflectivity of the walls and other losses.

Developing a Simple Resonant Cavity

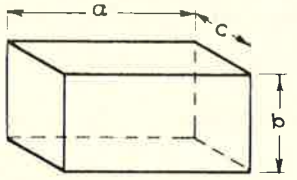


WHERE $f_R = \frac{1}{2\pi\sqrt{LC}}$

If it is desired to increase the resonant frequency, we can parallel the inductances "L," thus making the equivalent "L" quite small. There are limits as to how small "C" may be made practically, so the only thing left to do is to decrease the effective "L" of the circuit in order to tune to a higher frequency. See diagram at right.

As shown at the right above, more and more inductive stubs may be added, thus decreasing the effective "L" and increasing the resonant frequency. If this is carried on to the extreme, a closed chamber or resonant cavity results. (Strictly speaking, it is not proper to talk of the inductance of a cavity resonator.)

Modes of Operation of Cavity Resonators



Consider a rectangular section. The description of the "Mode" would be given in terms of electromagnetic fields and various frequencies. Various frequencies of oscillation (different modes) are possible, because wave energy may be propagated and reflected from various surfaces. There is also the possibility of an oscillation that is a harmonic of the basic wave. Two important points in cavity resonators are (1) how oscillations are forced and (2) how energy is removed. They will effect the mode of operation.

The r-f energy may be introduced to or removed from the resonant cavity, inductively, capacitively, or by radiation.

The energy is in the electrostatic field at one instant and in the magnetic field an instant later, oscillating from one field to the other at the frequency of the applied energy.

In referring to cavity resonators the idea that a conductor is always an "equipotential surface" is untrue; voltages and currents reverse themselves in a space measured in centimeters. An *electrostatic field* can terminate only on electrical charge, hence there must be appropriate distribution of charge on the surface. A *magnetic field* can cease suddenly only on a surface carrying current, hence there must be current flowing in the inside surface of the resonator (only penetrates a very thin skin of the metal surface and cannot be detected on the outside of the resonator).

A general statement, for simple resonators, can be made that it is necessary to have a dimension of an electrical half wave or multiples of a half wave since the electrostatic field is a maximum at the center and minimum at the sides of a simple resonator, otherwise the electrostatic field would be shorted out. (Refer to the data covering a section of wave guide used as a tuned circuit.)

Resonant Frequency "Q" and "Ro"

The resonant frequency can be calculated from the shape of the cavity for very simple types of resonators possessing symmetry.

The "Q" can be determined through knowledge of the rate at which energy is lost. A large "Q" may be obtained when the ratio of volume to surface is large. Approximate values of "Q" may be 28,000, 31,000 and 26,000, for a cube, cylinder, and sphere, respectively, when not loaded.

High "Q" does not necessarily imply high shunt resistance ("Ro") in a resonant cavity.

Forms of Actual Cavity Resonators

Cavity resonators may take various shapes such as, cube, sphere, cylinder, sphere dimpled on top and bottom, cylinder "dented" at one end (with ends forming grids as in an HF tube, for instance), etc.

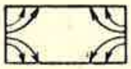
Several possible types of cavity resonators are shown below. The electrostatic lines of force are shown for one possible mode of operation:


CUBE

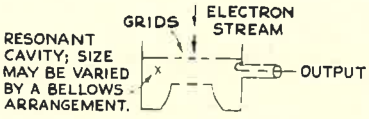

CYLINDER

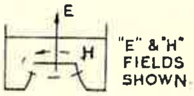

SPHERE

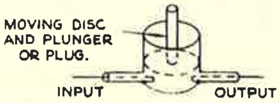

DIMPLED SPHERE

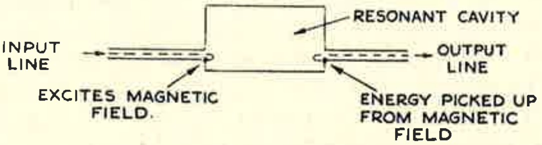

RECTANGULAR


RECTANGULAR









The cavity (box) cannot resonate if it is too small for the wavelength concerned. If the r-f energy is the correct frequency for the cavity, high amplitude centimeter waves (fields) will propagate across and from top to bottom of the cavity.

Tuning Slugs for Cavity Resonators

For the purposes of explanation, a metal sphere is shown in a rectangular resonant cavity, with the "E" lines of force for this particular operation as shown below:


BEFORE


AFTER


AFTER

(1) The slug shortens electrostatic (E) lines of force, hence the capacity is said to *increase.*

(2) The magnetic (H) lines of force are normally weak at the center and are not appreciably affected.

(3) The wavelength *increases* (frequency lower).

If the slug is inserted at one side or the other, the result is as follows:

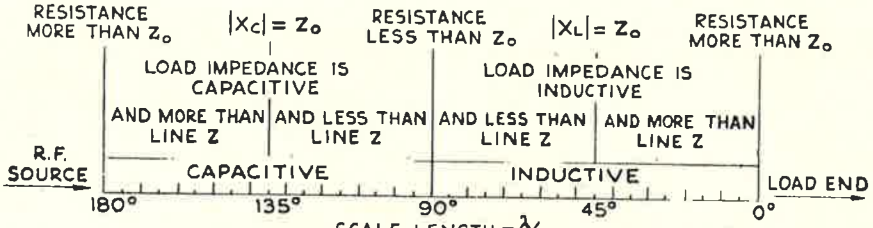
(1) The slug shortens the magnetic lines of force (H), hence the effective inductance is said to *decrease.*

(2) The electrostatic lines of force (E) are normally weak at the side and are not appreciably affected.

(3) The wavelength *decreases* (frequency higher).

Since the two positions of the slug, namely, in the maximum (E) field, and maximum (H) field, change the resonant frequency in opposite directions, it would be expected that a position where no change in wavelength would result might be found.

Appendix on Use of Slotted Line



SCALE LENGTH = $\lambda/2$

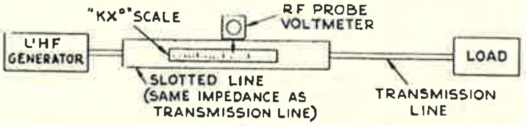
MAKE SCALE LENGTH = ½ WAVE LENGTH AT OPERATING FREQ.

$$\tfrac{1}{2}\text{ WAVE IN INCHES} = \frac{5908}{\text{FREQ. IN MC}}$$

(This length is based on the use of an air-dielectric slotted line with velocity constant practically the same as air.)

By making a few simple measurements on the slotted line (at the desired frequency) it is possible, with the aid of the charts at the end of this booklet, to determine the impedance and phase angle of any load.

