SECTION 3

SPECIALIZED BROADCAST RADIO ENGINEERING

TRANSMISSION SYSTEMS FOR U.H.F. BROADCASTING

313D

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TECHNICAL ASSIGNMENT

TRANSMISSION SYSTEMS FOR U.H.F. BROADCASTING

TRANSMISSION SYSTEMS FOR U.H.F. BROADCASTING

GENERAL DISCUSSION: The ever-increasing expansion of radio applications have necessitated, from time to time, the reallocation of radio channels and also the opening up of new portions of the spectrum to meet the increased demands for facilities. The trend has obviously been to the higher frequencies, since the original trend was to the lower frequencies, and these have een used up down to the lowest feasible, namely, about 10 k.c.

It will be recalled from an earlier assignment that with the advent of broadcasting, the amateurs were assigned frequencies in that portion of the spectrum now known as the short waves (3 m.c. to 23 m.c.). They found that it was possible to transmit over surprisingly long distances, and investigation indicated that this was owing to reflections of the sky wave from the ionosphere. As a result, commercial use began to be made of these frequencies for international broadcasting, radio telephone, etc.

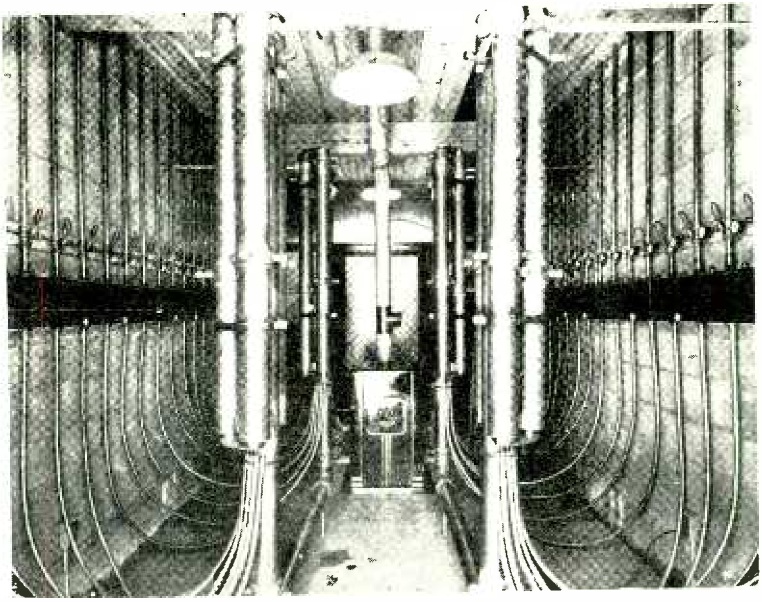
With the advent of television and F.M., which require rather wide bands for each channel (particularly television), it became necessary to employ still higher frequencies up to 100 m.c. and beyond in order that such wide bands be a suitably small fraction of the carrier frequency. These are known as ultra-high frequencies, although today frequencies as high as 30,000 m.c. are being employed! At present these are called by various names, such as u.h.f. (very high frequencies), hyper frequencies, and microwaves.

Broadcasting today, other than in the standard and international short-wave broadcast bands is confined mainly to frequencies around 90 m.c. in the form of F.M. Additional uses of ultra-high frequencies in the broadcast field are for relay transmitters that "beam" the signal output of the studio to the transmitter by means of radio rather than by wire lines. U.H.F. antennas for the production of beams are reasonably compact and hence not too expensive. The signal is picked up at the transmitter by a u.h.f. receiver, and the audio output used to modulate the carrier, after which it is of course broadcast. Thus the ultra-high frequencies are employed directly for F.M. broadcasting and also for point-to-point relaying in place of telephone lines. Radio relaying appears to have a promising future in this field after the war.

At ultra-high frequencies broadcasting is limited to distances that do not very greatly exceed line-of-sight transmission, at least as regards the primary service area. This might appear to be a serious limitation. However, it will be recalled that in the standard broadcast range the primary service area is limited to the distance where the ground wave still exceeds the reflected sky wave in intensity, and that this distance is at most on the order of 100 to 150 miles. Line-of-sight transmission, on the other hand, is about 50 miles for suitably elevated antennas, as is the case at least for the transmitting antenna. Thus the coverage in the case of the standard broadcast band is not so very greatly in excess of that for frequencies around 90 m.c.

Another factor, particularly pertinent in the case of F.M. broadcasting, is that the F.M. receiver tends to discriminate against the weaker of two signals on the same frequency, particularly if the stronger is 6 db. or more above the weaker. This effect is not present in the case of amplitude-modulated signals. As a result F.M. transmitters operating on the same frequency can usually be spaced closer together geographically than A.M. transmitters, and thus more stations accomodated in the same portion of the radio spectrum. Thus the use of u.h.f. for F.M. broadcasting not only permits the necessarily large number of side-bands to be accommodated, but also permits the use of more stations on each frequency, so that ultimately as comprehensive a network on F.M. may be expected as is to be found at present in the standard broadcast band.

Transmission lines are of great importance in these applications. Actually, it is evident from previous assignments that transmission lines are important in all branches of radio. For example, it has been shown how every antenna is essentially a transmission line from which radiation takes place, and usually antennas are fed from the transmitter through transmission lines. But at ultra-high frequencies transmission lines can be employed for many other purposes, such as resonant tank circuits of very high Q, as matching networks, and as means for cancelling out the reactances of loads, etc., in the form of matching stubs. In this lesson will be discussed the underlying theory of transmission lines, with particular reference to the application of this theory to the above-mentioned uses of the lines at U.H.F.



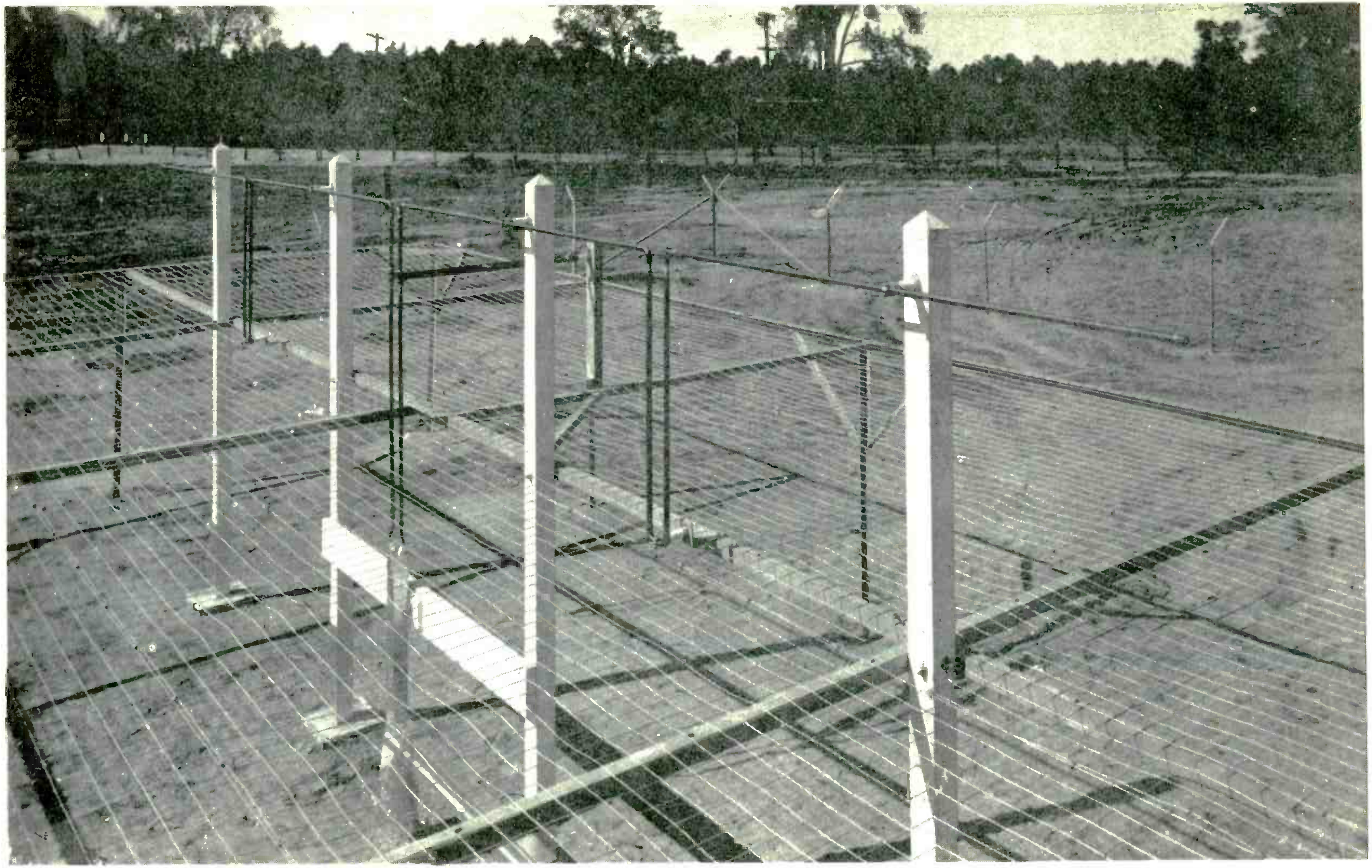
Matching and Phasing Networks, Coupling House, 50 KW. F.M. Transmitter, Paxton, Mass.

Figure 1

TRANSMISSION LINES

Although the study of transmission lines has always been of major importance to the *telephone engineer* it is a subject which in recent years has become of increasing importance to *radio engineers*, especially to those working with ultra-high frequency circuits as, for example, in the television channels, in F.M. broadcasting and aeronautical radio navigation apparatus.

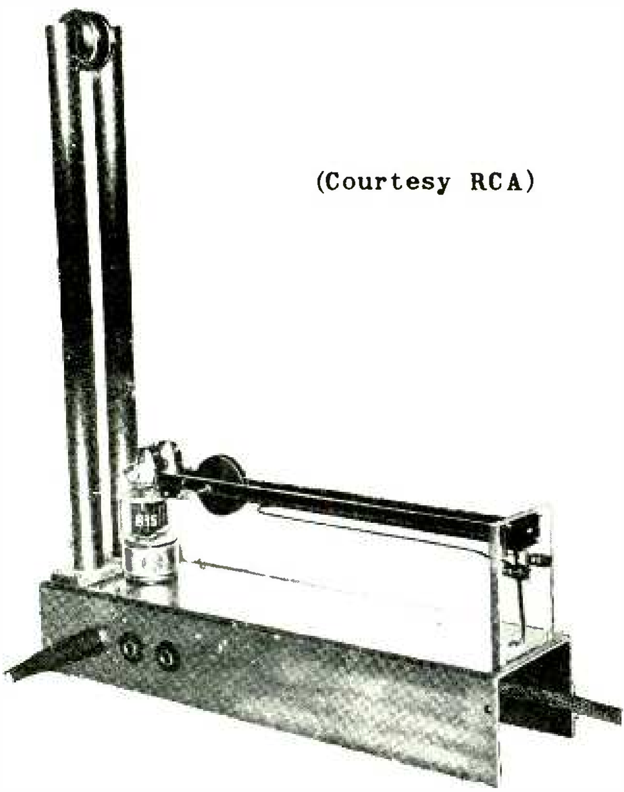
Transmission lines (although they sometimes are not thought of as such), are used in one form or another throughout the entire circuit of all transmitters and receivers. They are the



B 13 Fig. 2. Photograph of Antenna Installation lines which couple one amplifier stage to the next in cascade; transfer the power from output amplifier to the antenna; transfer energy from antenna to input terminals at the receiver--in short, a transmission line is simply a pair of conductors for transferring electrical energy from one point in a circuit to another.

Transmission lines are also used for a number of purposes in addition to that of transmitting energy from point to point. At the ultra-high frequencies (above 30 MC/s), where the LC circuit constants required become extremely small, the distributed capacity and inductance of a comparatively short two-wire line is quite sufficient to tune to the operating frequency.

The single-tube oscillator shown in Figure 3 is a typical arrangement of such a circuit. An RCA 815 tube, with an output of 25 to 35 watts at 120 MC/s is used with tuned grid and plate circuits consisting of tuned lines. These tuned lines consist of copper tubing, the vertical two conductor line being the tuned grid circuit and the horizontal tuned line that of the plate circuit. The resonant frequency of these two circuits is adjusted by the "opposing disc" condensers shown connected to the ends of the lines.



(Courtesy RCA)

Figure 3

The aeronautical marker beacon transmitter operating in an ultra-high frequency channel is a typical example of the use of transmission lines as circuit elements. Throughout the entire transmitter straight two-wire conductors may be used as tank circuits, filter circuits, coupling devices, and as impedance and voltage transformers. Tuned circuits consisting of lines are actually preferable to other types at the ultra-high frequencies because of the extremely high Q that can be obtained. (A circuit Q of 10,000 is easily obtained.)

LINE PROPAGATION: At the lower radio frequencies the engineer is accustomed to thinking of the transmission line as merely a conductor with negligible losses, which will transfer an impressed voltage immediately down to the receiving end. This is a concept which proves quite satisfactory for d.c. or low radio frequency circuits. However at the higher frequencies, especially where the line is of considerable length, this idea becomes decidedly in error and a more general concept must be taken for the study of transmission lines.

The velocity of an electrical impulse along the ordinary transmission line is nearly that of light (3×10^8 meters/sec.). It is considerably less for lines having high losses and particularly for those having an insulating medium whose dielectric constant or permeability is markedly greater than that of free space, e.g., cables.

To show the effect of line propagation at the ultra-high frequencies, a 150 MC/s oscillator feeding a two-wire open-circuited line may be considered as an example. In Figure 4 the

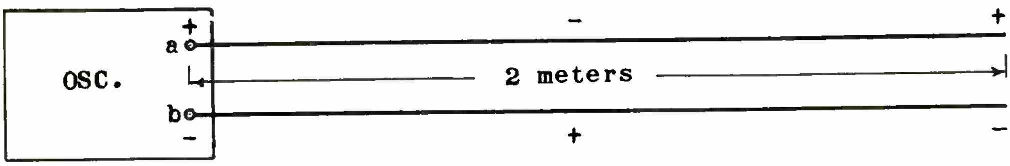


Figure 4

oscillator is shown connected to the two-wire line which is about 2 meters in length. Bearing in mind that the voltage output of the oscillator is in the form of a sine wave, when the voltage is a maximum positive at terminal "a" (and a maximum negative at "b"), this voltage starts propagating energy down the line at approximately the speed of light (3×10^8 meters per second).

But, since the oscillator has a frequency of 150 MC/s the voltage goes through a complete cycle every 1/150 millionth of a second and alternates in polarity every 1/300 millionth of a second. At the end of 1/300 millionth of a second the voltage polarity at terminals a-b will be reversed although *the original voltage has traveled only one meter down the line.* Here is a situation where the terminal voltage at a-b is of one polarity while half-way down the line is a voltage just the opposite in polarity.

Now observe the voltage for another 1/300 millionth of a second. The terminal voltage at a-b has by this time again reversed in polarity and the other voltages have traveled another meter down the line. Terminal "a" is positive, but half-way down the line the voltage is negative and at the end of the line the voltage is positive again. The condition at that time is shown schematically in Figure 5. It will be seen that the voltage across the line, between the points shown, varies sinusoidally and *is actually a sine wave of voltage in motion*, each portion of this voltage wave propagating down the line at approximately 3×10^8 meters/second.

