PROPERTY OF
M/SGT. WALTER F. PUDLIK

PHILCO

TECHREP DIVISION BULLETIN

1. magnetron.

MAY
1953

PHILCO
TECHREP DIVISION
BULLETIN

Published Monthly by

The TechRep Division of Philco Corporation

Philadelphia, Pennsylvania

Volume III — MAY, 1953 — Number 5

CONTENTS

Editorial 2
By John E Remich

Basic Magnetron Operation 3
By John E. Marchesano

What's Your Answer? 18

Simultaneous Transmission Systems
(Part I of Two Parts) 21
By Francis R. Sherman

Solution to Last Month's "What's Your Answer?" . . 37

Transmitter Monitor 38
By Harold Gullstad

An Emergency Teletype Relay System 43
By Perry S. Gaye

Editor
John E. Remich

Managing Editor
Robert L. Gish

Technical Editors
Francis R. Sherman
Gail W. Woodward

EDITORIAL OFFICE:
TECHREP DIVISION
PHILCO CORPORATION
22nd and LEHIGH AVENUE
PHILADELPHIA 32, PA.

If any information contained herein conflicts with a technical order, manual, or other official publication of the U.S. Armed Forces, the information in the official publication should be used. Views expressed in this publication are the opinions of the authors and do not necessarily represent those of Philco Corporation. All technical information and circuits are presented without responsibility as to patent infringement.

Editorial

THE IMPORTANT JOB

by John E. Remich, Manager, Technical Department

It is an unfortunate, but unmistakable, fact that regardless of the over-all importance of any job, there are some phases of every job that do not require the full abilities of the men performing them. Unfortunately, these routine details are often not recognized as the first steps to more important jobs. Everyone likes to feel that he is being fully utilized, and nearly everyone likes to do a job that is important; yet, often the most important job (from a management point of view) is the job which the man filling it believes to be the least important. Even more often, the important job is not the most glamorous one in sight, nor the most enjoyable.

In any profession, and particularly in field engineering, the man who wants to get ahead must be willing to accept the less glamorous job, when required, as well as the ones with the "sparkle." Frequently, a new man feels that he is not being properly utilized, that the requirements of his job are not sufficiently demanding, and that his abilities and potential value are going unnoticed and unappreciated by his supervisor simply because he is not being given an opportunity to display them. Although this may actually be the case in rare instances, it is almost invariably true that such individuals are placed in what they believe to be unimportant jobs, deliberately, for planned periods of indoctrination, familiarization, and testing. Many a man has passed up his opportunity for a far more important job because he formed a quick decision on the basis of his first few weeks on the new job.

In short, any individual who wants increased responsibilities must be free of any allergy for hard work or occasional dull chores.

BASIC MAGNETRON OPERATION

by John E. Marchesano
Philco Field Engineer

The basic theory of magnetron operation, maintenance, and testing.

WITH ENGLAND'S ENTRY into World War II on September 3, 1939, the need for an accurate method of detecting and locating enemy air and surface craft became imperative. The searchlight system used at that time in a chain along the coast of England was impressive, but it had the serious disadvantage of being limited to the range of human vision. It was believed that radio detection and location using a narrow beam of radio-frequency energy to scan the surrounding area would be the ideal system. However, with the radio frequencies then in use, the antenna system required to produce a narrow beam would have necessitated an immense radiating structure of several thousand cubic yards, installed at the top of a very tall tower. The engineering problem of swinging such a system about in various directions would have been formidable.

It was soon realized that the solution to the problem of building an antenna of a practicable size depended upon the development of an oscillator capable of producing usable amounts of r-f power at extremely short wavelengths. The shorter the wavelength, the smaller the antenna could be, and the more practicable would be the installation. In fact, if the wavelength were short enough, installations in aircraft would also be practicable, and the door to many other possible applications, both in peace and in war, would be opened. All of these factors were the incentive for the development of the oscillator known as the magnetron.

THE MAGNETRON

The basic theory of magnetron operation was first introduced by Hull*, in 1921. He published a mathematical analysis of an arrangement which consisted of a cylindrical anode and an axial filamentary cathode with a relatively strong magnetic field parallel to the cathode, as shown in figure 1. Many others became interested in the magnetron and a variety of arrangements were investigated. These investigations indicated the possibility of using the magnetron for the generation of high frequencies of considerable power. The sole drawback was that the efficiency of operation was low, being on the order of 10 to 20 percent near 10 cm. With the impetus of wartime need, the Brit-