The equipment is set-up as shown:



The procedure is as follows:

1. ***Make a scale*** as shown at top of page.
2. ***Locate a "reference point"*** as follows:
   - (a) Short circuit the far end of the transmission line, at the load. The short circuit must be as direct and effective as possible for accurate results.
   - (b) Adjust the generator for correct frequency. Move the probe to a voltage max. point, and adjust generator output for exactly full-scale reading on probe voltmeter.
   - (c) Move the probe along the slotted line and note the position of min. voltage points. Select a min. near the center of the slotted line and locate this point accurately. This min. will be referred to as the "reference point."
3. ***Fasten the prepared scale*** underneath the probe pointer so the 90 degree mark is at the "reference point."
4. Remove the short circuit.
5. Check and if necessary readjust the generator for correct frequency.
6. Move the probe along the scale and accurately locate the point of min. voltage. The reading in degrees on the scale at this min. voltage point is referred to as "KX°."
7. Note the voltmeter reading at the min. voltage point. Determine the ratio of $\frac{\text{min. voltage}}{\text{max. voltage}}$ This is the standing-wave ratio, referred to as "R." (If the rectifier in the r-f voltmeter is non-linear, the ratio may be corrected as described previously.)

Having determined values of "KX°" and "R," it is then necessary to use the correct chart to determine the impedance and phase angle of the load.

There are four charts included in this booklet.

Use #1 if KX° is between 0 and 45°, or 135° and 180°.

Use #2 if KX° is between 45° and 135°.

(Charts #3 and #4 are enlargements of #1 and #2, respectively, for greater accuracy when the standing-wave ratio is between 0.7 and 1.0.)

As an example, if—

R, the standing wave ratio, $\frac{\text{Min. voltage}}{\text{Max. voltage}} = 0.5$

KX°, $= 60°$

Zc, the line impedance, $= 40$ ohms

As KX° is 60°, use chart #2.

Locate R = 0.5 on top of #2 chart and follow this line around until it crosses the KX = 60° line. Mark this point and from it go straight across the chart to find that the ratio of the

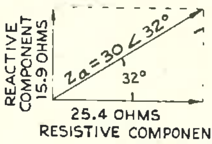
$$\frac{\text{load impedance}}{\text{line impedance}} = 0.75$$

As the line impedance is 40 ohms, the load impedance is 40 x 0.75 or 30 ohms.

From the marked point (where R crosses KX°), drop straight down and note that the angle of the load is 32° (inductive).

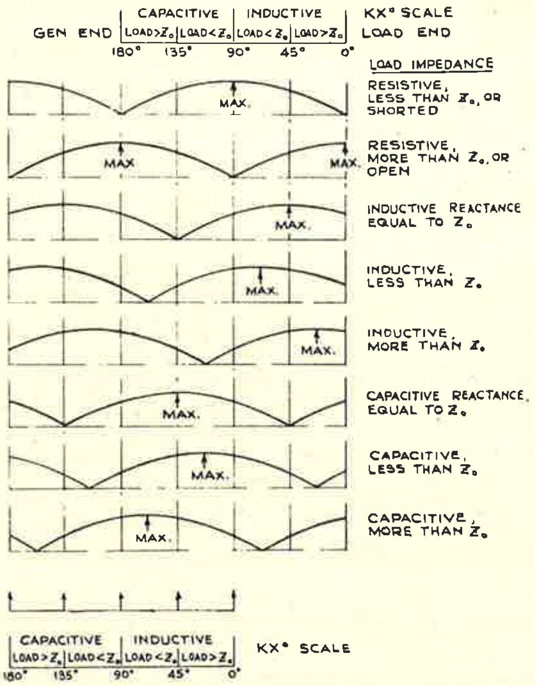
The load impedance (Za) may be expressed in terms of resistive and reactive components:

$$Za = 30 \cos 32° + J\,30 \sin 32°$$
$$= 25.4 + J15.9 \text{ ohms.}$$



EXPRESSED VECTORIALLY

Position of Standing Waves for Various Loads



With the load *short circuited*, any convenient point of *minimum voltage* on the slotted line may be used as the *"reference point."*

To determine KX° by the position of voltage maximum, place the scale so 0° is at a reference point, as shown at top of following chart.

To determine KX° by the position of voltage minimum, place the scale so 90° is at a reference point, as shown at bottom.

The answer is the same in either case.

Owing to the "sharper" indication, it is generally preferable to use a minimum voltage point to determine KX°. However, when using a relatively low frequency, a half-wavelength may be longer than the slotted line and the minimum voltage points may fall beyond the ends of the slotted line. In such cases, it is possible to use a maximum voltage point in determining KX°

The procedure is to short circuit the load and locate the minimum voltage point nearest the load end of the slotted line. Use this as the "reference point." (If the minimum voltage reference point is not close to the load end of the slotted line, change the length of the transmission line so that the minimum voltage reference point is near the load end of the slotted line. Place the prepared scale (shown in previous sketch) so that the *zero-degree mark is at the reference point.*

Remove the short-circuit and locate the first maximum voltage point from the reference mark (on generator side of reference mark). The reading on the degree scale at this maximum voltage point is KX°.

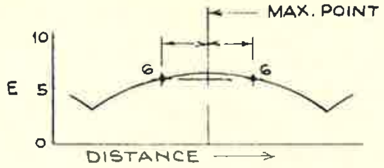
Load Connected Directly at End of Slotted Line

In some applications, the load is connected directly to the end of the slotted line, without using a transmission line. In this case the question of using the minimum or maximum voltage point in determining KX° again depends on the required operating frequency and the length of the slotted line.

With relatively high frequencies, use the first procedure (in which a voltage minimum is used in determining KX°).

With relatively low frequencies, use the point where the load is connected to the slotted line as the "reference point." Place the scale so the zero degree mark is at this reference point. The first maximum voltage point, from the reference point, is KX°.

Note on Obtaining Maximum Voltage Point



As an aid in obtaining accurate location of maximum voltage points, it is suggested that two voltage points of equal magnitude be selected, one on each side of the maximum point, then choose the distance half way between these two points as the maximum point. This same system may be used to determine the location of *minimum* voltage points.

Charts for Use with Slotted Line

Z_C = CHARACTERISTIC IMPEDANCE (SURGE IMP.) OF CONCENTRIC TRANSMISSION LINE OR PARALLEL WIRE LINE.