REFLECTION: Realizing, then, that a sine wave of voltage, is actually travelling down the length of the line, (its position at some later time is shown, for example, by the dotted line of

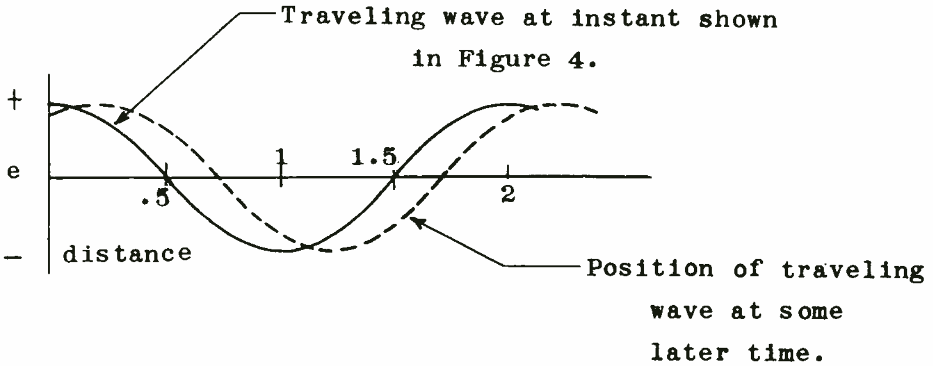


Figure 5

of Figure 5) it now becomes extremely interesting to observe what happens as each portion of the wave arrives at the end of the line. Since the passing of the voltage wave down the line is accompanied by a corresponding flow of current, power is being transferred along the line and if the line is not properly terminated by a load resistance which will absorb all of this power, reflections will occur.

In other words, if the travelling wave is not completely absorbed at the receiving end by the correct load impedance, the portion of this energy not dissipated in the load will then start back toward the generator, again in the form of a travelling wave. This condition is shown in Figure 6, where the end of the

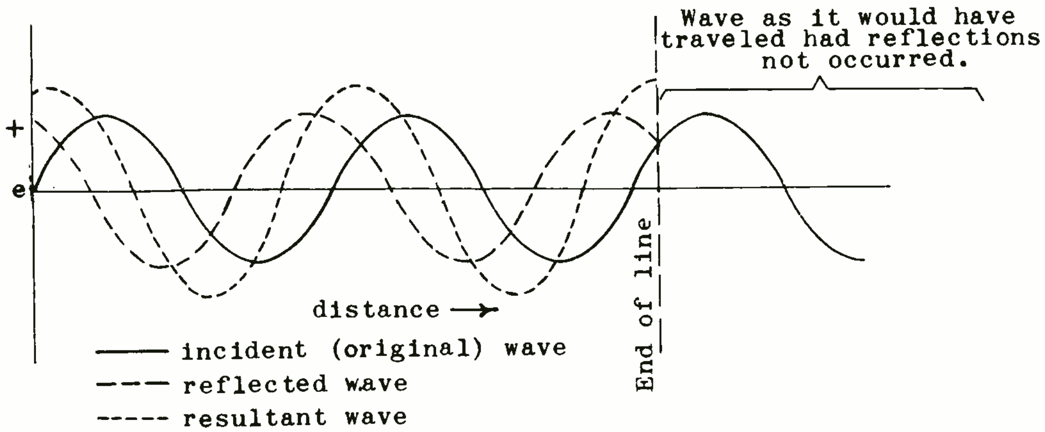


Figure 6

line is open circuited and all of the original travelling wave is reflected back toward the generator. It will be noted that the resultant of these two waves is a maximum at the end of the line. Furthermore, it can be shown that as the incident wave travels down the line, the sum of the incident and reflected waves will give rise to a voltage at the end of the line that, while varying sinusoidally with respect to time, has a peak value greater than that of neighboring points of the line. However, other such maximum peak values will be found at points distant by a half-wavelength, or integer multiples of a half-wavelength, from the end of the line. The resulting sinusoidal distribution of voltage or current along the line, with maxima and minima at certain points along the line, is known and referred to as a *standing wave*, even though it is produced by an incident wave travelling continuously down the line.

RESONANT LINES: One of the outstanding facts about the operation of a transmission line, which is long in comparison with the operating wavelength, is the manner in which the voltage and current distribute themselves along a line which is not properly terminated.

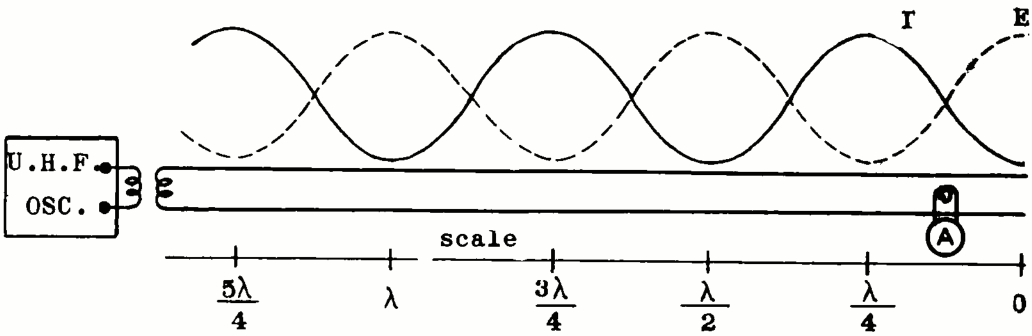


Figure 7

This effect may be readily seen by experiment, as shown in Fig. 7, where a two-wire transmission line is inductively coupled to an U.H.F. oscillator. The "receiving end" is left open circuited. In this particular case the line is slightly more than 5/4 wavelength as shown by the scale. An ammeter is coupled to the line by a coil consisting of several turns of wire. This coil is placed between the two wires and moved up and down the length of the line. The meter, which reads a value proportional to the current flow at any point on the wire, will be observed to change in reading as shown by curve I. That is, at the end of the line the current will read a minimum.

As the coil is moved from the receiving end the meter reading will increase and reach a maximum at λ/4 from the receiver. As the coil is moved still farther from the receiving end the current will go through successive minima at distances from the receiver corresponding to an *even number* of quarter-wavelengths and through maxima at distances corresponding to *odd numbers* of quarter-wavelengths, *always measured from the receiving end.*

With a voltage indicating device (such as a neon tube or lamp) it also can be shown that the voltage across the line goes through maxima and minima, as shown by curve E. The voltage distribution has its maximum values where the current is minimum and its minimum values where the current is maximum.

The graph shows the relative *amplitudes* of the voltage and current in both wires at various distances from the receiver. However it must be remembered that the meter will not show phase relationships. Actually the current and voltage both reverse in phase as they pass through their respective minimum values. Also, the current (or voltage) at any point in one wire of the line is 180° out of phase with that in the second wire at the same point.

Such distributions of E and I are referred to as *standing waves* of voltage and current. The ratio of the maximum to minimum values depends on the resistance per unit length of line, being greater with lower resistance.

If the receiving end of the two-wire line is *short circuited,* as shown in Figure 8, a different current and voltage distribution is observed. In this case, with the aid of an ammeter,

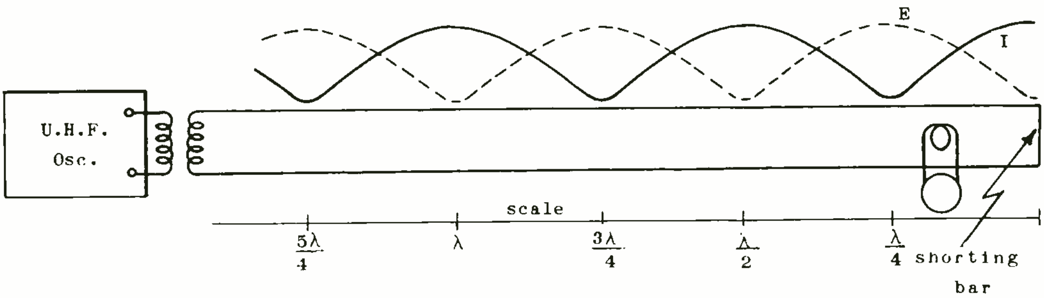


Figure 8

it will be noted that the current goes through maxima at distances an even number of quarter-wavelengths from the receiving end and minima at distances an odd number of quarter-wavelengths from the receiving end. As before, the voltage has its minimum values where the current is maximum and its maximum values where the current is minimum.

To obtain ammeter readings that are proportional to the current in the line, care must be taken to keep the ammeter "pick up" coil at the same distance from the two wires as it is moved along the line. Otherwise, faulty readings might be made which would not show the correct distribution of the current.

One method which may be used with bare wire lines to eliminate this possibility of error is shown in Figure 9 where the meter is connected across a portion of one of the lines (several inches in length) acting as a shunt. Another useful and inexpensive device that may be employed in place of a meter is a small neon lamp. If one terminal of the lamp is moved along the line in contact with one of the wires, as shown in Figure 9,

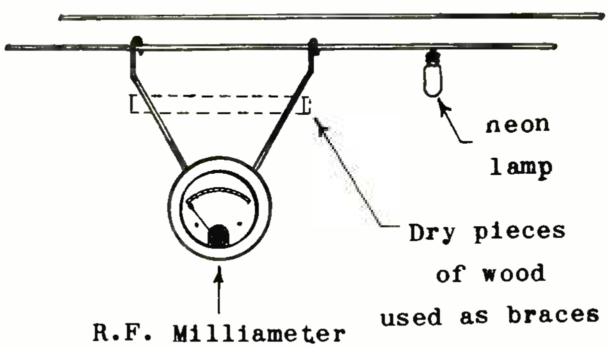


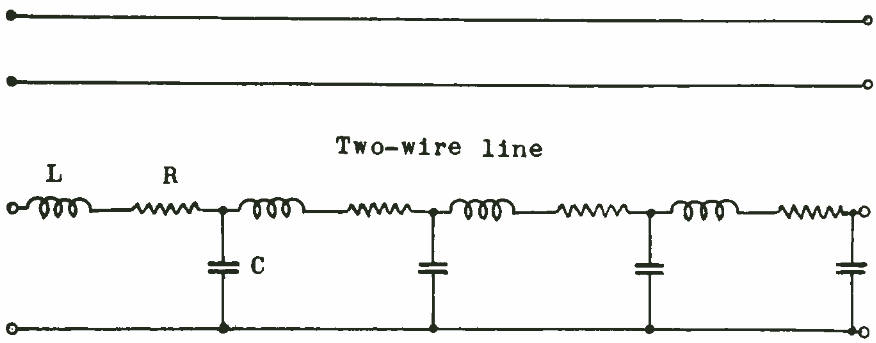
Figure 9

the lamp will glow proportionately to the *voltage* at its terminal. It should be noted that the actual voltage measured by the neon lamp is that across the very small capacity which exists between the wire and the operator's hand.

It may be asked why standing waves do not appear on low frequency lines as well as those operating at the higher frequencies. Actually these standing waves appear on *all* lines which are improperly terminated but they usually are not observed on low frequency lines because the operating wavelength is ordinarily quite long in comparison to the line length.

LINE CONSTANTS (Fundamental Relations): The distribution of the current and voltage in standing waves is greatly affected by the distributed constants of the line. When the inductance, capacity, and resistance of a circuit are distributed throughout a common conductor rather than being substantially in separate circuit components as in the case of conventional series or parallel circuits, it is said that the circuit has *distributed* constants; that is -- distributed inductance, capacity, and resistance.

The two-wire line, for example, is actually a rather complex network made up of shunt capacity, series inductance, and resistance. Since these constants are distributed throughout the entire length of the line, it is considered as having so much shunt capacity and series inductance and resistance *per unit of length*. Thus the two-wire line in Figure 10 may be represented



Equivalent circuit

Figure 10

by the equivalent iterative (repeating) network as shown. For example, if the line were four feet in length, the L, R, and C would represent the series inductance, series resistance, and and shunt capacity per foot of length, since four identical networks are shown.

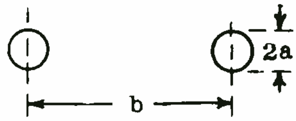
Actually there is no such network, there being a distributed capacity between adjacent wires throughout the entire length of the line. Similarly with the resistance and inductance, which of course are not present in lumps as shown. However, the two-wire line behaves exactly as though it were made up of such an equivalent network, and so is treated as such for simplicity. The student must remember, however, that the two circuits are equivalent at only one frequency for given values of the circuit constants. But since we are concerned with the behavior of the transmission line at one frequency (or at most over a very narrow band,) the above limited equivalence is sufficient for our discussion.

These distributed constants may be measured, or computed by well known formulas. For the two-wire line they are as follows:

$$L = (0.281 \log_{10} b/a) + .03 \ \mu h/\text{foot}$$

$$C = \frac{3.680}{\log_{10} b/a} \ \mu\mu f/\text{foot}$$

$$R = \frac{\sqrt{f}}{a} \times 10^{-6} \ \text{ohms/loop foot}$$



Two-wire line dimensions

L is the series inductance, C the shunt capacity, and R the series resistance. All dimensions are in inches and the frequency is measured in cycles/second. (It should be understood that the value of R obtained in this equation is for a "loop foot"; that is, it expresses the RF resistance introduced by both wires in the foot of two-wire line, a total of two feet of wire. Another expression frequently used is, $R = (4.16)(10^{-8})\frac{\sqrt{f}}{a}$ ohms/meter where a is the radius of the wire in meters and f is the frequency in cycles/second. This equation expresses the resistance of one meter of conductor; for a two-wire line the value of R should be multiplied by 2 to obtain the resistance of one "loop meter".) A practical idea of these values may be obtained by the consideration of a typical two-wire line; for example, one made up of No. 10 B & S bare copper wire, the radius (a) of which is .051 inches, spaced 5 inches center to center. Assume frequency to be 50 MC/s.

$$L = .281 \log_{10} 5/.051 + .03 = .281 \times 1.999 + .03 = .591 \ \mu h/\text{foot}$$

$$C = \frac{3.680}{\log_{10} 5/.051} = \frac{3.68}{1.999} = 1.84 \ \mu\mu f/\text{foot}$$

$$R = \frac{\sqrt{50 \times 10^{6}}}{.051} \times 10^{-6} = .1411 \ \text{ohms/foot at 50 MC/s}$$

It is interesting to note what would happen to these values of L, C, and R if the wires were placed closer together. An examination of the given equations will show that the inductance would decrease since the ratio b/a becomes smaller in value. This seems reasonable, for then the magnetic fields of the two wires would be closer together and would tend to cancel, *thereby diminishing the effective inductance.*

The capacity would increase. This also seems reasonable, for the opposing surfaces of the two parallel wires which make up this capacity would then be closer together.

However, an examination of the equation for resistance shows that it does not depend on the wire spacing. This is not strictly correct because of what is known as the *proximity effect*, that is, the non-uniform current distribution over the cross section of each wire caused by the flux of the adjacent wire. (For all practical purposes, however, this proximity effect may be neglected.)

Particularly observe the effects of variations of factors a and f upon the resistance of the conductor and the wide divergence that occurs between the R and the d.c. value of resistance. The resistance of No. 10 copper wire to the flow of direct current is .001 ohm/foot while at the radio frequency of 50 MC/s the resistance becomes approximately .07 ohm/foot, (1 foot of conductor is one-half of 1 foot of two-wire line), an increase of 70 times. As might be expected, the radio frequency resistance R varies inversely as a, because as a is increased the surface area of the conductor upon which the high frequency current travels increases in proportion. (The circumference of a circle varies in direct proportion to its radius.) The radio frequency resistance R varies directly as the square root of the frequency f. This is because the cross-section area of the current carrying conductor varies inversely as $\sqrt{f}$ due to the decreased depth of penetration of the current below the surface with increase in f. This is expressed by the equation for copper,

$$d.p. = \frac{.0664}{\sqrt{f}}$$

where d.p. is depth of penetration in meters
f is the frequency in cycles/second.

At 50 megacycles per second,

$$d.p. = \frac{.0664}{\sqrt{50 \times 10^6}}$$

$$= 93.9 \times 10^{-7} \text{ meters}$$

$$= .00037 \text{ inch}$$

Since the diameter of B. and S. Gauge No. 10 wire is approximately .1 inch it will be seen that an extremely small part of the cross section of the copper is usefully employed at this frequency.

CHARACTERISTIC IMPEDANCE: In Figures 7 and 8 the effect on the current and voltage distribution along the line was observed when the receiving end was open- or short-circuited. The third case is now considered, where a pure resistance equal to $\sqrt{L/C}$ (L is the series inductance and C the shunt capacity per unit length) is placed across the receiving end terminals. As shown in Figure 11, all resonances in the line are destroyed while the gradual

Figure 11

attenuation of the current and voltage towards the receiving end is caused by the line loss.