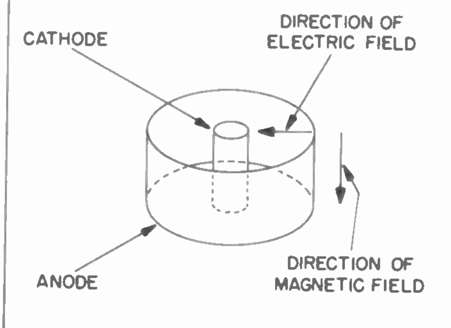


Figure 1. Basic Magnetron Arrangement

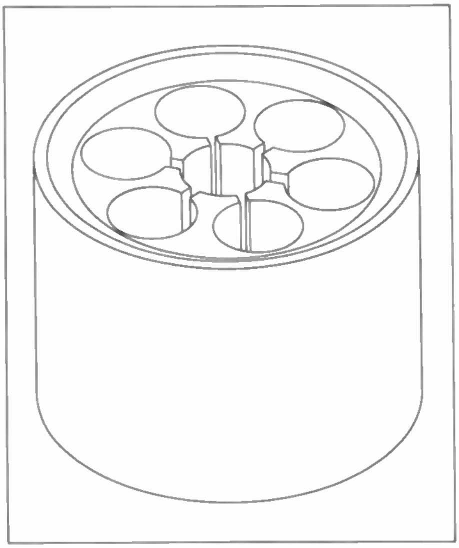


Figure 2. Anode Block of the First British 10-Cm. Magnetron

ish, late in 1939, developed a magnetron for radar use capable of very high pulsed-power outputs at wavelengths of 10 cm. or less. The magnetron had internal resonators which, when pulsed, produced microwave radiation. A drawing of the first anode block used by the British is shown in figure 2. Modifications in the design increased the efficiency of operation to about 50 percent.

This article will cover the operation of the class of magnetrons that employ internal resonant systems of the type shown in figure 2.

REVIEW OF BASIC PRINCIPLES

Before a physical description of what occurs in a magnetron is presented, a few fundamental principles that are involved in magnetron operation will be reviewed. Consider first a simple diode that is connected in series with a battery, as shown in figure 3. With the cathode heated, electrons will pass from cathode to anode in the diode. However, instead of being concerned with the passage of a great many electrons, it will be best to consider the effect of each electron. Assume, therefore, that one electron has just left the cathode surface. Since it is in the electric field existing between the anode and cathode, it will start accelerating toward the anode in exactly the same manner that a freely falling body accelerates in the earth's gravitational field. As the negatively charged electron approaches the anode, it induces a positive charge on the anode. This electrostatic induction by the approaching electron causes electron motion in the external circuit, such motion constituting a current. When the electron strikes the anode, it no longer induces electron motion, and therefore current ceases to exist. The kinetic energy acquired by the electron is dissipated in the form of heat and secondary emission at the anode surface. It is important to note that current existed only while the electron was in motion from cathode to anode, and that it ceased when the electron struck the anode. Also, if the electron had proceeded toward the anode, but had reversed direction by some means before it struck the anode, the current induced would have flowed first in one direction and then in the other. Thus, the direct passage of an electron to the anode induces a direct current, while the passage of an electron toward, and then away from, the anode induces an alternating current.

Consider next the motion of an electron in a magnetic field, as shown in

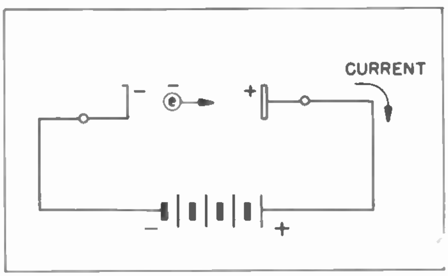


Figure 3. Diode Circuit to Indicate the Effect of Electron Motion

---

* Hull, A. W. "Phys. Rev.," pp. 18, 31, 1921.

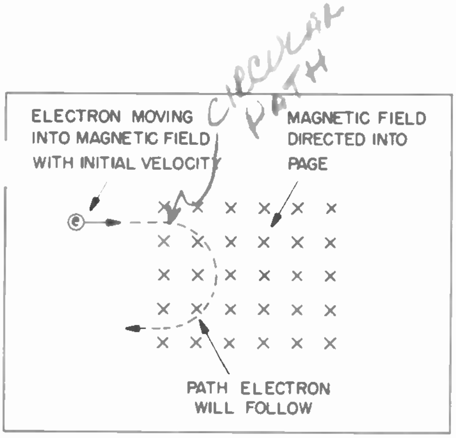


Figure 4. Path of Electron Projected into Magnetic Field Oriented Perpendicularly to Electron's Initial Path

figure 4. Assume that the electron enters at right angles to the magnetic field with a certain initial velocity. The interaction of the moving electric charge and the magnetic field causes the electron to experience a force which is always at right angles to the direction of motion. Thus, the electron follows a circular path, just as does a weight attached to the end of a string and rotated about one's head. The rotating weight also experiences a force (centrifugal) which is always at right angles to its direction of motion.