θ_a = ANGLE OF LOAD IMPEDANCE

Z_a = MAGNITUDE OF LOAD IMPEDANCE (TERM IMP.) AT END OF TRANSMISSION LINE

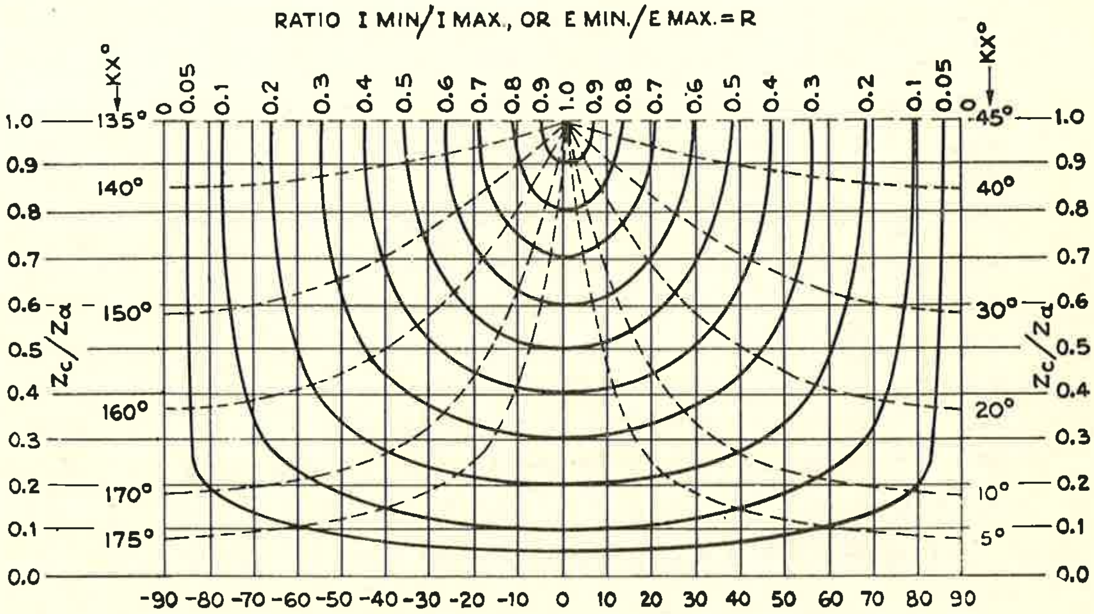


CHART #1.