This particular value of receiver resistance ($R = \sqrt{L/C}$) is called the *characteristic impedance* (frequently called surge impedance) of the line and destroys all resonances, no matter what line length or frequency is being considered, since reflections cannot occur.

If the load at the receiving end of the line is a resistance, either greater or less than the characteristic impedance, the current and voltage distribution will resemble either Figure 7 or Figure 8 respectively, but the resonances in either case will not be so pronounced as that obtained with the open or short-circuit termination.

It is highly desirable to terminate transmission lines with their characteristic impedance for a number of reasons. If standing waves are allowed to appear on the line, the voltage stress between adjacent wires will greatly increase the leakage losses in the insulating spacers. These spacers, usually consisting of wood or some ceramic material, are placed at certain intervals along the line to maintain the spacing. When operating wavelength is short the spacers may be placed at half-wavelength intervals, in positions where the voltage is minimum, in order to reduce the leakage and possible breakdown caused by standing waves. Such arrangement is shown in Figure 12.

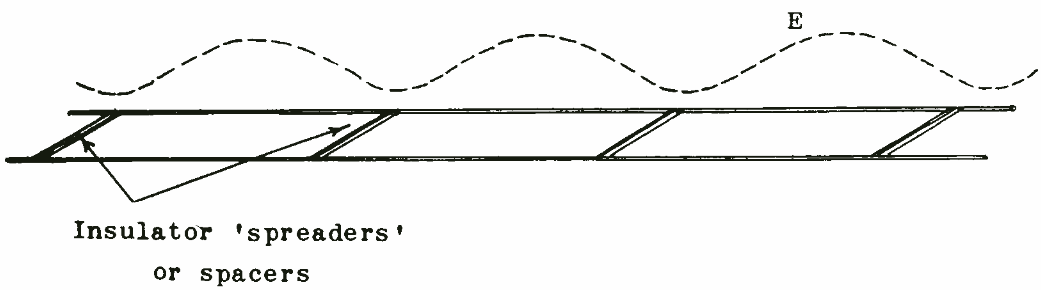


Figure 12

However, in addition to the other possible factors which may produce loss in the line, radiation losses must be considered. This loss of power from the line in the form of electromagnetic radiation varies roughly as the square of the maximum current in the resonant line.

For these reasons, *since standing waves always greatly in-*

crease the radiation and leakage losses in line, it is highly desirable properly to terminate a transmission line with its characteristic impedance.

Practical values of characteristic impedance (written Z_0) in r-f transmission lines will range anywhere from 70 to 800 ohms approximately. For example, consider the No. 10 two-wire line discussed above, whose distributed constants were found to be: L = .591 μh/foot, C = 1.84 μμf/foot.

$$Z_0 = \sqrt{\frac{L}{C}} = \sqrt{\frac{.591 \times 10^{-6}}{1.84 \times 10^{-12}}} = 567 \text{ ohms}$$

Since computing the distributed constants, L and C, in order to find the line impedance is a lengthy process, the equations for L and C may be substituted in the equation for Z_0 and an approximate formula derived which is much easier to use:

$$Z_0 = 276 \log_{10} b/a$$

This formula, and the one to be given later for Z_0 of a coaxial line, neglect the effects of the magnetic field inside the metal of the conductor and the distortion of current distribution on the conductors due to the proximity effect. (Because of the symmetry of the electric field within the line, the latter effect is not present in a coaxial line.) In a two-wire line the error introduced by neglecting this effect is 5.5 per cent when b/a = 4, becoming progressively less as the separation between conductors is increased. In most two-wire lines used for power transfer, b/a is quite large--150 in the case of a 600 ohm line--and the error due to neglecting these effects is negligible.

Following are several problems which illustrate the usefulness of this formula in determining line dimensions.

1. Determine the surge impedance of a two-wire line using

1 inch (diameter) copper tubing spaced 4 inches (center to center)

$$Z_o = 276 \log_{10} 4/.5 = 276 \times .903 = 249 \text{ ohms}$$

2. Determine the spacing required for a two-wire line using No. 12 B & S wire if a characteristic impedance of 600 ohms is desired.

$$600 = 276 \log_{10} b/a \quad \text{(diameter of No. 12 wire = .0808 in.)}$$

$$b/a = \text{antilog} \frac{600}{276} = 149.2$$

$$b = 149.2a = 149.2 \times .0404 = 6.03 \text{ inches}$$

(6 inch spacers, commercially available, would be used in this case to maintain the line spacing.)

On the following page is a graphical plot of the equation $Z_o = 276 \log_{10} b/a$ which will be found convenient to use for calculations involving transmission lines. Below are several examples which show the use of this graph.

1. What should be the spacing between wire centers in a two-wire 400 ohm line employing 1/4 inch tubing? (Graph is read along dotted lines.)

Answer: 3.5 inches

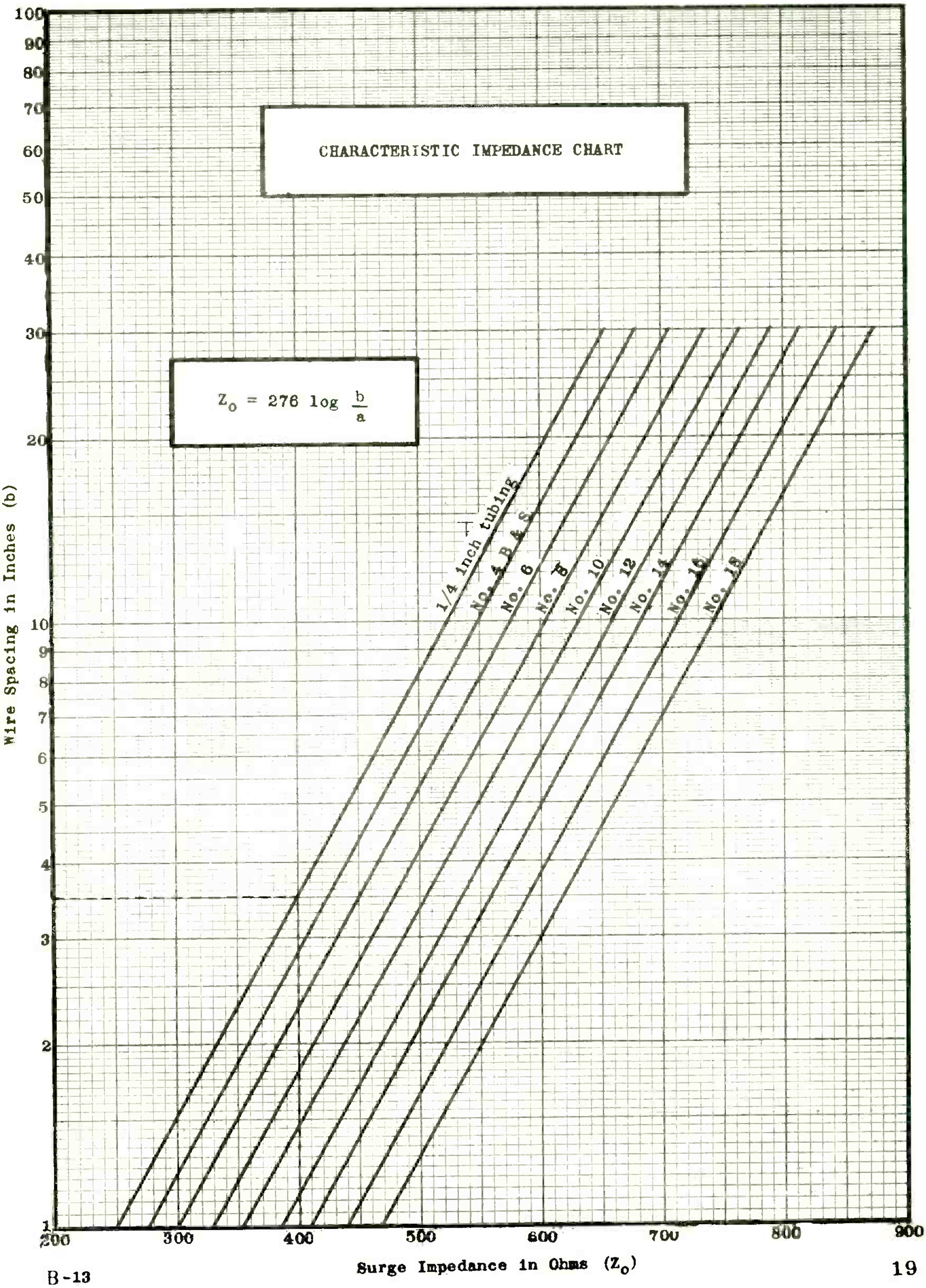
2. What wire size should be used in a two-wire line with 6 inch spacers if the desired surge impedance is 600 ohms?

Answer: No. 12

3. Same as example 2 for 5 inch spacers.

Answer: No. 14

4. Same as example 2 for 3 inch spacers.


CHARACTERISTIC IMPEDANCE CHART
$Z_o = 276 \log \frac{b}{a}$
Wire Spacing in Inches (b)
Surge Impedance in Ohms (Z_o)
1/4 inch tubing
No. 4 B & S.
No. 6
No. 8
No. 10
No. 12
No. 14
No. 16
No. 18
100
90
80
70
60
50
40
30
20
10
9
8
7
6
5
4
3
2
1
200
300
400
500
600
700
800
900

Answer: No. 18

5. What will be the surge impedance of a two-wire line if No. 4 wire is used with a spacing of 10 inches (center to center)?

Answer: 550 ohms

6. What spacing should be used in a 600 ohm line employing No. 10?

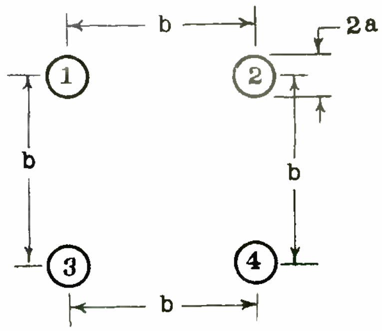
Answer: 7.6 inches exactly although in actual practice commercial 8 inch spacers would probably be used.

7. What size wire must be used in a two-wire line if 500 ohms surge impedance is desired and 3 inch spacers are to be used?

Answer: No. 11 is the nearest size although in practice the nearest available even wire size may be used--No. 10. This would give an impedance of 490 ohms which is still quite close to the desired value.

OTHER TYPES OF LINE: Since the magnetic field which surrounds a transmission line as the result of current flow will produce undesired radiations and crosstalk with adjacent lines, it is desirable to make this field as weak as possible. One method of doing this is to place the two wires as close together as possible so that the fields will neutralize. However, this lowers the line break-down voltage and also greatly increases the insulation losses since *the electrostatic field will become stronger* as the wires are brought closer together. More-over, this will also reduce the surge impedance unless the wire diameter is suitably reduced. But, reduction of the wire diameter increases the I^2R losses and thus the total losses. Hence this is not a good practical solution.

Another method would be to use multiple wire lines by making up several pairs of go and return circuits in parallel. The cross section of such an arrangement is shown in Figure 13 where four wires are used. Numbers 1 and 4 are connected together at each end of the line to make up one path and similarly wires 2 and 3 are tied together at each end of the line to make up the return path.



$$Z_o = 138 \log_{10} \frac{b}{a} - 20.8$$

Figure 13

CONCENTRIC LINES: Although the above arrangement of multiple wires greatly reduces the magnetic field, from the standpoint of isolation an ideal arrangement for a two conductor circuit is approached when one conductor completely encloses the other. This type of line, shown in Figure 14, is widely used in radio

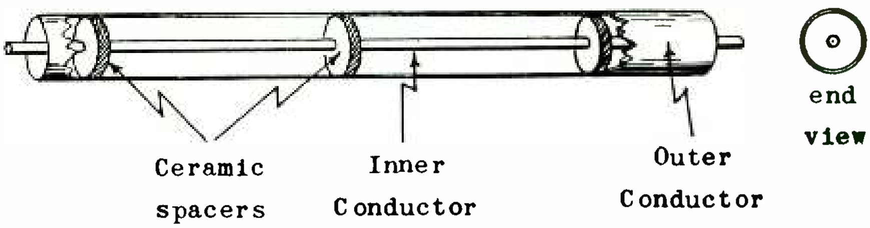


Figure 14

circuits and is known as a "coaxial" or "concentric" line since one conductor is concentric about the other, that is the two are coaxial.

For large diameter concentric lines, copper tubing is normally used for both conductors. For small sizes, however, solid copper wire is used for the inner conductor. The inner conductor is held in position by ceramic spacers placed at regular intervals along the line. These spacers, as shown in Figure 15, may be any one of several different shapes.

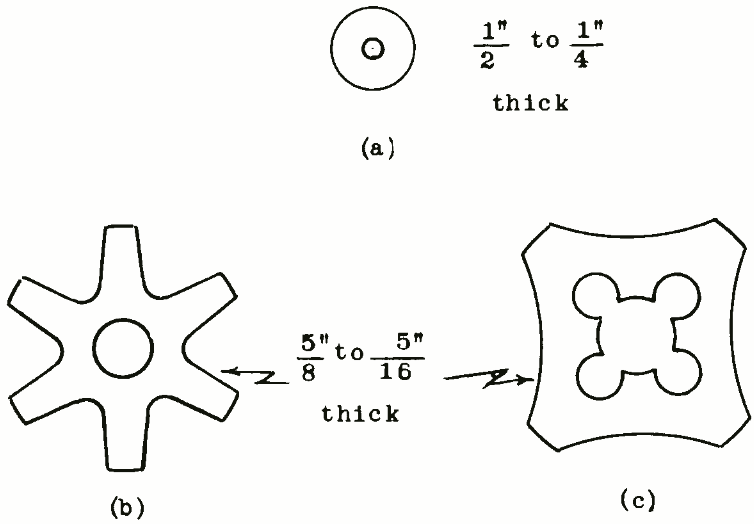


Figure 15

Beads of high grade porcelain (a) in diameters up to one inch are satisfactory insulators for low power and receiving lines. For higher voltages, which necessitate greater spacing, this type of spacer is undesirable because the greater volume of the dielectric may cause sufficient heating to break down the insulator. It is therefore desirable to use type (b) or (c) where the line is to withstand high voltage stresses.

In choosing a concentric line for any particular operation it is important that the construction be sufficiently rugged for the job at hand. If the line is to be employed for high power transmitting purposes the voltage safety factor may be so low that accidental dents in the sheath may lead to breakdown. Sterba and Feldman have found that for outer sheaths a diameter of 2.5

inches and radial thickness of .0875 to .10 inch provides lines which are sufficiently rugged for transmitting 15 K.W. of modulated power at 16 meters wavelength.

Concentric lines have several advantages over two-wire lines. The construction minimizes skin effect by arranging the surface of the conductor parallel with the magnetic flux lines. *The electrostatic and electromagnetic fields between the two conductors are confined by the outer conductor which furnishes a very effective shield, usually grounded.* For this reason radiation losses and pickup from external sources are minimized. In addition, the line constants are unaffected by the operator's hands or adjustment tools in the vicinity, which is a decided advantage in concentric lines used as tuned elements for oscillator circuits. Concentric lines may be constructed so as to be weatherproof whereas the ice, snow, or moisture which collects on open two-wire lines may cause a serious change in the line constants, line loss, and breakdown voltage.

For a given minimum conductor radius the resistance of a coaxial line is considerably less than that of an equivalent two-wire line; if the radius of the inner conductor is the same as that used in the two-wire line, the radius of the outer conductor will be 3.5 to 4 times greater. The equation:

$$R = (4.16)(10^{-8})\frac{\sqrt{f}}{a} \text{ ohms/meter}$$

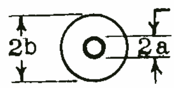
applies to the outer conductor of the coaxial line when a equals the inner radius of that conductor. Thus the resistance of a "loop meter" of coaxial line will be the sum of the resistance of the inner and outer conductors, a for the former being computed from the outside dimension and a for the latter from the inside dimension. The resultant R will be very substantially less than that of a two-wire line made up of conductors equivalent to the inner conductor of the coaxial line. This is an important factor in some applications.