If the electric and magnetic fields are combined, as in figure 5, and it is assumed that an electron starts from rest at the plate on the left, another type of motion is obtained. As soon as the electron starts to accelerate, under the influence of the electric field, it experiences a force at right angles to its direction of motion because of the magnetic field. The kinetic energy received by the electron as it is moving in a direction opposite to the electric field is given up when the electron curves to a direction which is with the electric field. Thus, the electron finally comes to rest, and is again accelerated by the electric field and completes another cycle. The resulting motion is exactly described by a cycloidal curve, as shown in figure 5. The same motion is obtained if one traces the path of a point on the rim of a rolling wheel. Actually every point on a rolling wheel describes a cycloidal path, and it can be shown that all points, from the axle on out to the rim, describe motions that electrons with different initial velocities would take in combined electric and magnetic fields. The important facts to note here, however, are that the electron drift is at right angles to the direction of the electric field, and that the electron started with zero velocity and after each cycloidal cycle again had zero velocity.

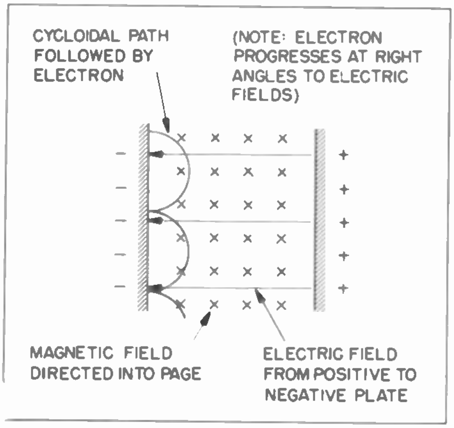


Figure 5. Motion of an Electron in Combined Electric and Magnetic Fields

EQUIVALENT CIRCUIT OF THE CAVITY MAGNETRON

The tuned circuits in the cavity magnetron (see figure 2) may be in the form of holes and slots cut in a solid beryllium-copper block. Consider one hole and slot apart from the others. See part A of figure 6. The slot may be regarded as a capacitor, since most of the electric field is located in it, and the hole as a single-turn coil. The combination of the two constitutes a resonant circuit.

Two other forms of resonant circuits used in cavity magnetrons are shown in Parts B and C of figure 6. The main difference found in various cavity shapes is the ratio of L to C, and hence

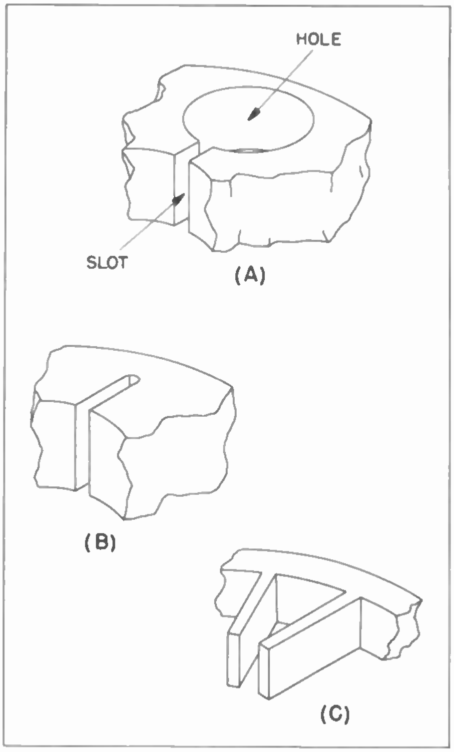


Figure 6. Various Magnetron Cavity Types
A. Hole and Slot
B. Slot
C. Vane

~~the Q of the~~ cavity. Qualitative operation of the various types ~~is identical~~.

The equivalent circuit of the magnetron as a whole is more complicated than just simple parallel-resonant circuits representing each hole and slot. First, there is mutual induction between holes, since the flux emerging from one hole passes through an adjacent hole to form a closed loop. Second, there are capacitances to the end plates and to the cathode, which result in capacitive coupling between cavities. Also, there is coupling through the space charge itself. However, for the purpose of this article, it is not necessary to consider all of these factors in the equivalent circuit. An equivalent circuit does not have to be an exact representation of the circuit, as long as it gives results which closely approximate the true operation. For example, in determining the operation of an amplifier over a wide band of frequencies, three different equivalent circuits are employed—one for the low frequencies, one for the middle frequencies, and one for the high frequencies. None of these alone would give a true picture of the over-all amplifier operation. In the case of the magnetron, if the effect of inductive and capacitive coupling between cavities is neglected, an equivalent circuit can be drawn which is useful in demonstrating the operation of the actual circuit, even though it is not sufficiently accurate for quantitative calculations. Thus, for the purpose of this article, the equivalent circuit of a six-cavity magnetron can be represented as shown in figure 7.