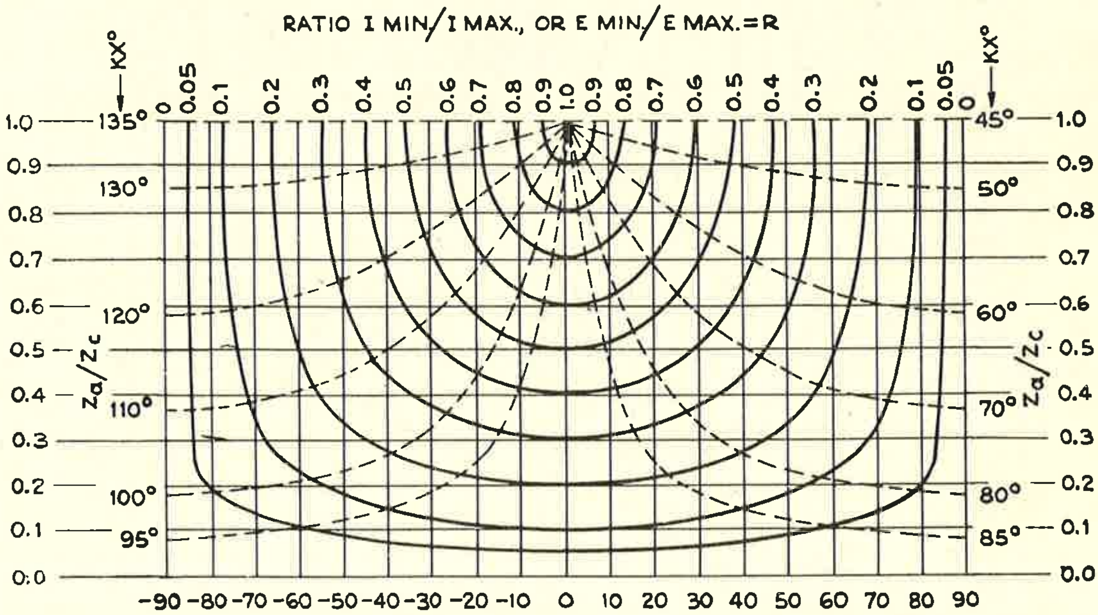


CHART #2

Charts for Use with Slotted Line

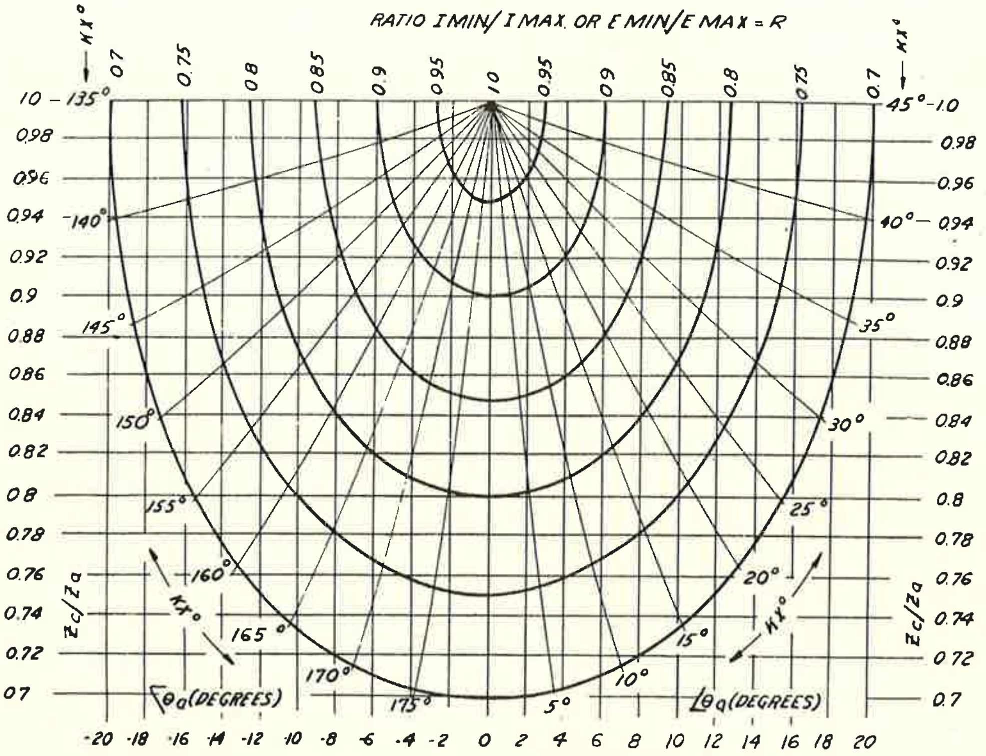


CHART Nº 3 - ENLARGEMENT OF PART OF CHART Nº 1

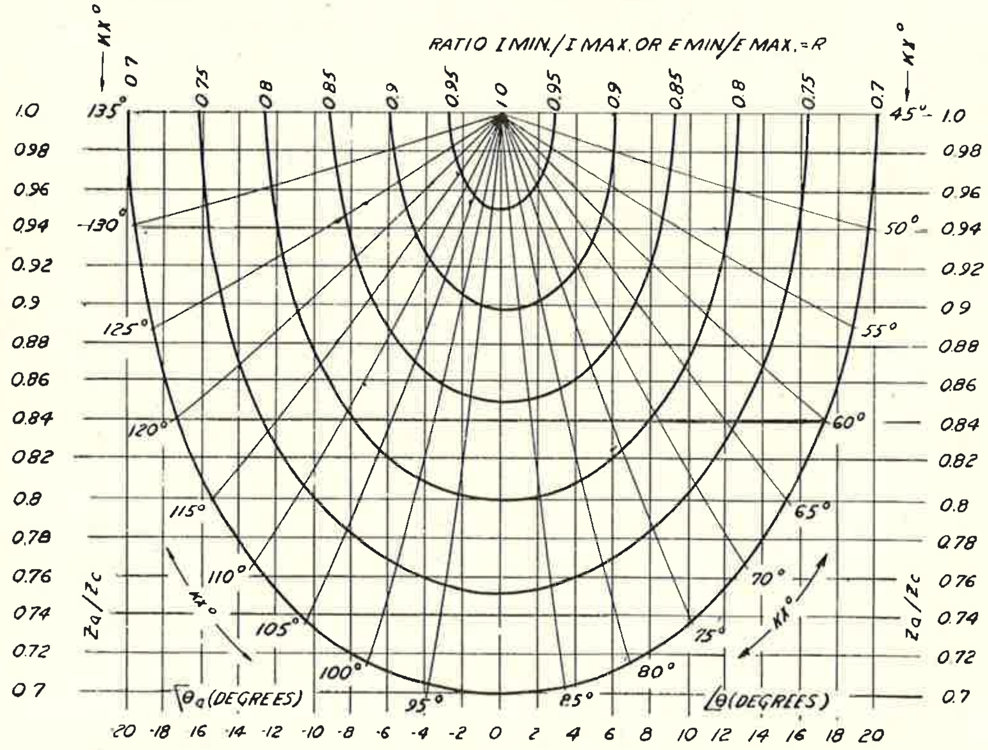
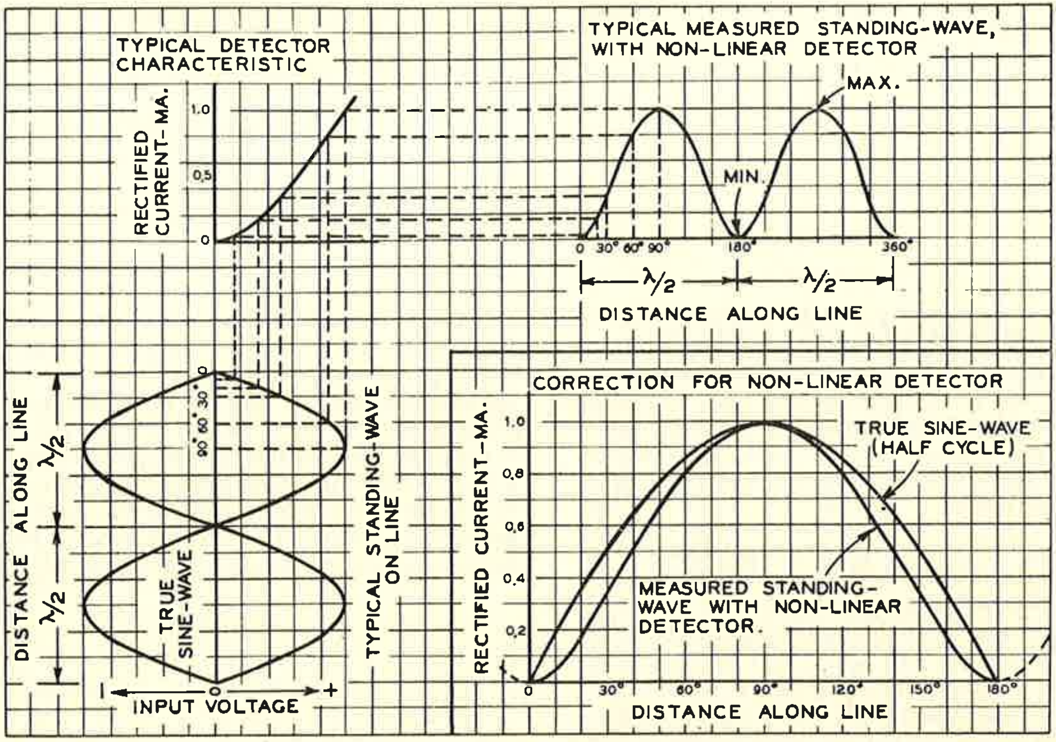
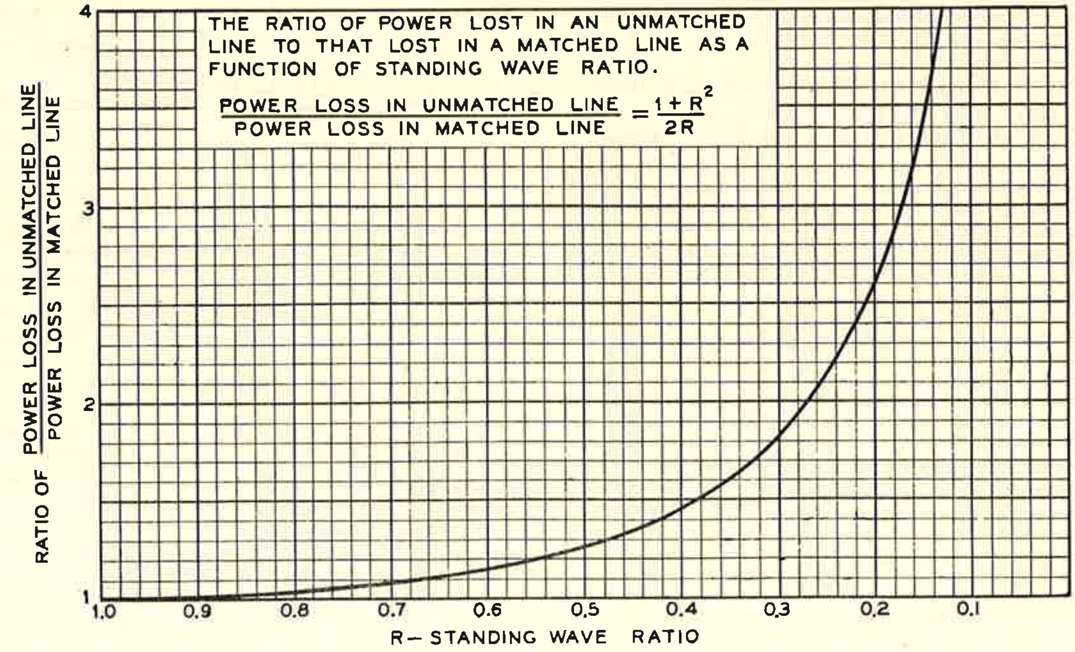