Concentric transmission lines are more expensive than two-wire lines since they must be carefully constructed to operate satisfactorily. They must be closed tightly to exclude moisture and often are filled with some inert gas, such as nitrogen, under pressure in order to increase the breakdown voltage. On the other hand, although concentric lines are more costly than open wire lines, they permit the installation of a number of radio units within a single structure without incurring difficulties from cross talk between lines. For this reason the compact installation obtainable with concentric lines as compared with that of widely separated structures necessary with open lines may more than offset the additional cost of the lines.

However, in spite of increased loss and cross talk difficulties encountered in "open" two-wire lines, they are used extensively for transmission purposes for two reasons. First, they are easily constructed from inexpensive available materials. Second, the line's dimensions may be easily changed if a higher breakdown voltage, Q, or different characteristic impedance is desired.

The characteristic impedance of a concentric line is found approximately by:

$$Z_o = 138 \log_{10} b/a$$



In this case b is the inner radius of the outer conductor and a is the outer radius of the inner conductor. It has been found that both concentric and two-wire lines have the highest Q (lowest loss) when b/a is approximately 3.6. For this reason, practically all concentric lines commercially available have a dimension b/a between 3 and 4.

For example, consider a typical concentric line whose dimensions are 2b = 2.9 inches, 2a = .875 inches,

$$Z_o = 138 \log_{10} \frac{2.9}{.875} = 138 \times .520 = 71.8 \text{ ohms}$$

It is to be noted that since b/a for concentric lines will usually be between 3 and 4, the characteristic impedance under that condition will always be approximately 70 to 90 ohms.

USES OF TRANSMISSION LINES

LINES AS TRANSMISSION ELEMENTS: The fundamental purpose of transmission lines is that of transmitting energy from one point to another. For this purpose both open two-wire and concentric lines are widely used. A primary consideration in a line used for such a purpose is the proper termination in order to prevent undesirable reactions, previously discussed. Second, the line should have a sufficiently high breakdown voltage and should not offer too much loss.

As a practical example of such use, consider the line which is to link a radio transmitter to its antenna, Fig. 16. If

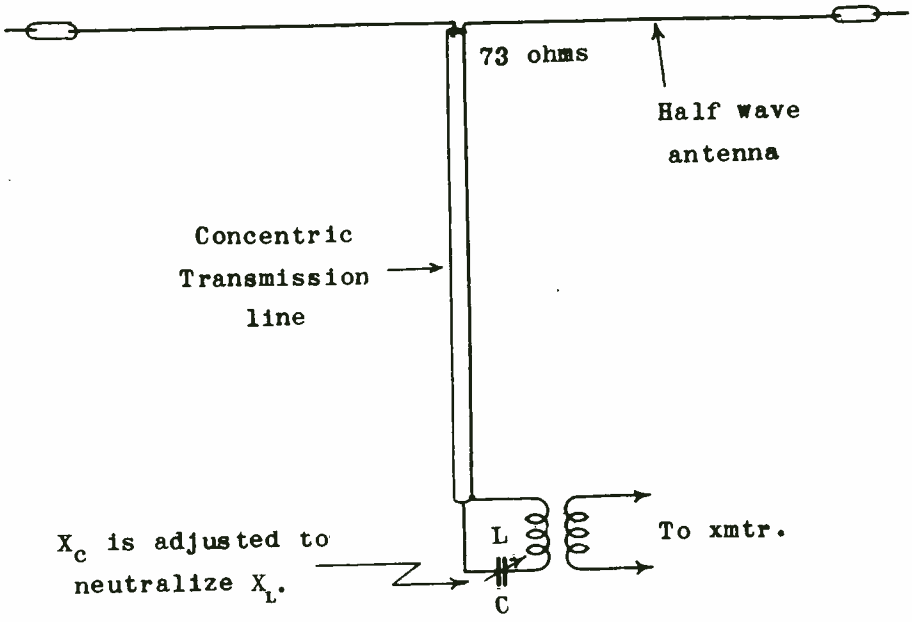


Figure 16

the antenna is a simple halfwave, center-fed, Hertz type, the input resistance will be approximately 73 ohms. It is therefore desirable to use a transmission line whose characteristic impedance is approximately 73 ohms for then the line will look into an impedance at the receiving (antenna) end which properly terminates it and standing waves will not appear. If this is not done, the standing waves which result cause undesirable radiation losses in the line, possible breakdown of the insulation at points of maximum voltage stress, *and reduced power transfer to the antenna* due to the mismatch (unless a correction is made) between the line and the antenna impedance. (It must be remembered that to the antenna the transmission line is the apparent source *and therefore the line impedance must be matched for maximum power transfer.)*

Since most concentric lines used for power transmission have a surge impedance of approximately 70 ohms, any line of this type having the necessary breakdown voltage will be satisfactory since the line termination usually is not very critical. (Methods of impedance matching where an exact match is required, will be discussed later.)

At 100% modulation, since the peak power is four times the r.m.s. value, the line voltage becomes:

$$E = \sqrt{4PZ_0} \quad \text{(r.m.s.)}$$

$$E = \sqrt{2}\,\sqrt{4PZ_0} \quad \text{(peak)}$$

$$E = \sqrt{8PZ_0} \quad \text{peak voltage}$$

where P = unmodulated carrier power

For example, if the transmitter output is 1KW into a 72 ohm line, the peak voltage across the line at 100% modulation will be

$$E = \sqrt{8 \times 1000 \times 72} = 760 \text{ peak volts}$$

The 72 ohm concentric line to be used, then, should have a breakdown voltage of at least 1000 volts. It should be noted that *760 volts is the peak value with proper termination and no standing waves.* This voltage may be far exceeded under conditions of mismatch because of the standing waves which will appear. For this reason preliminary adjustments should be made with reduced power and full power applied only after it is assured that the line is properly terminated.

It may be desired to use a two-wire line in the above circuit; what dimensions would be required?

$$72 = 276 \log b/a$$

$$b/a = \text{antilog } 72/276$$

$$= 1.823$$

It can be seen that for the two-wire line, b/a cannot be made less than 2.0, even with the two wires touching. However, as shown in Figure 17, the equation used for Z_0 is only approximate, and in actual practice certain twisted pair transmission lines may have an impedance as low as 70 ohms. For a short run

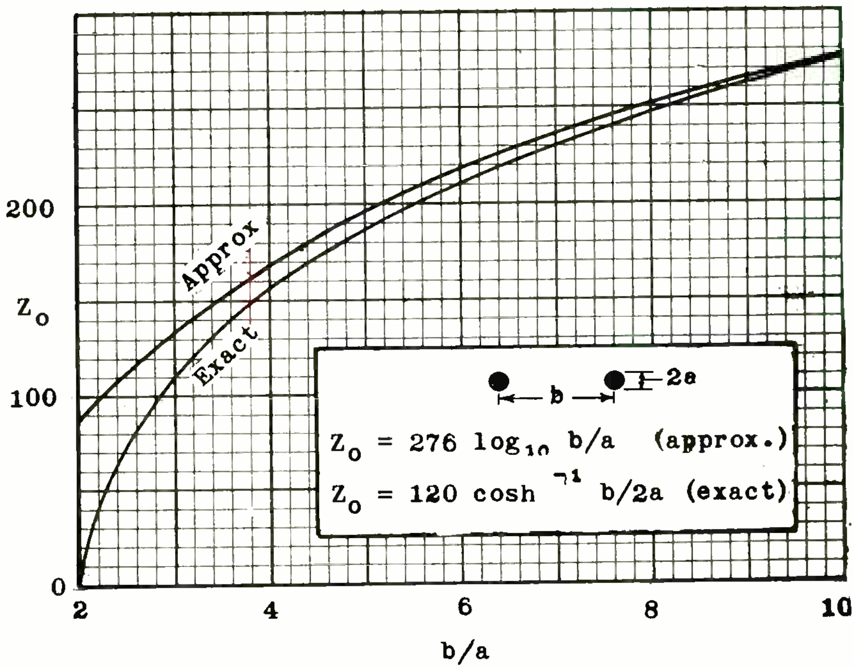


Figure 17

at low power, twisted pairs often will be quite satisfactory and inexpensive in an amateur or emergency installation, but for aeronautical radio transmission or for any long run between transmitter and antenna the loss in such a line is too high. Twisted pair lamp cord, for example, has a loss of about 1.5 db per wavelength when dry, and a characteristic impedance of about

130 ohms. With such a line used to deliver the 1000 watts output from transmitter to antenna the power loss in one wavelength would be

Since $10 \log P_{(in)}/P_{(out)}$ = attenuation in decibels

$$P_{(out)} = \frac{P_{(in)}}{\text{antilog db}/10}$$

$$P_{loss} = P_{(in)} - P_{(out)} = P_{(in)} \left(1 - \frac{1}{\text{antilog db}/10}\right)$$

$$P_{loss} = 1000\left(1 - \frac{1}{\text{antilog } 1.5/10}\right) = 293 \text{ watts.}$$

At 60 MC/s, a 5 meter (16.4 feet) length of the above twisted pair line would give 293 watts loss. A 10 meter length would offer 500 watts loss! Also of course such a line could not be used at the peak voltages developed with 1000 watt output.

The loss in a terminated non-resonant transmission line may be determined by:

$$\text{Power loss} = \frac{8.686\ Rl}{2Z_o} \text{ db}$$

where

Z_o = surge impedance of the line in ohms
l = length of the line in feet
R = series resistance of the line per foot

Two-wire Line *Concentric Line*

$R = \frac{\sqrt{f}}{a} \times 10^{-6}$ ohms/loop foot $\qquad R = 5\sqrt{f}\,(1/a + 1/b) \times 10^{-7}$ ohms/foot

(a and b are measured in inches, f is in cycles/second)

For example, consider a 1000 foot length of concentric line whose dimensions are b = 2.75 in., a = .75 in. At 80 MC/s

$$R = 5\sqrt{80 \times 10^{6}}\ (1/.75 + 1/2.75) \times 10^{-7}$$

$$= 5 \times 8.94 \times 10^{3} \times 1.694 \times 10^{-7}$$

$$= 7.58 \times 10^{-3} \text{ ohms/foot}$$

$$Z_o = 138 \log_{10} 2.75/.75 = 138 \times .565 = 78 \text{ ohms}$$

$$\text{Power loss} = \frac{8.686 \times 7.58 \times 10^{-3} \times 10^{3}}{2 \times 78}$$

$$= .422 \text{ db}$$

For 1000 watts input to the line, the line loss is

$$P_{loss} = 1000\ (1 - \frac{1}{\text{antilog } .422/10})$$

$$= 92 \text{ watts line loss}$$

Such a line loss is quite tolerable for such a long line. Compare this figure with the loss in a two-wire line of the same length, 168 watts; and with that of a 1000 ft. length of twisted pair, which would attenuate the input down to essentially zero.

Power loss or attenuation in a transmission line usually is expressed in decibels per meter, (db/meter). The line attenuation of a *properly terminated* two-wire line is expressed in decibels/meter as

$$\frac{(1.3)\,(10^{-9})\,\sqrt{f}}{a \log_{10} b/a} \text{ db/meter} \qquad \text{(a and b are in meters)}$$

The attenuation varies directly as $\sqrt{f}$ just as does the line resistance; it varies inversely as a, as does the line resistance because as a is increased the resistance is decreased; and it varies inversely as the $(\log_{10} b/a)$ because as the ratio b/a is increased the proximity effect of the two conductors is decreased and the line resistance is decreased.

The line attenuation per meter of a *properly terminated* coaxial line is given by the expression,

$$\frac{(1.3)\,(10^{-9})\,\sqrt{f}}{\log_{10} b/a}\left[\frac{1}{a} + \frac{1}{b}\right] \text{ db/meter} \qquad \text{(a and b are in meters)}$$

For a given value of diameter of outside conductor the attenuation in a coaxial line is a minimum when b/a is approximately 3.5.

Example 1: A two-wire line consisting of No. 12 wire spaced .33 inch is 400 feet long. What is the line attenuation in decibels when operated at 70 MC/s?

a for No. 12 wire = 40.4 mils = .001028 meter

$b/a = .33/.0404 = 8.2$

$\log_{10} 8.2 = .914$

$f = 70 \times 10^6$

$$db = \frac{(1.3)\ (10^{-9})\ \sqrt{70 \times 10^6}}{.001028 \times\ .914} = .0116 \text{ db/meter}$$

400 feet = 122 meters

Line attenuation = 122 x .0116 = 1.41 db.

Example 2: A coaxial line is to be substituted for the two-wire line in the example above. The inner conductor is to have an outside diameter (2a) of .28 in., the outer conductor is to have an inside diameter (2b) of 1 in. What is the line attenuation at 70 MC/s?

a = .14 in. = .00356 meter

b = .5 in. = .0127 meter

$b/a = .0127/.0035 = 3.57$

$\log_{10} 3.57 = .553$

$$\sqrt{f} = \sqrt{70 \times 10^6} = 8370$$

$$\frac{1}{a} + \frac{1}{b} = \frac{1}{.00356} + \frac{1}{.0127} = 360$$

$$db = \frac{1.3 \times 10^{-9} \times 8370 \times 360}{.553} = .00708 \text{ db/meter}$$

Line attenuation = 122 x .00708 = .864 db

The above examples show that the coaxial cable has a lower attenuation than the two-wire line ---.864 db attenuation as compared to 1.41 db. If the two types of lines were compared on a basis of equal surge impedance the advantage of the coaxial line would appear even greater. However, in the case of the two-wire line, considerations of practical wire size and spacing favor a higher surge impedance than that for coaxial lines. A higher surge impedance means a higher voltage and a smaller current for a given amount of power transmitted. For example, compare the voltages and currents when 1000 watts are transmitted over a 600 ohm two-wire line and a 70 ohm coaxial line:

$I = \sqrt{P/R} = \sqrt{1000/70} = 3.79$ ampere (coaxial)

$E = IZ = 3.79 \times 76.4 = 290$ volts r.m.s. (coaxial)

$I = \sqrt{1000/600} = 1.292$ amperes (two-wire)

$E = 1.292 \times 600 = 775$ volts r.m.s. (two-wire)

The smaller current of the two-wire line will reduce the copper losses, and if the surge impedance is made quite high, these losses in the two-wire line may actually be less than that in the coaxial line. For large amounts of power, however, the high voltage present in a high impedance line may produce excessive corona and dielectric losses so that the coaxial line of lower surge impedance may in this case actually have less attenuation. From the above discussion it may be concluded that there is little to choose between the two types of line in terms of attenuation. Other practical considerations, such as compactness of structure, freedom from radiation, independence of climatic conditions (sleet, rain etc.) tend to make the more expensive coaxial cable give a superior performance.

It probably is assumed by this time that the study of transmission lines is simply the study of mathematical formulas. This is far from the truth, however. Even though the relations in a transmission line are best studied by means of formulas, these formulas are not exact, and at best will serve only as first approximations in practical design. Their error in some cases may be as high as 20 percent. The engineer who works with transmission lines is truly one "with a slide rule in one hand and a soldering iron in the other". In all work at the ultra-high frequencies one must necessarily strike a happy medium between first approximation and practical adjustment, since all formulas may be quite in error.

For example, the loss in a transmission line is actually dependent upon a number of other factors which working formulas cannot take into account--degree of oxidation on the conductor surface, moisture in the air or on the conductors, dielectric constant of the ceramic spacers in the coaxial line and across the two-wire line--and many other items.

With so many factors to limit the accuracy of purely theoretical formulas it is little wonder that the engineer adopts simpler though approximate thumb-rule equations such as the following:

For the two-wire line:

$$\text{Power loss} = 0.1\sqrt{f}\ \text{db per 1000 feet.}$$

For the concentric line:

$$\text{Power loss} = \frac{0.128}{b}\sqrt{f}\ \text{db per 1000 feet.}$$

where f is the operating frequency in megacycles per second and b is the inside radius of the outer conductor in inches.