CHART Nº 4 - ENLARGEMENT OF PART OF CHART Nº 2

Effect of Non-Linear Detector



Ratio of Power Loss in Unmatched Line



Intermodulation Measurements on Radiotron Amplifier A515

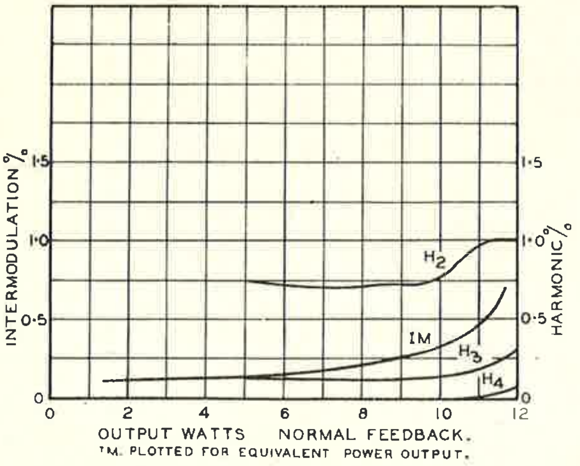
By R. H. ASTON, A.M.I.R.E. (Aust.).

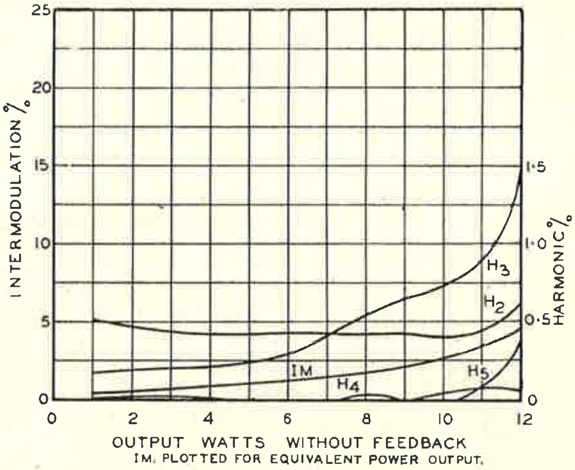
When the Radiotron high fidelity amplifier A515 was described in Radiotronics 128, intermodulation tests were not included because suitable measuring equipment was not available at that time. An Altec Lansing signal generator type T1 401 and intermodulation analyzer T1 402 were subsequently purchased and intermodulation measurements made as published below.

As a guide to the amount of intermodulation which is tolerable, Hilliard states that less than 2 per cent. cannot be detected by ear. It is further claimed that this method of measuring distortion is a very sensitive one and agrees well with the aural impression of distortion. For those readers who require additional information on the subject of intermodulation testing we publish below a bibliography of some of the articles which have appeared in the literature.

Measurements were made of intermodulation and individual harmonics present for various output powers both with and without feedback. These results are published so that some indication of the correlation between intermodulation and harmonic analysis may be obtained.

It should be noted that the intermodulation tests show overload conditions for a lower power output than that indicated by the harmonic measurements of a single frequency input. This arises from the fact that distortion occurs with high peak voltages rather than high powers. When two input signals are applied simultaneously and one of them is four times the amplitude of the other, the smaller signal makes a negligible contribution to the power output, but adds directly to the peak voltage amplitude. As the waveform of the signals encountered in practice is generally of a complex nature, the overload indicated by intermodulation tests is a more practical one.





The only change in the amplifier was the replacing of the output transformer by one designed and manufactured for this amplifier by Fergusons Radio. With the new transformer the amplifier was even more stable than with the Goodmans transformer used previously.

The input signal consisted of the simultaneous application of a 60 c/s sine wave and a 2,000 c/s sine wave. The 2,000 c/s signal was 12 db lower than the 60 c/s input. The combined signals passed through the amplifier and to the intermodulation analyzer where the 60 c/s fundamental was filtered out. The 60 c/s sideband components of the 2,000 c/s, considered as a carrier, were then measured and expressed as a percentage modulation factor.

The output power was calculated from a measurement of the rms voltage, due to the two signals, developed across a 10 ohm resistance.

So that a closer practical comparison can be made between the intermodulation measurements and the harmonic analysis, it is better to calculate the equivalent power output which would result from a single sine wave signal having the same peak amplitude as the sum of the peak voltages of the two signals used for the intermodulation tests. The intermodulation distortion resulting with this equivalent power output can then be compared with the harmonic distortion resulting with the actual power output from a single sine wave.

The equivalent power output may be derived as follows:

let E_1 = 2,000 c/s output voltage
E_2 = 60 c/s output voltage
E_T = measured rms total of E_1 and E_2

$$\text{that is } E_T = \sqrt{E_1^2 + E_2^2}$$

$E_2 = 4E_1$ since E_1 is 12db below E_2

$$\therefore E_T = E_1\sqrt{1 + 16} = E_1\sqrt{17}$$

$$\text{and } E_1 = \frac{E_T}{\sqrt{17}}$$

$$E_2 = \frac{4E_T}{\sqrt{17}}$$

$$E_1 + E_2 = \frac{5}{\sqrt{17}}E_T$$

Equivalent power output =

$$\frac{\left(\frac{5}{\sqrt{17}}E_T\right)^2}{R_L} = \frac{25}{17}\,\frac{E_T^2}{R_L}$$

MEASUREMENTS

Output watts	Equivalent Output watts	Intermodulation per cent. normal	no feedback
1.0	1.47	0.10	0.27
2.0	2.94	0.11	0.37
3.0	4.42	0.13	0.69
4.0	5.88	0.17	1.20
5.0	7.35	0.22	1.65
6.0	8.83	0.27	2.15
7.0	10.30	0.35	2.8
8.0	11.78	0.72	4.3
9.0	13.24	1.75	7.2
10.0	14.71	3.70	11.0
11.0	16.18	7.2	17.5
12.0	17.65	8.8	23.2
13.0	19.12	14.5	29.0
14.0	20.59	22.0	34.0
15.0	22.2	26.0	35.0

◆

HARMONIC ANALYSIS

60 c/s output into 10 ohm resistive load. No feedback.

Output		Harmonics per cent.		
watts	H_2	H_3	H_4	H_5
1.0	0.52	0.17	—	0.025
3.0	0.44	0.22	0.02	—
5.0	0.41	0.24	—	0.025
6.0	0.42	0.29	—	0.02
7.0	0.42	0.40	—	0.04
8.0	0.42	0.54	0.02	0.05
9.0	0.42	0.65	—	0.05
10.0	0.40	0.72	0.04	0.055
11.0	0.43	0.88	0.07	0.07
12.0	0.62	1.48	0.06	0.35
Oscillator	0.66	0.06	—	0.025

60 c/s output into 10 ohm resistive load. Normal feedback.

Output		Harmonics per cent.		
watts	H_2	H_3	H_4	H_5
5.0	0.75	0.11	—	0.025
7.0	0.73	0.13	—	0.025
8.0	0.73	0.14	—	0.025
9.0	0.73	0.14	—	0.025
10.0	0.73	0.16	—	0.025
11.0	1.0	0.18	—	—
12.0	1.0	0.30	0.04	0.05
Oscillator	0.73	0.07	—	0.025

◆

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1. Harries, J. H. O., "Amplitude Distortion", Wireless Engr., Vol. 14 (Feb. 1937), 63.
2. Hilliard, J. K., "Distortion Tests by the Intermodulation Method", Proc. I.R.E., Vol. 29 (Dec. 1941), 614.
3. Scott, H. H., "Audible Audio Distortion", Elec. Vol 18 No. 1 (Jan. 1945), 126.
4. Hilliard, J. K., "Intermodulation Testing", Electronics Vol. 20 No. 7 (July 1946), 123.

Compensation of Frequency Drift

R.C.A. Application Note AN-122 reprinted by courtesy of Radio Corporation of America.

In local oscillator circuits of high-frequency receivers, trouble may be encountered with frequency drift during warm-up. This drift falls into two general classes: The first is that due to the oscillator tube and its socket and is characterized by a short-time frequency drift as the tube warms up; the second is that due to the other components of the oscillator circuit and is characterized by a longer-time drift as the chassis comes up to temperature. Both types of drift generally tend to lower the oscillator frequency.

This Note describes a method for substantially compensating the drift due to the warm-up of the tube and socket, and illustrates the application of the method in a receiver. Compensation of the drift attributable to other circuit components is not discussed.

Short-Time Drift Considerations

The construction of a miniature tube lends itself well to the compensation of that part of the frequency drift due to the tube and socket. Heat is conducted directly from the tube elements through the base pins to the tube socket contacts. Because the socket dielectric material is a poor heat conductor, little heat is lost to the chassis. Consequently, the changes in temperature at the socket terminals are closely related both in time and in relative value to the temperature changes of the tube elements to which the terminals are connected. A capacitor with a negative temperature coefficient mounted directly across two appropriate socket terminals will have a very appreciable temperature rise and, therefore, a small value of capacitance can provide substantial compensation for frequency drift due to the tube and socket.

Tests Without Frequency-Drift Compensation

A receiver using a 6BE6 as a local oscillator was used in the following tests. The oscillator was operated with its fundamental frequency above the signal in a conventional cathode-feedback circuit as shown in Fig. 1. The oscillator tube socket was a phenolic wafer socket. Tests were made at an oscillator frequency of 110.7 Mc. Because the long-time drift was not under consideration, the tests were made with the receiver chassis out of the cabinet. The observed results follow. For purposes of comparison, the frequency drift values reported are those read from the curves at the ten-minute test point.

(1) The warm-up drift of the unmodified receiver was about 200 kc as shown by curve 1 of Fig. 2.

(2) When a cold 6BE6 was substituted for a hot tube in the set after it was thoroughly warmed up, the measured drift was about 140 kc as shown by curve 2 of Fig. 2.

(3) After a mica-filled rubber socket was substituted for the phenolic socket, the drift was again measured with a cold 6BE6 inserted in a hot chassis. The drift was reduced to about 100 kc as shown by curve 3 of Fig. 2.

(4) The changes in frequency produced by connecting small capacitors between the grid-socket terminal and the chassis were measured to determine the rate of change of frequency with capacitance. In this manner it was determined that a 100-kc frequency drift corresponded to a change in effective tank capacitance of 0.05 $\mu\mu$f.

Several conclusions may be drawn from the data:

(1) Tests involving the insertion of a cold tube in a hot socket do not necessarily give the drift due to the tube alone. That this is true, is shown by the difference between curves 2 and 3 of Fig. 2.

(2) Phenolic sockets may contribute a significant part of the frequency drift.

(3) The mica-filled rubber socket tested gave reduced drift.

(4) A capacitor having a negative temperature coefficient and connected across the socket terminals will receive heat from the tube in the same manner as the socket. On the basis that a compensation of 100 kc is required, a 3.3-$\mu\mu$f capacitor with a negative temperature coefficient of 0.00075 $\mu\mu$f/$\mu\mu$f/°C across the tank circuit will suffice if its temperature can be raised 20 degrees C during warm-up.

Tests With Frequency-Drift Compensation

The actual compensation was accomplished by soldering the compensating capacitor across the terminals of a mica-filled rubber socket as shown in Fig. 3. The capacitor is connected between socket pins 3 and 6. This is equivalent electrically to connecting the capacitor between the cathode tap on the coil and ground, as may be seen by reference to Fig. 1.

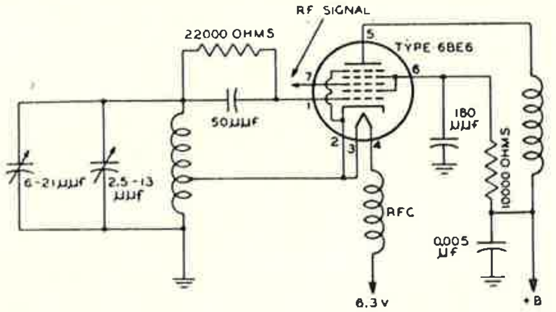

Fig. 1 - Schematic diagram of oscillator circuit.

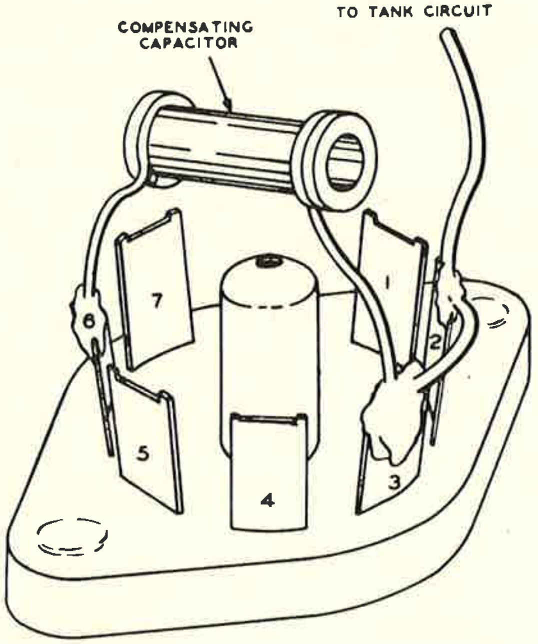

Fig. 3 - Location of the compensating capacitor across the terminals of the socket.