For example, by the use of the approximate equation in the case of the 78 ohm 1000 foot concentric line discussed above, the power loss at 80 MC/s is found to be:

$$\text{Power loss} = \frac{0.128\sqrt{80}}{2.75} = .416\ \text{db}$$

(This agrees very closely with the calculation employing the other formula.)

RESONANT LINES AS CIRCUIT ELEMENTS: So far the discussion has been limited to lines as used for transmission purposes. However this is not always the purpose of a line. Transmission lines that are either open or short circuited at the receiving end have many of the properties of ordinary resonant circuits. For example, as shown in Fig. 7, if the line is open at the receiver and is an odd number of quarter-wavelengths long, the *receiving voltage is much higher than the sending voltage, giving a resonant rise of voltage like that obtained with a series circuit.*

The comparison is also shown by Fig. 18 (a) and (b). Fig. 18 (a) is a diagram of the normal grid circuit connections in an r.f. amplifier. The operation which takes place in this circuit is well known. The voltage induced in secondary L by transformer action causes a resonant current to flow around the series loop LC. The resonant rise in voltage across the condenser is then applied between grid and cathode of the tube and amplified.

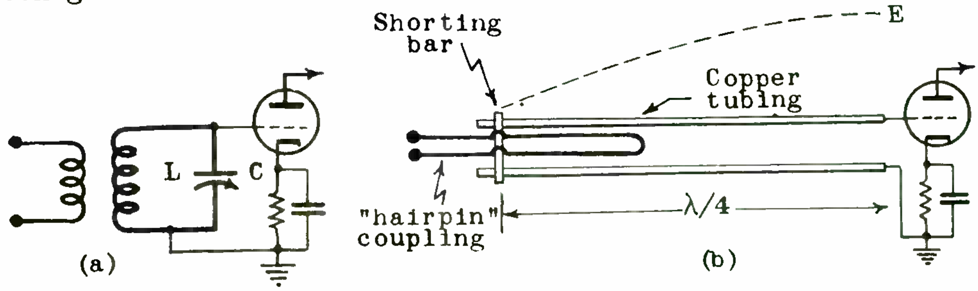


Figure 18

It will be recalled that in any series resonant circuit the voltage drop across the reactances is always Q times the applied voltage. In this case, for example, if the circuit Q is 100 and the voltage induced in L is 5 μ-volts, the voltage appearing across the condenser C will be 5 x 100 or 500 μ-volts. It is well to note that since the amplification factor of the triode may not be much above 20, *most of the stage gain occurs in the tuned circuit.* For this reason one should give particular attention to the circuits associated with any r.f. amplifier, always attempting to keep the circuit losses to a minimum, or in other words, to keep the circuit Q as high as possible.

Fig. 18 (b) shows a quarter-wavelength tuned line used in place of the tuned circuit LC. This line is excited by a single loop of wire inductively coupled to the tuned line and commonly referred to as a hairpin coupling. The shorting bar is a wire strap connecting the lines which may be shifted up or down the tubing as necessary in order to adjust the line to exactly one-quarter wavelength.

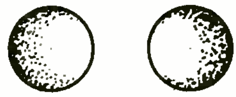
Reference to Fig. 7 indicates that the voltage will distribute itself along the line as shown by curve E. That is, a small voltage induced at the sending end will appear as a much larger voltage across the receiving end of the line. A resonant rise in voltage always takes place in any line an odd number of quarter-wavelengths long that is open-circuited at the receiving end.

Since the voltage amplification by such a line is inversely proportional to the series resistance per unit length, every effort is made to keep this resistance to a minimum. The lines are usually made of copper tubing to reduce the skin effect losses to a minimum; at the higher frequencies it is found economical to use iron rods with a very thin surface of silver plating. This is made possible by the fact that at higher ultra-high frequencies the current penetration below the surface is practically infinitesimal.

It should be remembered that this resonant rise which takes place in the line is due to the fact that its far end is *open circuited.* Actually the tube's input impedance (grid to cathode) is not infinite and therefore will load the line somewhat. As a matter of fact, if the grid is driven positive the tube's input impedance will become quite low and the voltage gain of the line will be considerably reduced.

In addition to the above factors which must be considered, one of the most important is the line spacing. If the lines are spaced far apart the loss due to proximity effect will be small but radiation losses will be high because the electromagnetic fields of the two conductors do not cancel each other very effectively. On the other hand, if the two conductors are placed very close together the radiation loss will be small but the loss due to proximity effect will be greatly increased. The result of proximity effect is shown in Figure 19.

As the conductors are placed close together the *current is redistributed over the conductor cross section in such a way as to make most of the current flow where it is encircled by the smallest number of flux lines.* This general principle determines the distribution of current in any conductor, irrespective of the shape. Note in Figure 19 that the current tends to distribute itself on the surface of each conductor away from the other. This substantially reduces the available current-carrying cross section and therefore *the conductor's effective resistance is increased.*



Current density is shown by the density of shading.

Figure 19

From the above discussion it will be seen that there is an optimum value of the ratio b/a that will give minimum line loss. This value will lie somewhere between the limits b/a = 3 and b/a = 4 depending upon which of the line dimensions, a or b, is held constant and which is being adjusted.

As mentioned previously, the optimum ratio with a fixed line spacing, or outer sheath diameter in the case of a concentric line will be approximately 3.6.

The maximum Q obtainable with a tuned line (b/a = approx. 3.6.) will be, approximately,

$$Q = 220b \sqrt{f} \text{ MC/s} \qquad \text{where b is in inches.}$$

For example, consider the "resonant line" tuned circuit of Figure 18(b). Using two inch diameter copper tubing spaced 3.6 in. center to center; the Q at 100 MC/s would be

$$Q = 220 \times 3.6 \times \sqrt{100} = 7920$$

A more general equation for the Q of a resonant line is given by

$$Q = \frac{2\pi Z_o}{R\lambda_o}$$

where R is the resistance per loop meter of the line, Z_o is the line characteristic impedance, and λ_o is the wavelength of the propagation measured at line velocity. If it is assumed that line velocity equals velocity of light, calculate the Q of the line as specified on the preceding page.

$$Z_o = 276 \log_{10} b/a$$

$$b/a = 3.6 \text{ and } \log_{10} 3.6 = .5563$$

$$Z_o = 276 \times .5563 = 153.5 \text{ ohms}$$

$$\lambda_o = 3 \text{ meters}$$

$$R = \frac{\sqrt{f}}{a} \times 10^{-6} \times 3.28 \text{ ohms/loop meter} \quad \text{(a in inches)}$$

$$= \frac{\sqrt{10^8}}{1} \times 10^{-6} \times 3.28 = .0328 \text{ ohms/loop meter}$$

$$Q = \frac{2\pi Z_o}{R\lambda_o} = \frac{6.28 \times 153.5}{.0328 \times 3} = \frac{964}{.0984} = 9800$$

It will be observed that there is a discrepancy between the values of Q obtained from the two formulas. This discrepancy is not more than is to be expected in ultra-high frequency calculations where unmeasured approximations are used. For example, in the second calculation it is assumed that the line velocity equals the velocity of light--while it is well known that the line velocity at the specified frequency is substantially below the velocity of light. However these calculations give with usable accuracy the order of the value of Q that will be obtained and the general formula demonstrates clearly the manner in which Q will vary with variations in line constants and frequency.

The calculated values of Q are much higher than could be obtained with the coil and condenser of Figure 18(a), since it is impractical to wind a coil with a Q much greater than 500. However such tuned lines are only practical at the ultra-high frequencies. When the frequency is sufficiently high so that the length required is physically short, it is possible by suitable design to obtain circuit Q's ranging from 1000 to 100,000, with values in the order of 10,000 quite practical under most circumstances.

The increase in Q as the frequency increases results from the fact that, although the skin effect resistance is proportional to the square root of frequency, the length of line required for a quarter-wavelength is inversely proportional to frequency, so that, as the frequency is increased, the length reduces faster than the skin effect increases. Resonant lines are therefore particularly suitable for use at extremely high frequencies, since then the Q is high and the physical dimensions are small.

Since tuned lines have such high Q they are especially useful in oscillator circuits in order to obtain very high frequency stability. For such purpose they simply replace the tuned circuits ordinarily used. For example, note the Hartley oscillator shown below in Figure 20 (a). In Figure 20 (b) the

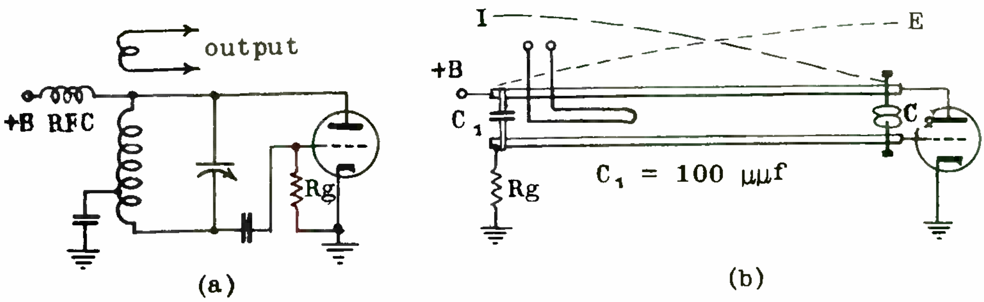


Figure 20

same oscillator is shown with the tuned circuit replaced by a resonant quarter-wavelength line, "shorted" at one end by the 100 μμf condenser which is sufficiently large to have a negligible reactance at the operating frequency (120 MC/s). It will be noted that a shorting bar could not be used in this case since the d.c. plate voltage would then be applied directly to the grid. The r.f. choke shown in Figure 20(a) to keep r.f. current out of the power supply is unnecessary in the circuit of Figure 20(b) since the +B voltage is applied to a point which is at low r.f. potential. (It is interesting to note that a choke coil, even if used, would act as a condenser at such a high frequency due to its distributed capacity.)

A comparison of Figures 20(a) and 20(b) may raise the question in the student's mind as to how the potential of the resonant line in Figure 20(b) is fixed with respect to ground and therefore the cathode. In Figure 20(a) there is a by-pass condenser from a *tap* on the tank coil to ground. In Figure 20(b) there is a grid resistance R_g from *one* side of the resonant line to ground. This grid resistance does not tie that end of the line to ground, however. Instead, at this high frequency, the grid-to-cathode interelectrode capacity C_{gf} and the plate-to-cathode interelectrode capacity C_{pf} serve as a low reactance capacitive voltage divider to set the potential of the two sides of the line with respect to the cathode and hence with respect to ground. Thus, if at some instant the top conductor has an a.c. polarity positive to ground, then the bottom conductor has a negative a.c. polarity to ground, just as in the case of Figure 20(a), where the by-pass condenser mentioned above performs this service.

The load is inductively coupled to the tuned line oscillator by hairpin coupling. Since the voltage induced in the coupling link is proportional to the magnetic flux which links with it, the "hairpin", is always placed near the shorted end of the line where the current producing this flux will be maximum.

It will be recalled that the mutual inductance required for maximum power transfer from primary to secondary is given by the relationship, $M = \sqrt{R_p R_s}/\omega$, where M is the mutual inductance, R_p is the primary resistance, and R_s is the secondary resistance. It is readily seen that for extremely high frequencies very little mutual inductance is required and it has been found experimentally that a single loop or "hairpin" is quite sufficient at U.H.F. The hairpin may be bent closer to or away from the tuned line as is necessary in order to adjust the mutual to the required value. Note, for example, the position of the loop in Figure 3. A hinged mounting may be used to simplify this adjustment.

The correct adjustment of the line length for resonance can be made by adjusting the position of the shorting condenser at the end of the line, or by means of the disc-type trimmer condenser C_2. The mechanical arrangement of this condenser, as shown in Figure 3, consists of two copper discs about 2 inches in diameter mounted on brass screws. These screws are threaded through the copper tubing and the plate spacing may be adjusted by means of a screwdriver. Since this condenser, together with C_{pf} and C_{gf} in series, increase the capacity (and consequently the LC product) of the line, *the electrical wavelength of the line is effectively increased.*

By using a tuned line physically somewhat shorter than a quarter-wavelength, shunted at the open end by a small trimmer condenser to increase the effective length to the required value, very accurate adjustment of the resonant frequency may be obtained. In practice C_2 may be used for vernier tuning after preliminary adjustment of the shorting bar to the approximate position.

CONCENTRIC LINE OSCILLATORS: Concentric lines also may be used as oscillator tuned circuits. A practical oscillator circuit utilizing a resonant concentric line to control the

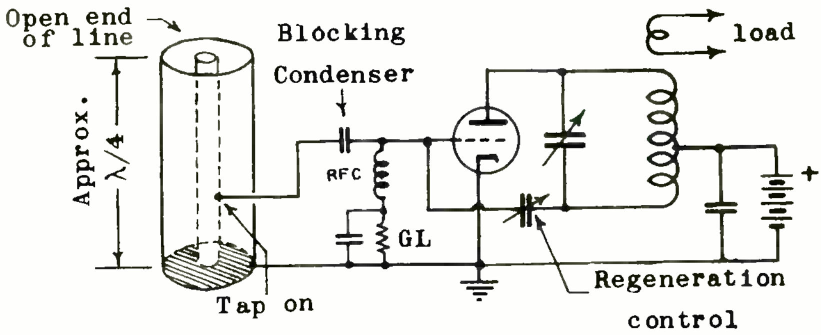


Figure 21

the frequency is shown in Figure 21. The grid tuned circuit controls the frequency, and, because of the extremely high Q obtained with the resonant line, the frequency is substantially independent of ordinary variations in power supply voltages or load coupling.

It will be recalled that the standing wave of voltage on a quarter-wave resonant line is maximum at the open circuited end and minimum at the shorted end. For maximum frequency stability the resonant line must be loosely coupled to the grid of the tube in order that the grid circuit losses will not be excessive. This is accomplished by connecting the grid to the line at a point relatively close to the shorted end where the voltage tapped off to drive the grid will not be excessive.

In operating this circuit it is best to adjust the regeneration control condenser to give about the smallest feedback from plate to grid circuit which can be used to make the oscillator work efficiently. Any excess feedback reduces the ability of the line to stabilize the frequency. The circuit will function with the regeneration control condenser set either above or below the capacity value required for a balance but one adjustment or the other will be preferable depending upon the ratio of effective resistance in plate and grid circuits and the frequency.

Tuned lines, especially of the concentric type, are very well suited for the control of ultra-high frequency oscillators and will perform satisfactorily at frequencies far higher than can be reached by crystals. As one goes to higher frequencies, the crystals used become increasingly thin and fragile, gradually becoming less useful in stabilizing the oscillator frequency, while lines increase in their stabilizing ability as the frequency is increased.

Hansell and Carter, discussing frequency control, state: "It is interesting to note that any phase or frequency modulation noise introduced in any early stage of a transmitter will be increased in proportion to the amount of frequency multiplication used after that stage. A crystal-controlled transmitter having an output of 100,000 kilocycles would probably start out with an oscillator frequency of about 3125 kilocycles. One degree of phase modulation in the output of the crystal oscillator would then appear as thirty-two degrees in the output of the transmitter and produce side frequency energy equivalent to that obtained with about sixty per cent amplitude modulation. With line control the oscillator may be operated at the output frequency so that one degree of phase modulation in the oscillator will appear as one degree in the transmitter output."

TEMPERATURE COMPENSATION: To obtain excellent frequency stability with tuned lines the linear expansion of the line with temperature must be compensated. It has been observed that concentric lines have a temperature coefficient of frequency variation almost equal to the mechanical temperature coefficient of linear expansion for the material of which the line is made.

With a concentric line, so long as both conductors have the same temperature, the ratio of their diameters and, therefore, the electrical constants per unit of length do not change with temperature. To a reasonable degree the change in frequency with change in temperature can be considered as due only to change in length.

In Figure 22 are shown two examples of temperature compensation which operate on the principle that aluminum has a greater coefficient of expansion than copper. Two concentric quarter-wavelength lines are shown. One end is shorted and the open-

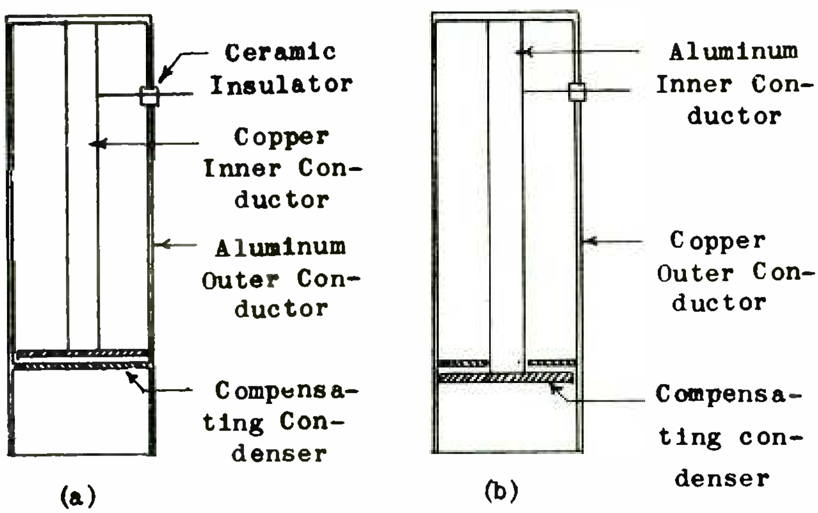


Figure 22

circuited end is shunted by a compensating condenser consisting of opposing plates. In Figure 22(a), if the line temperature rises the aluminum expands more than the copper and so increases the spacing of the plates. This decreases the compensating capacity and tends to increase the resonant frequency of the line, compensating the tendency for the frequency to decrease as the copper inner conductor increases in length. Figure 22(b) is an inverted variation of the arrangement shown in Figure 22(a). Both these arrangements require careful design, construction, and adjustment.

In Figure 23 is shown a more convenient arrangement of the concentric line used to nullify the effects of temperature change. A portion of the inner conductor is made in the form of a flexible metal bellows and the conductor is then maintained at a fairly constant length by a rod of some material, such as invar, having a very low temperature coefficient of expansion. This invar rod is not in the electrical circuit of the line, because it is enclosed by the inner conductor. Since the line's electrical length is determined by the length of opposing surface of the inner and outer conductors, the line frequency will be determined by the *inner conductor's* length.

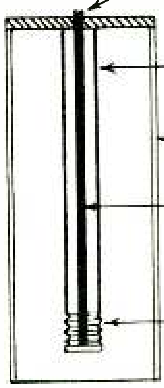


Figure 23

This construction is also particularly well adapted to making exact adjustments of frequency by adjusting the free length of the invar rod to stretch or compress the flexible bellows.

PUSH-PULL OSCILLATORS: In Figures 24 and 25 we are shown two examples of push-pull oscillators employing concentric line tank circuits. In both cases, except for the tuned line grid circuits, the connections are of the usual type for push-pull oscillators.

The oscillator in Figure 24 employs a half-wavelength concentric line open-circuited at both ends. Since the line acts simply as two quarter-wavelength sections connected in series, each being driven at one end and open-circuited at the other end, the voltage distribution will be as shown by curve E. The two conductors may be shorted together at the center of the line and grounded (as shown). This is not necessary, however, and if

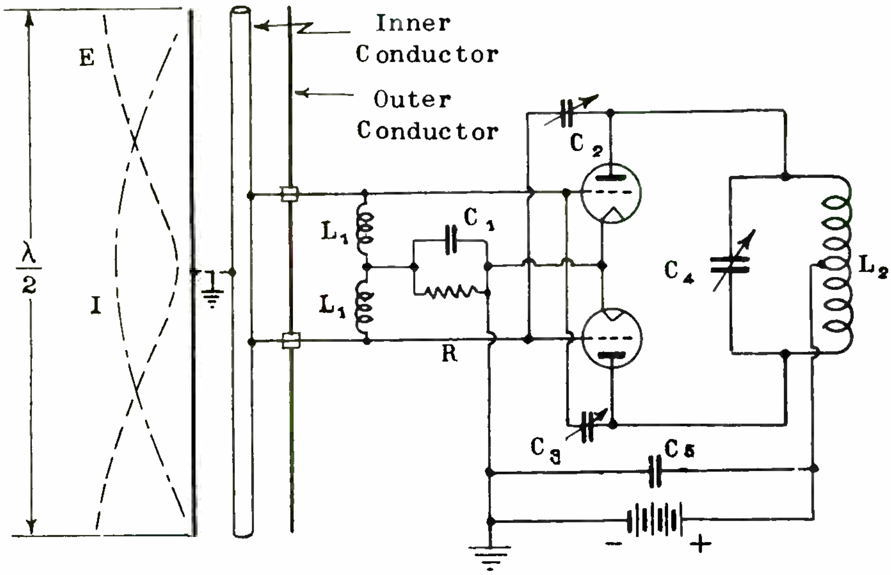


Figure 24

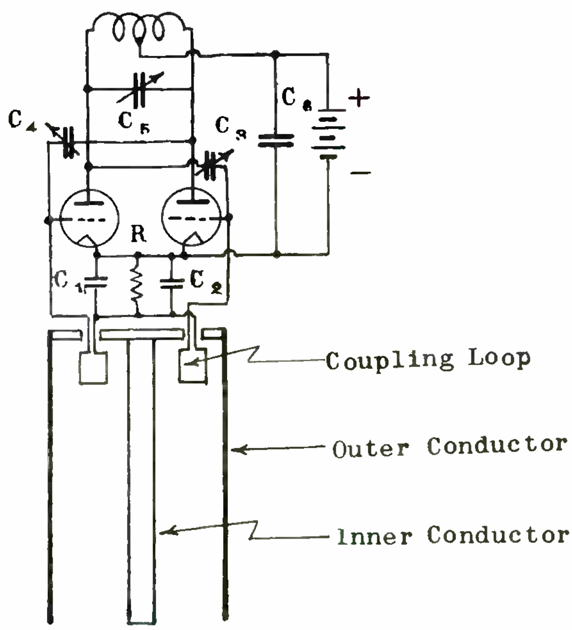
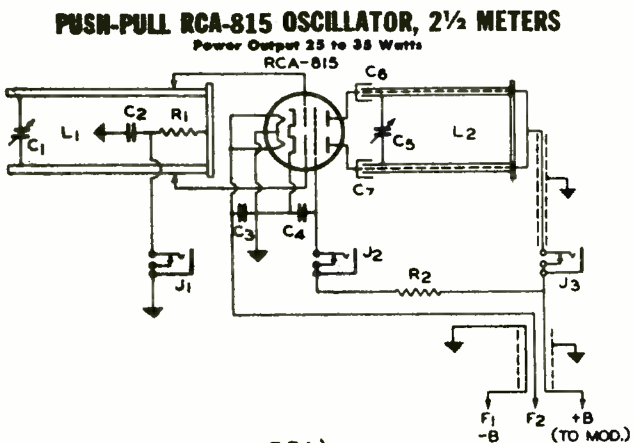


Figure 25

done will require the use of grid condensers to prevent shorting of the d.c. bias developed across R.

Figure 25 shows a push-pull oscillator with grid circuits excited by a single quarter-wavelength concentric line. The grids are inductively coupled to the tuned line by means of the coupling loops shown in the diagram. Note that the connections between one coupling loop and grid are reversed so that the two grids will be driven by voltages 180 degrees out of phase.

In Figure 26 is shown a push-pull oscillator employing two-wire tuned lines in both the plate and grid circuits. (A picture of this oscillator appears in Figure 3.) The oscillator is



C_1 C_5=Line Condensers. 2 Discs of 1/8" Copper 2" Diameter mounted on 10-32 Brass Screws.
C_2 C_3 C_4=0.001 µf mica. 600 v. (Aerovox #1467)
C_6 C_7=Plate line d-c isolation condensers 0.010" copper sheet 1" long wrapped around outside of ½" tubing. 0.002" mica insulation.
L_1=Grid Line. 1" O.D. Copper Tubing 21" long. spaced 1⅞" between centers.
L_2=Plate Line. ½" O.D. Copper Tubing 13" long. spaced ⅞" between centers.
R_1=10000 ohm. 1 watt (IRC #BT-1).
R_2=15000 ohm. 25 watt (IRC #DHA) (Adjust to 14000 ohms).
J_1 J_2 J_3=Meter jacks (Mallory "Midget" #A-2).

ADDITIONAL PARTS

1 RCA-815.
1 Chassis 1/16" aluminum or 16-gauge sheet iron. 4" x 4¼" x 17½".
1 Ceramic socket (National #XC-8).
1 Shielded microphone plug (Amphenol #MCIF and standard microphone plug attachment)
1 Rubber insulating "Boot" (Mueller #29).
NOTE: The various components which have been mentioned by manufacturers' trade names in this unit are the parts that were actually used. Other parts may be substituted with equally good results, provided they have similar characteristics.

(Courtesy RCA)

Figure 26.

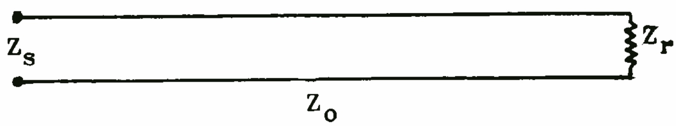
tuned over the 112-116 MC/s amateur band by means of disc-type condensers (C_1 and C_5) mounted near the open end of each resonant line. (These circuit constants are quite similar to those required in the 108-112 MC/s aeronautical band.) In order to remove the d.c. plate voltage from the plate lines, the plates of the oscillator are capacitance-coupled to the plate line by means of condensers C_6 and C_7. The d.c. plate voltage is fed to the plate side of these condensers by means of insulated wires (shown in dotted lines) running through the center of each plate rod. The condensers are made by wrapping and bolting a piece of one inch copper strip around each plate rod at the plate end.

Mica, 0.002 inch thick, is used to insulate the copper strip from the rod.

(The above description is an excerpt from specifications by R.C.A.)

By removing the d.c. voltage from the plate tuned line the possibility of dangerous shocks during tuning or adjustment of this line is avoided. The grid excitation may be altered by adjustment of the taps on the grid line. It will be recalled that standing voltage waves on the two wires of a tuned line are 180 degrees out of phase so that they may be used, as shown, to drive both tubes of a push-pull oscillator. The necessary feed-back coupling is provided by locating a portion of the grid line near the plate line as shown in Figure 3.

TRANSMISSION LINES AS IMPEDANCE TRANSFORMERS: One of the most important relations to be studied in a transmission line is is that of its input impedance Z_s looking into the line at the sending end. This impedance is dependent upon a number of factors--line length (l), characteristic impedance of the line



(Z_o), and the impedance connected at the receiving end (Z_r). For any set of values for these factors the input impedance is found by the general formula:

$$Z_s = Z_o \frac{Z_r + Z_o \tanh j2\pi l/n}{Z_o + Z_r \tanh j2\pi l/n}$$

For any line an *even number* of quarter-wavelengths long this becomes--

$$Z_s = Z_r$$

For any line an *odd number* of quarter-wavelengths long this becomes--

$$Z_s = Z_o^2/Z_r \quad \text{or} \quad Z_o = \sqrt{Z_s Z_r}$$

The general equation is too cumbersome to be of much practical value but the two derived equations are extremely useful and should be studied carefully. It is seen that with any line an *even* number of quarter-wavelengths long, no matter what impedance (Z_r) is terminating the line, the *same impedance will be measured at the input terminals.* On the other hand, if the line is an *odd* number of quarter-wavelengths long *the sending end impedance* Z_s *will always vary inversely with the receiver impedance.*

For the condition where the line is properly terminated ($Z_r = Z_o$) *the input impedance* Z_s *will be equal to the characteristic impedance of the line* Z_o, *no matter what line length is being used.* This can be seen by substituting $Z_r = Z_o$ in either of the two given equations.

These very important principles are made use of in transmission lines operated as impedance transformers. For example, consider the problem of coupling a 73 ohm antenna to a 73 ohm

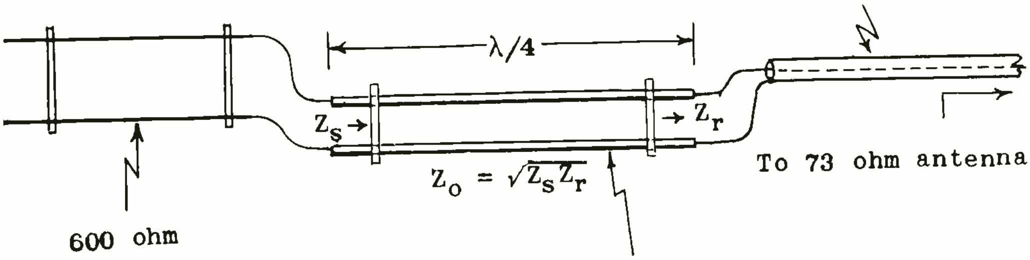


Figure 27

concentric line to a 600 ohm two-wire line. A quarter-wavelength line will be used for an impedance matching section as shown in Figure 27.

Since the 73 ohm concentric line is properly terminated with a 73 ohm antenna impedance, its input impedance will be 73 ohms. But--it is desired to terminate the two-wire line with its characteristic impedance of 600 ohms in order to prevent

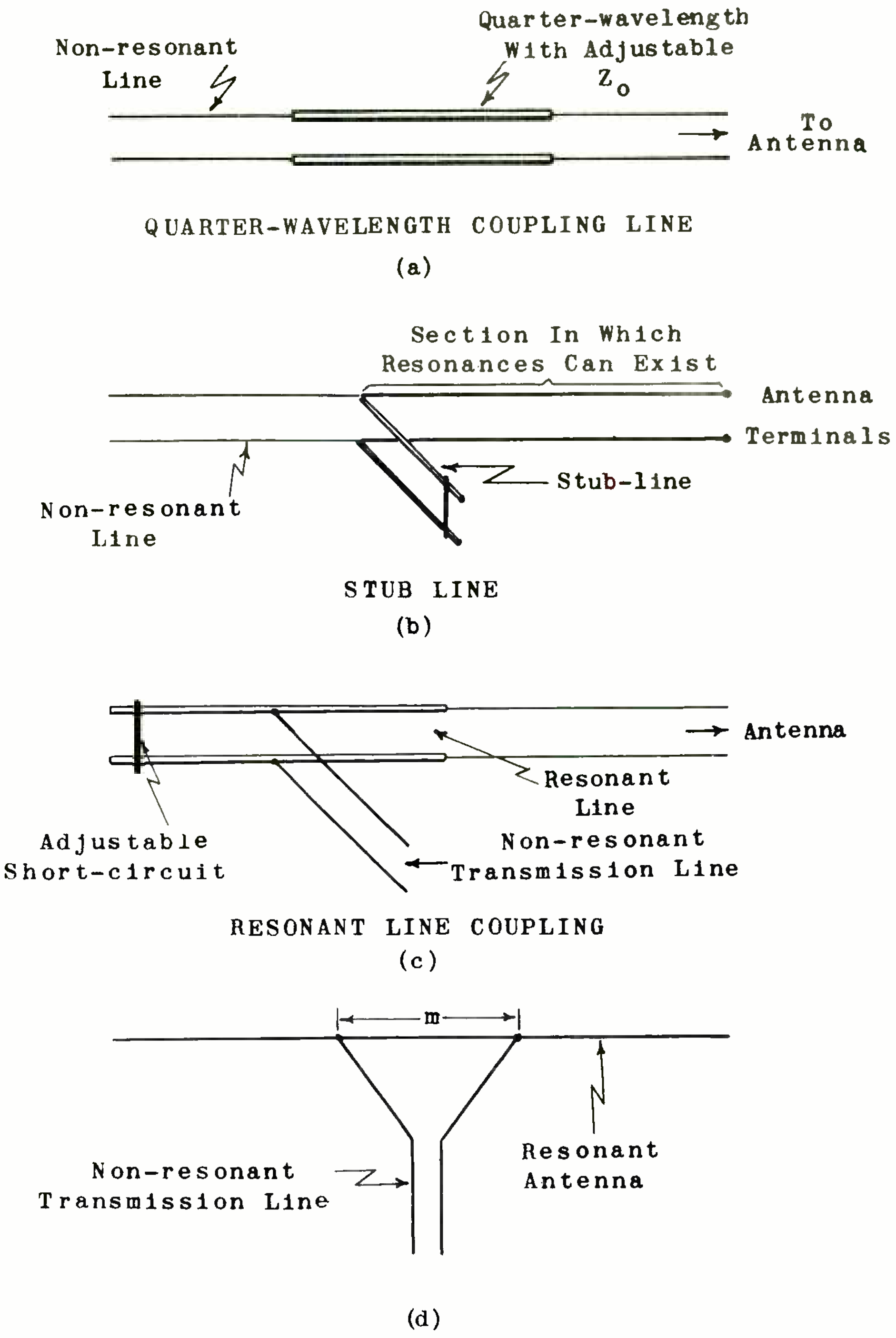

Non-resonant
Line
Quarter-wavelength
With Adjustable
Z_o
To
Antenna
QUARTER-WAVELENGTH COUPLING LINE
(a)
Section In Which
Resonances Can Exist
Antenna
Terminals
Stub-line
Non-resonant
Line
STUB LINE
(b)
Antenna
Resonant
Line
Non-resonant
Transmission Line
Adjustable
Short-circuit
RESONANT LINE COUPLING
(c)
m
Resonant
Antenna
Non-resonant
Transmission Line
(d)