The capacitor leads should be as short as is feasible. This connection was desirable for several reasons. Since the capacitor is across only part of the tank circuit, it can have a larger, more convenient value. However, its effect can be made the same as that of a smaller capacitor across the whole tank circuit. Further, the capacitor receives more heat by being connected to the heater terminal instead of the oscillator-grid terminal because the heater terminal runs hotter. An 8.2-μμf capacitor with a negative temperature coefficient of 0.00075 μμf/μμf°C was selected on a trial-and-error basis. The compensating capacitor effectively added less than two μμf across the terminals of the tuning capacitor, a value within the adjustment range of the trimmer capacitor. The frequency drift obtained with this compensating capacitor was about 50 kc as shown in curve 1, Fig. 4 for a cold tube plugged into a hot socket. Curve 3, Fig. 2 gives results before compensation.

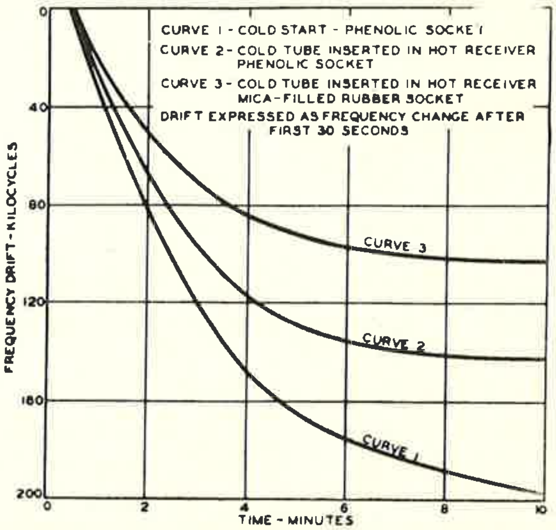

Fig. 2 - Frequency drift of receiver before compensation

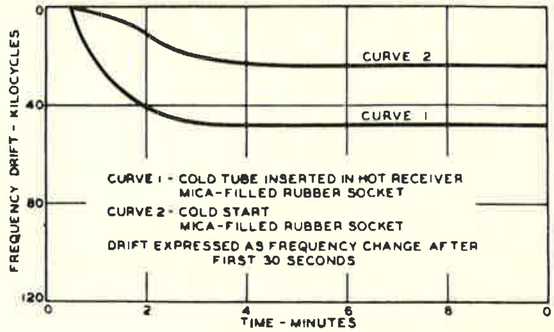

Fig. 4 - Frequency drift of receiver after compensation.

The overall frequency drift from a cold start obtained with this compensation was 25 kc as shown in curve 2, Fig. 4. The frequency drift after compensation is greater when a cold tube is inserted in a hot chassis than when both the tube and chassis have a cold start. This difference in frequency drift occurs because the compensating capacitor is not completely cooled by insertion of the cold tube into the hot socket.

Receiver Microphonics Caused by Heater-Cathode Capacitance Variations

R.C.A. Application Note AN-123 reprinted by courtesy of Radio Corporation of America.

One of the sources of microphonics in conventional superheterodyne receivers is small variations in the heater-cathode capacitance of the oscillator tube which are produced by vibration and which occur at an audio frequency. When the vibration is caused by the loudspeaker, acoustic feedback may take place and an objectionable howl will result. In an FM receiver, microphonics can be readily detected as frequency modulation of the oscillator. In an AM system, microphonics are evident when the receiver is slightly off tune. Under this condition, the centre frequency is located on the side of the intermediate-frequency response curve and frequency variations cause changes in amplitude. In the 500-to-1600-kilocycle standard broadcast band, microphonics caused by variations of the heater-cathode capacitance are negligible and, therefore, ordinarily require no special design precautions.

This Note describes design and construction precautions for minimizing microphonics by eliminating the effects of heater-cathode capacitance variations in an oscillator tube of a receiver operating in the FM, television, or other short-wave bands.

The heater-cathode capacitance for tube types 6SB7-Y and 6BE6 is approximately three micromicrofarads. If the oscillator of an FM receiver operating at 100 megacycles has a total tank capacitance of 25 micromicrofarads, a change as small as 0.00375 micromicrofarads causes a frequency deviation of 7500 cycles. If this deviation is recurring, it corresponds to a five-per cent. modulation of a 150,000-cycle FM bandwidth. If the deviation takes place at an audio-frequency rate, such a capacitance change, although small, is of sufficient magnitude to cause a microphonic howl in a high-frequency receiver.

A typical 100-megacycle converter circuit employing a Hartley-type oscillator is shown in Fig. 1. In a pentagrid converter such as the 6SB7-Y or the 6BE6, grid No. 2 (screen) functions as the oscillator anode and is at rf ground. The cathode is connected to a tap on the tank coil and the heater-cathode capacitance is, therefore, across part of the tank circuit. If there is movement of the heater in the cathode sleeve at an audio frequency due to vibration, the heater-cathode capacitance will vary at an audio frequency and will in turn modulate the frequency of the oscillator. Microphonics can be minimized in this circuit if an arrangement is utilized which limits the effect of heater-cathode capacitance variations on the oscillator. This effect can be eliminated almost entirely by tying one heater lead to the cathode. The only other circuit change required is the insertion of an rf choke in the second heater lead to prevent the heater from loading the oscillator. See Fig. 2. A ground return for the heater is provided through the cathode tap on the oscillator.

Suggestions and considerations for utilizing this method in various types of receivers follow.

a. In single-band ac FM receivers, use circuit of Fig. 2. To prevent the heater from loading the oscillator, the inductance of L_1 should be large compared with the inductance of the tapped section of the oscillator coil. In the FM band, a value of one microhenry for L_1 is adequate.

b. In short-wave multi-band (AM/FM) ac receivers, use circuit of Fig. 2. The value of L_1 should be sufficient to prevent the heater from loading the oscillator at the lowest frequency band.

c. In ac/dc receivers in which the oscillator or converter tube is at the ground end of the heater string, use circuit of Fig. 2. To prevent the heater from loading the oscillator, the inductance of L_1 should be large compared with the inductance of the tapped section of the oscillator coil.

d. In ac/dc receivers in which the oscillator or converter tube is not at the ground end of the heater string, use circuit of Fig. 3. Representative values for an ac/dc receiver operating in the FM band are:

L_1, L_2 = 1 microhenry
C_1 = 68 micromicrofarads

In this circuit, the value of C_1 is selected so that the heater is effectively short-circuited to the cathode throughout the FM band.

Because the oscillator coil in the AM broadcast band is often made of very fine wire not capable of carrying heater current, it may be desirable to use a switching arrangement which ties the heater to cathode when the receiver is operating in the FM band but which disconnects the heater and cathode in the AM band. A typical switching arrangement is given in Fig. 4.

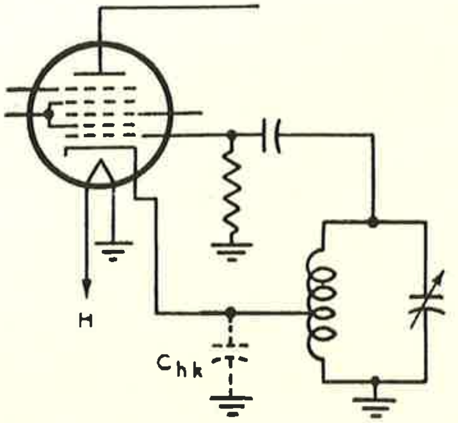

Fig. 1

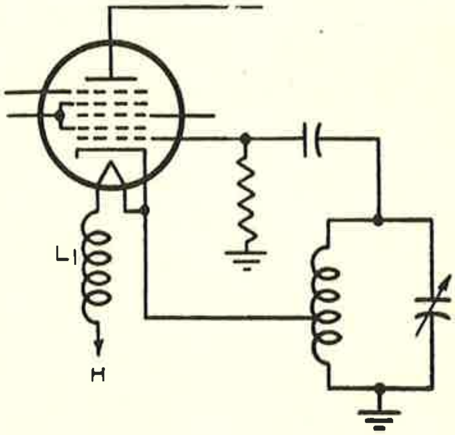

Fig. 2

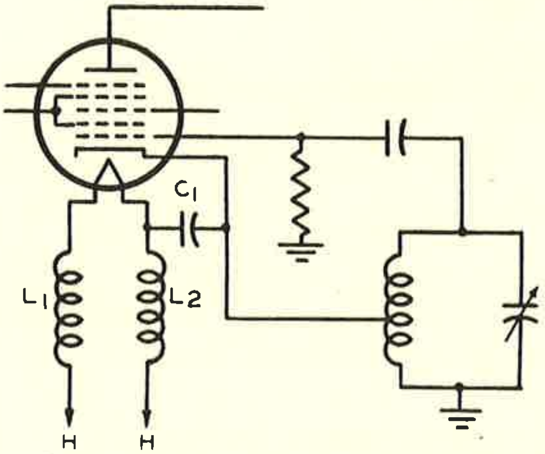

Fig. 3

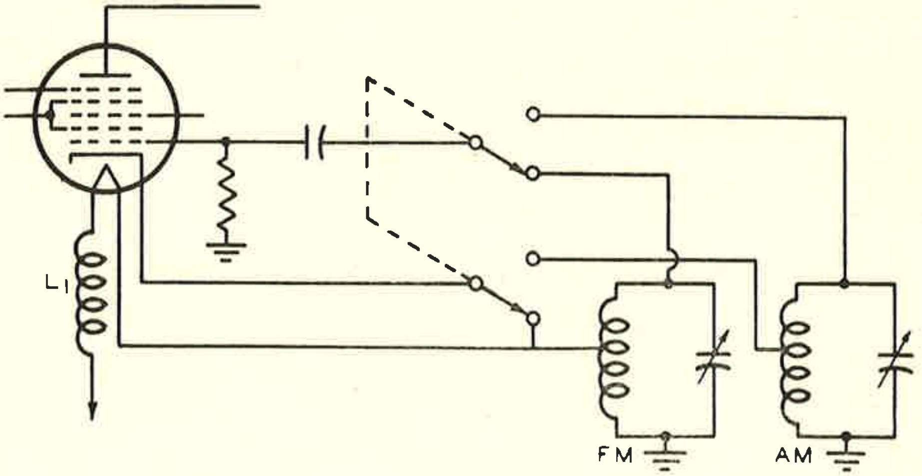

Fig. 4

Some precautions to observe in the design and construction of these circuits follow.

1. When the circuit of either Fig. 2 or Fig. 4 is used, the choke L_1 and the high-frequency oscillator coil have to carry heater current and should be designed accordingly.

2. The resistance of the chokes and coils in series with the heater of the tube should be kept low in order to avoid operating the tube at reduced heater voltage. If the heaters are in parallel, it is generally not difficult to keep the voltage drop across the choke below 0.05 volts. If the heaters are in series, the voltage drop is not important but the heat dissipation in the choke and coil windings should be kept to a safe value.

3. The coils employed for one band should not produce resonant effects in other bands of the receiver.

4. Precautions against microphonics caused by variations in heater-cathode capacitance should be incorporated in the early stages of design. Later adoption may result in tracking difficulties or loss of sensitivity.

New R.C.A. Releases

Radiotron types OA2, OB2 — are miniature types of cold-cathode, glow discharge tubes. They are intended for use as voltage regulators in applications where it is necessary to maintain across a load a d.c. voltage substantially independent of load current and moderate line-voltage variations. The type OA2 maintains a d.c. operating voltage of approximately 150 volts, while the type OB2 operates at 108 volts.

Because of their miniature size, these tubes can be utilized to provide regulated B and C voltages in compact equipment where space previously precluded use of the larger voltage regulator tubes.

Like other glow-discharge tubes types OA2 and OB2 may also be used for spark-over protection.

Radiotron type 6BH6 is a miniature type of r-f pentode featuring high transconductance, a sharp cut-off characteristic, and a 150 mA heater. It is particularly useful in ac/dc receivers and mobile equipment where low heater drain is important.

Radiotron type 5651 — is a miniature, voltage-reference tube of the cold-cathode, glow discharge type designed for extreme voltage stability. Its exceptional stability, not only initially but also throughout life, makes the type 5651 particularly useful as the voltage-reference tube in d.c. power supplies incorporating electronic voltage regulation.

The voltage stability of the type 5651 is such that voltage fluctuation at any current within the operating-current range of 1.5 to 3.5 mA is less than 0.1 volt. Type 5651 maintains an operating voltage of approximately 87 volts over its operating-current range, and has an operating characteristic which is essentially independent of ambient temperature.

Radiotron type 5671 is a forced air cooled power triode designed to meet the exacting requirements of high power broadcast stations. In unmodulated class C telegraph service it has a maximum plate dissipation of 25 kilowatts. It can be used at maximum rated plate voltage and plate input at frequencies as high as 10 Mc/s.