Figure 28

standing waves on the line and to obtain maximum power transfer. Obviously the concentric line could not be connected directly to the two-wire line. Therefore a device is needed which, with 73 ohms coupled to one end, will present an impedance of 600 ohms at the other end. A quarter-wave matching section will do this very satisfactorily. All that is necessary is that the quarter-wavelength line have a characteristic impedance equal to

$$Z_o = \sqrt{Z_s Z_r}$$

$$= \sqrt{600 \times 73} = 210 \text{ ohms}$$

If one inch copper tubing is used for the leads, the spacing center to center for the matching line will be

$$Z_o = 276 \log_{10} b/a$$

$$210 = 276 \log_{10} b/.5$$

$$b = .5 \text{ antilog } \frac{210}{276} = 2.88 \text{ inches spacing}$$

This spacing should be considered only as a first approximation and insulating spacers which hold the $\lambda/4$ line rigid should be so selected that the spacing may be adjusted to slightly more or less than 2.88 inches, as desired. The exact spacing will be that which results in a minimum of standing waves on the two connecting lines.

IMPEDANCE MATCHING SYSTEMS FOR ANTENNAS: In order to avoid resonances in transmission lines associated with transmitters, it is necessary that the antenna be coupled to the line in such a way that the effective load impedance which is offered to the line is a resistance equal to the characteristic impedance of the line. Some typical arrangements using tuned lines are shown in Figure 28.

Figure 28(a) is an example of line transformer action, previously discussed. By proper adjustment of the matching line's length and characteristic impedance the transmission line will be properly terminated. For example, assume that the

73 ohm antenna of Figure 16 is to be fed by a 600 ohm two-wire transmission line, coupled to the transmitter. What should be the proper line dimensions of the matching section at a frequency of 100 MC/s?

At 100 MC/s:

$$\lambda = 300/100 = 3 \text{ meters}$$

$$\lambda/4 = 3/4 = .75 \text{ meter}$$

$$= .75 \times 39.37$$

$$= 29.5 \text{ inches}$$

$$Z_o = \sqrt{Z_s Z_r}$$

$$= \sqrt{600 \times 73}$$

$$= 210 \text{ ohms}$$

A reference to the Characteristic Impedance Chart will show that 210 ohms requires the use of tubing since copper wire gives an impractical line spacing at this low value of Z_o because of the small radius (a) of the wire. Using 5/8 inch copper tubing the separation between the centers of the conductors would be calculated as follows:

$$210 = 276 \log_{10} b/a$$

$$\log b/a = 210/276 = .76$$

$$b/a = \text{antilog } .76$$

$$= 5.76$$

$$b = (5.76)a$$

$$= 5.76 \times \frac{5/8}{2}$$

$$= 1.8 \text{ inches}$$

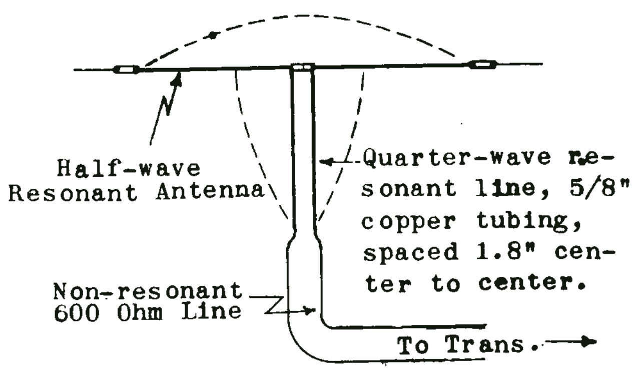


Figure 29

This arrangement would then be as shown in Figure 29. Note carefully the current distribution on both the antenna and the matching lines as represented in the diagram by the dotted lines.

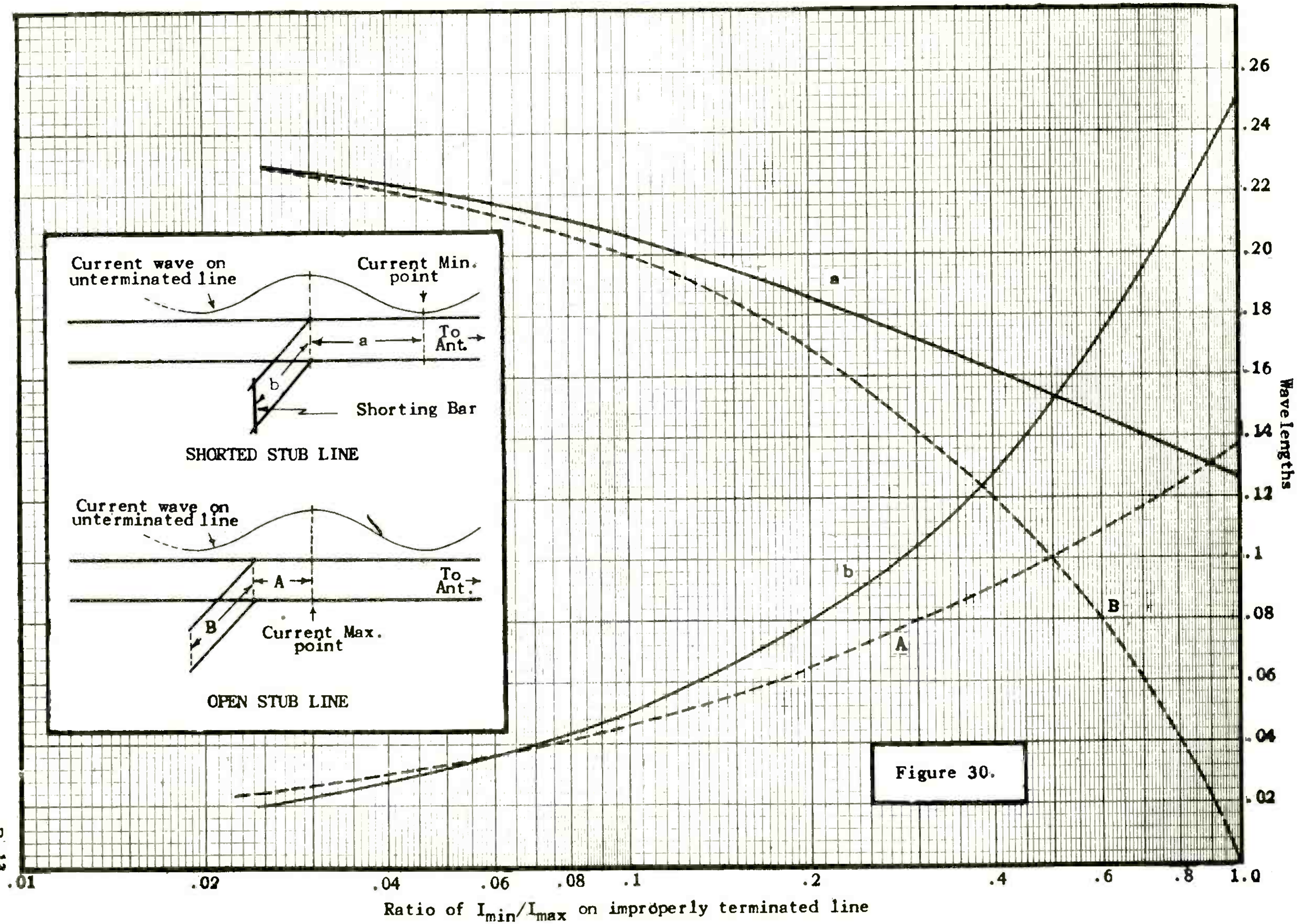

Current wave on unterminated line
Current Min. point
a
b
To Ant.
Shorting Bar
SHORTED STUB LINE
Current wave on unterminated line
A
B
To Ant.
Current Max. point
OPEN STUB LINE
a
b
A
B
.01
.02
.04
.06
.08
.1
.2
.4
.6
.8
1.0
Ratio of I_{min}/I_{max} on improperly terminated line
.02
.04
.06
.08
.1
.12
.14
.16
.18
.20
.22
.24
.26
Wavelengths

Figure 30.

STUB LINES: Another convenient type of matching section is the stub line, shown in Figure 28(b). The transmission line is connected directly to the antenna and a suitable shunt reactance, commonly in the form of a short length of transmission line (stub line), is used to eliminate resonances between the shunting point and the generator.

A transmission line shorter than one quarter-wavelength and open-or-short-circuited at one end presents, respectively, a capacitive or inductive reactance at the other end. If the line is greater than $\lambda/4$ and less than $\lambda/2$, these reactive relations are reversed.

This shunting reactance must be placed at a point such that, when the impedance looking toward the antenna is paralleled by the shunt reactance, the impedance of the combination is a resistance equal in magnitude to the characteristic impedance of the line. The exact length and position of the stub line (subject to minor adjustments) may be obtained graphically from Figure 30.

First the ratio of minimum to maximum current in the standing wave of the improperly terminated line is determined. This may be done as in Figure 7, since only a ratio is desired. The stub line (either open or shorted type) is then placed near the receiving end, its length and position being determined from Figure 30.

For example, with I_{min} and I_{max} as measured below (Figure 31), the position of a shorted stub for matching would be determined as follows:

$$I_{min}/I_{max} = \frac{100}{400} = .25$$

$a = .18$ (From the graph,

$b = .092$ Fig. 30)

It will be noted that when the matching stub is adjusted to the correct length by the shorting bar and is properly positioned, standing waves appear on the line only between the stub and antenna; the remainder of the line is now working into the proper terminating impedance.

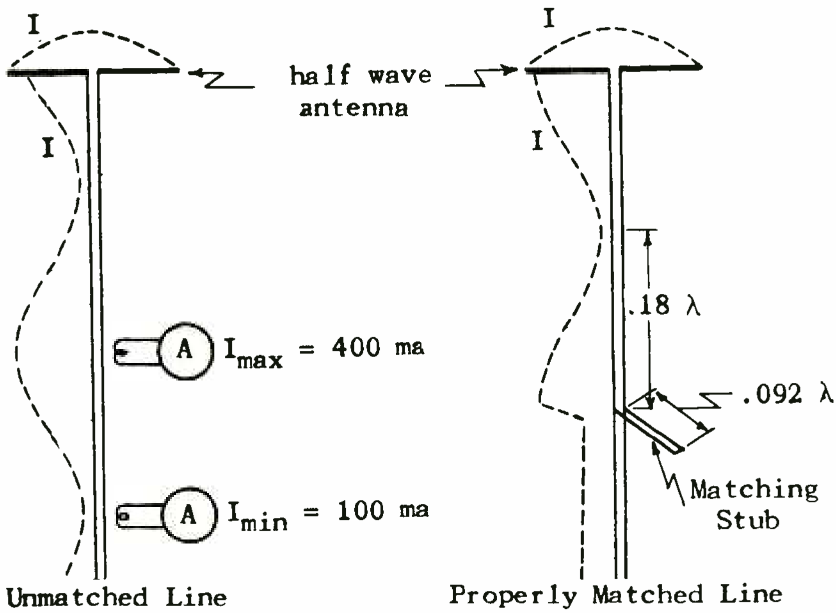


Figure 31

In the resonant coupling line of Figure 28(c) a resistive impedance is obtained by adjusting the short circuit on the coupling line to give resonance in conjunction with the antenna. The resulting load offered the non-resonant line is then resistive and has a magnitude determined by the point of connection. The length of resonant line should be approximately λ/4 when the antenna resistance is higher than the non-resonant line impedance, and a λ/2 when the antenna resistance is lower.

In Figure 28(d) the transmission line is directly coupled to the antenna, the matching being obtained by making the antenna length the exact value required for resonance and then connecting the two wires of the line symmetrically with a "spacing in" such as to give the required impedance match. Another type of antenna matching which functions in a somewhat similar manner

$$N = \frac{148}{fmc/s} \qquad M = \frac{123}{fmc/s}$$

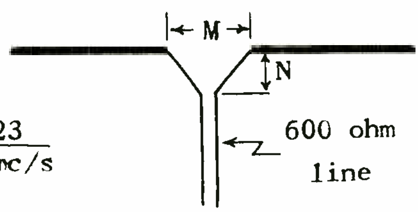


Figure 32

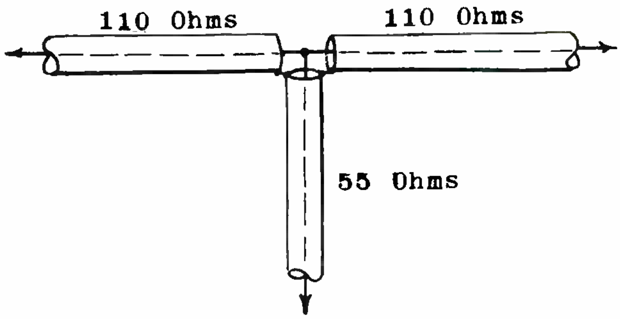


Figure 33.

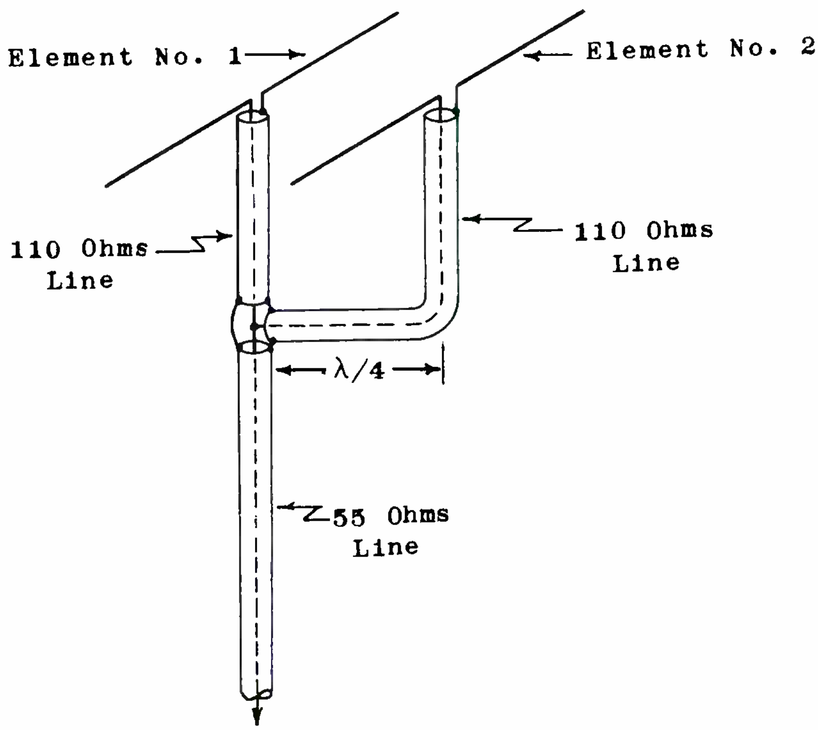


Figure 34.

to the above is shown in Figure 32. Dimensions for matching a 600 ohm line, which is the Z_c most commonly used, are also given.

In each of the matching systems mentioned above, the dimensions arrived at are only approximate and slight adjustments are made experimentally until current (or voltage) measurements show a minimum of standing waves existing on the transmission line.

Frequently in directional arrays it becomes necessary to feed more than one radiating system or element from a common transmitter output. This is done very simply by the use of parallel circuit calculations. For example, assume that the main line from the transmitter is designed for an impedance of 55 ohms. At some point near the radiating array it is to branch into two lines to feed two radiating elements. The two branch lines should be designed for Z_o = 110 ohms and connected in parallel to the main line. This is shown in Figure 33.

In other words, two 110 ohm properly terminated lines connected in parallel have an impedance of 55 ohms just as would any two 110 ohm resistances connected in parallel. As such, they form the proper terminating impedance for a 55 ohm line.

Very often it is desired to feed two elements of an array out of phase. For example, assume that it is required to feed Element No. 2 90^o behind Element No. 1. See Figure 34. This is done by introducing a 90^o lag ($\lambda/4$) in the line which feeds Element No. 2 as compared with that which feeds Element No. 1.

These same principles can be applied to meet any combination of number of elements, impedances of radiating elements, and lines and phase angles between elements. In a more involved network where the desired line impedances for proper transfer of power result in improper termination of the lines by the radiating elements, a matching section as shown in Figure 29 could be used between each radiating element and its line.

POWER AMPLIFIERS EMPLOYING LINEAR TANKS: Resonant lines may be used in tank circuits of amplifiers as well as oscillators at ultra-high frequencies such as those encountered in the aeronautical channels. It will be recalled that the input impedance Z_s of a line an odd number of quarter-wavelengths long is,

$$Z_s = \frac{Z_o^2}{Z_r}$$

Also the input impedance of a line an even number of quarter-wavelengths long is,

$$Z_s = Z_r$$

It will then be seen that theoretically a quarter-wavelength line *shorted* at one end should present a very high or infinite impedance at the open circuit terminals.

Similarly, a half-wave line *open circuited* at one end should present a very high or infinite impedance at the opposite terminals. These conditions are both shown in Figure 35.

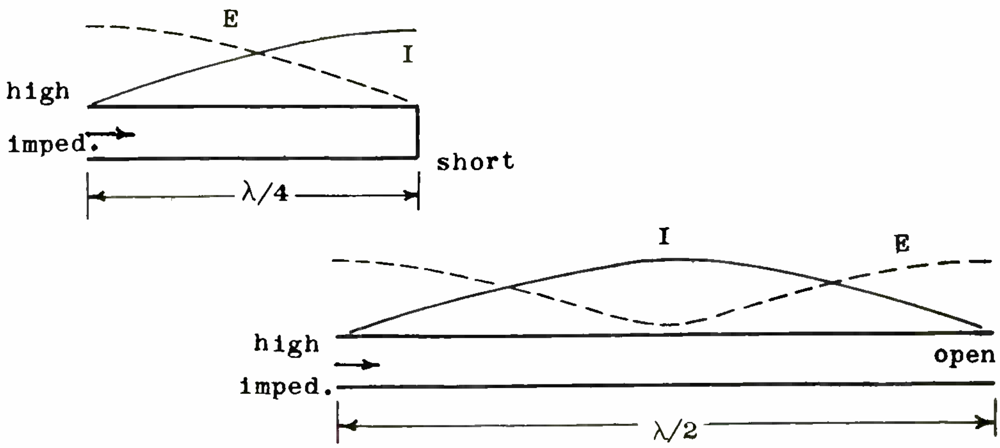


Figure 35

In practice, however, the input impedance obtained under such conditions is never infinite but will be some real value depending upon the line dimensions. Neglecting radiation, the input impedance of a quarter-wavelength line short circuited at its far end or of a half-wave line open at its far end is found to be--

$$Z_s = \frac{2Z_o^2}{Rl}$$

where

Z_o = characteristic impedance in ohms

R = line resistance in ohms per loop meter

l = length of the line in meters

For example, at 100 MC/s a quarter-wavelength, two-wire line using No. 12 B & S spaced 6 inches center to center is shorted at one end. The r.f. impedance looking into the line at the other terminals would then be

Z_o = 600 ohms (from chart)

$$R = \frac{\sqrt{f}}{a} \times 10^{-6} \text{ ohms/foot}$$

(from an earlier formula for two-wire lines)

$$R = \frac{\sqrt{10^2 \times 10^6}}{.04} \times 10^{-6} = .25 \text{ ohms/foot} = .82 \text{ ohms/loop meter}$$

At 100 MC/s a $\lambda/4$ line is approximately .75 meter long

$$Z_s = \frac{2Z_o^2}{Rl} = \frac{2 \times 600^2}{.82 \times .75} = 1.17 \text{ megohms}$$

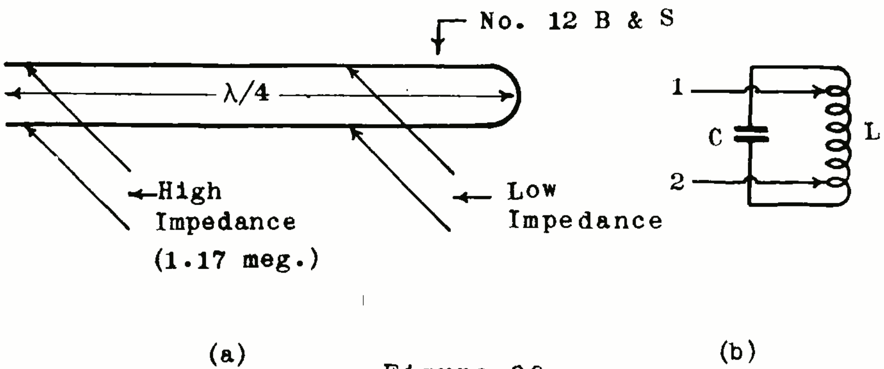
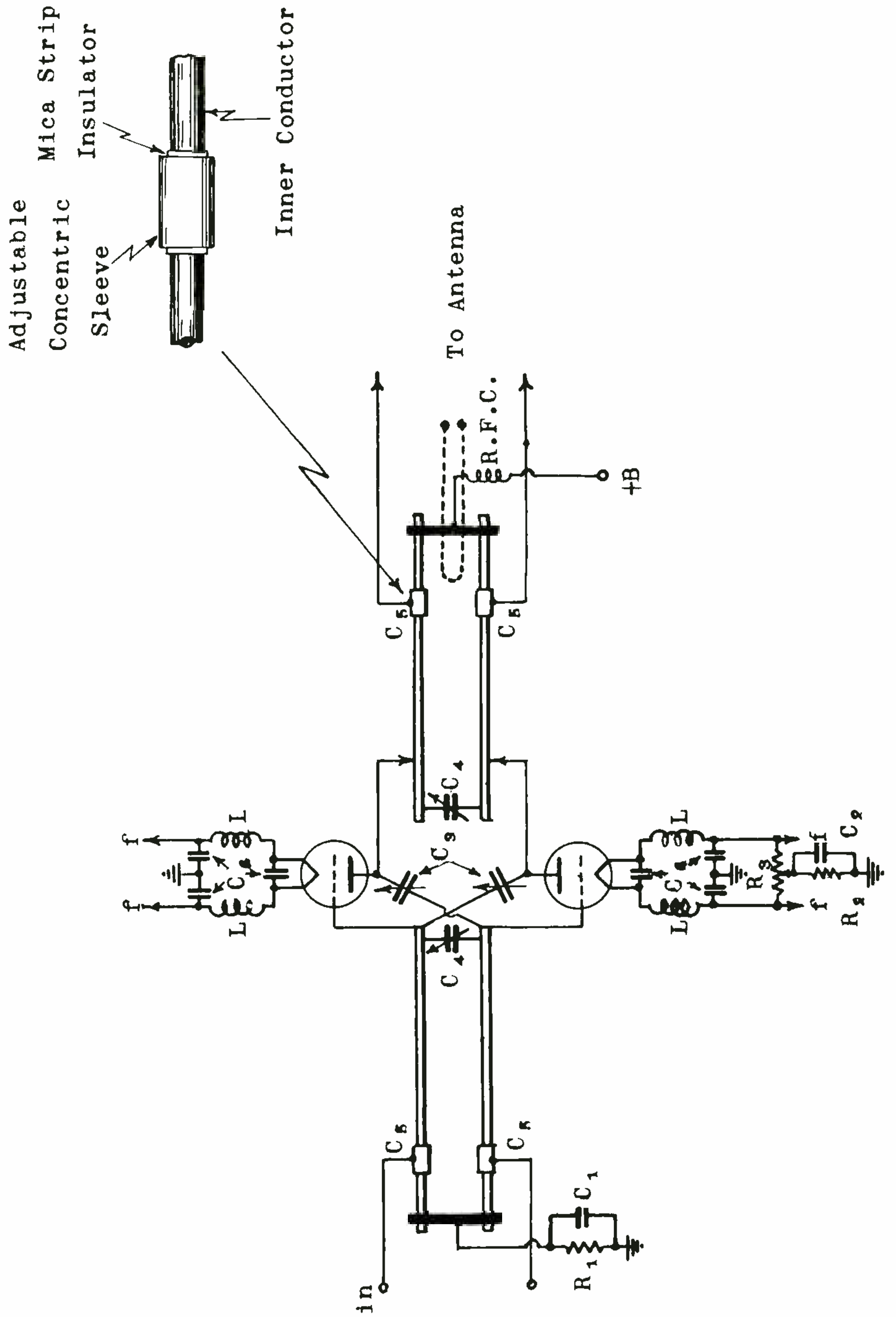


Figure 36



Adjustable Concentric Sleeve
Mica Strip Insulator
Inner Conductor
To Antenna
R.F.C.
+B
in

Figure 37

By means of adjustable taps, as shown in Fig. 36(a), the r.f. input impedance looking into such a line may be varied between zero and the maximum value of 1.17 megohms. This action will be clearly seen if it is recalled that the operation of a quarter-wave line is similar to that of a parallel tuned circuit containing L and C such as shown in Fig. 36(b). In this circuit the impedance between terminals 1 and 2 at resonance will be a pure resistance and maximum in value when the taps are placed at opposite ends of the coil. As the taps are moved toward the center of the coil, the impedance between terminals 1 and 2 becomes less and less, approaching zero.

The behavior of a quarter-wave line in a manner similar to a parallel tuned circuit is made use of in U.H.F. amplifier circuits. For example, in Fig. 37 is shown the schematic diagram of a typical U.H.F. power amplifier. Both resonant lines shown are quarter-wavelength sections. Tuning is accomplished by means of adjustable shorting bars and trimmer condensers C_4. It will be noted that the plate tap positions are adjustable to enable the tubes to work into the correct load impedance. Coupling to the antenna (or next stage) is accomplished by capacitive taps C_5. Link coupling may be used if desired. The grid driving power is adjusted by means of taps C_5 in the grid circuit. Grid leak and cathode bias are furnished by R_1 and R_2 respectively.

Of particular interest in this amplifier is the tuned filament circuit. The coils (L) in the filament circuit are frequently required at the ultra-high frequencies to compensate for the effects of the inductance of connecting leads, which in many cases are long enough to cause an appreciable phase shift, reducing the amplifier efficiency. The effective length of the filament circuit to the points of connection (f-f) should be approximately .5 wavelength to bring the filament to the same potential as the shorted ends of the line. The proper inductance must be determined by experiment, the coils being adjusted until optimum stability and power output are obtained.

At the higher frequencies where .5 wavelength is not excessively long, a straight line may be used to replace L. The by-pass condensers (C_6) should be small in size to reduce lead inductance; 500 $\mu\mu f$ is a satisfactory value.

The use of half-wave lines in the filament circuits of the amplifier is a useful device, employed extensively with connecting leads at the higher frequencies. It has already been shown that a line approximately .25 wavelength open-circuited at one end will appear as a short-circuit at the other terminals. It would not do, for example, to have interstage leads in a transmitter which were .25 wavelength long; for then the driver stages output would be effectively short circuited even though the leads were connected to the fairly high grid-circuit input impedance of the next stage.

As shown in Figure 38, if it becomes necessary for x to be

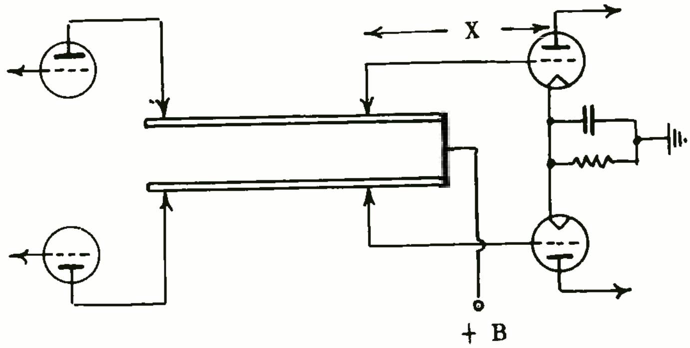


Figure 38

as long as .25 wavelength, that dimension would then be increased to .5 wavelength since *the input impedance of a half-wave line is equal to the terminating impedance.*

EXAMINATION

1. (a) Why are ultra-high frequencies required for F.M. broadcasting?

   (b) Why does F.M. permit closer geographical spacing of the broadcast stations?

2. Discuss the use of a transmission line at u.h.f. for purposes other than that of transmitting energy.

3. How are current and voltage loops (maxima) detected on a transmission line?

4. Given a two-wire line having the following operating features: Size of wire - No. 12; separation - 4 inches; frequency = 75 m.c.

   (a) Find L, C, and R per foot.

   (b) Calculate the characteristic impedance Z_0, and check by means of the chart.

5 (a) What are the advantages of a concentric line over a two-wire line?

   (b) Given a concentric line that is to have a characteristic impedance of 75 ohms. The inner conductor is a No. 10 copper wire. What should be the inner diameter of the outer shell?

6. (a) What is the attenuation in db. per *foot* of the two-wire line given in Question 4?

   (b) What is the attenuation in db. per foot of a coaxial cable having a No. 12 inner wire, and a characteristic impedance of 75 ohms, at 75 m.c.?

7. Given a concentric line having the following dimensions: b = 2.5 in., a = .75 in.. It is one-quarter wave length long, and shorted at its far end. The operating frequency is 200 m.c. Assume the wave velocity is .95 that of light,

7. (a) Find R, the resistance per loop meter of the line.

(b) Find the actual wavelength, λ_o.

(c) Find the characteristic impedance, Z_o.

(d) Find the Q of the line.

8. (a) What advantage has a resonant line over a crystal for stabilizing the frequency of a u.h.f. oscillator?

(b) Give a representative oscillator circuit employing a resonant line.

(c) Show two methods by which temperature compensation may be provided for a resonant line.

9. (a) Design a quarter-wave length two-wire line that will match a 600 ohm line to a 200 ohm load. Use 1/4 inch copper tubing for the matching line.

(b) A transmission line feeds an antenna, and it is desired to eliminate standing waves on the greater portion of the line by means of a matching stub. The current minimum is one-half of the current maximum. The operating frequency is 100 m.c., and the wave velocity is .95 that of light. What length of *shorted* stub is required, and how far along the line must it be located from the current maximum point. Is the location toward the load or toward the source?

10. Given a quarter-wave length concentric line for which b = 2.5 in. and a = .75 in., and which operates at a frequency of 200 m.c. The far end is short-circuited. What is the impedance looking into the open end of the line? *Note:*--Wave velocity is to be taken as .95 that of light.