CONDITIONS FOR OSCILLATIONS

In discussing the operation of a six-cavity magnetron, which uses an anode block similar to that shown in figure 2 it will be advantageous for the sake of simplicity to assume that the magnetron cathode and anode are continuous and parallel as shown in figure 8. The power

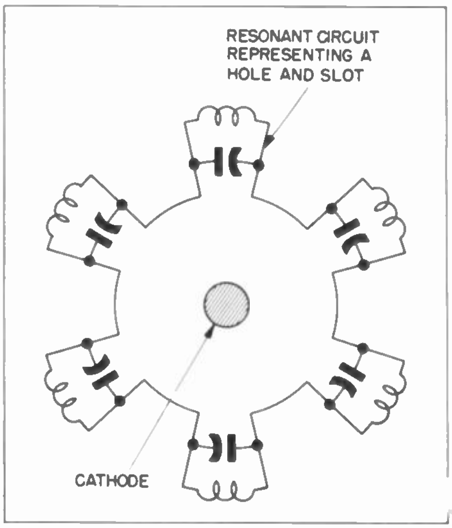


Figure 7. Equivalent Circuit of a Six-Cavity Magnetron

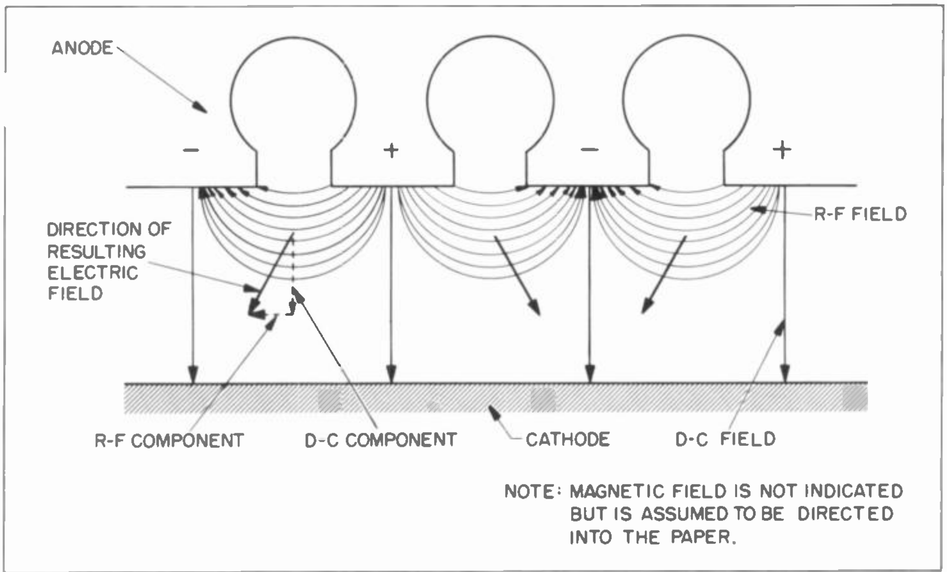


Figure 8. Simplified Drawing of a Cavity Magnetron, Showing Electric-Field Configuration

for the magnetron may be obtained from a direct-current source which is capable of supplying a heavy current (on the order of amperes) at a voltage of several kilovolts. The magnetic field may be obtained from either an electromagnet or a permanent magnet. Assume that the cavities have been shock-excited into oscillation by the abrupt application of a large voltage from the power supply, and that at the instant under consideration the voltages resulting from the oscillation are at their peak values and cause the cavity segments to have the polarities indicated. Under these conditions, there are two electric fields present in the region of each cavity: that due to the d-c voltage of the power supply, and that due to the cavity oscillations. The resulting field is the vector sum of the two, and is in the direction indicated.

In order to understand how the cavity oscillations are sustained, it is necessary to consider the action of the electrons in their passage from the cathode to the anode. It has already been shown that as an electron approaches an anode it will induce a positive charge, and as it recedes from the anode it will induce a negative charge. Thus, if an electron in the interaction space (space between cathode and anode) of a magnetron were to approach an anode segment as the segment is going positive (due to the resonance of the cavity), it would aid oscillations by making the segment more positive. If an electron were to recede from an anode segment as the segment is going negative, it would again aid oscillations by making the segment more negative. However, if an electron were to approach an anode segment as the segment is swinging negative, or recede from the segment as the segment is swinging positive, it would tend to stop oscillations. Therefore, for oscillations to be sustained, the conditions must be such that the electrons which tend to stop oscillations are eliminated and the others are permitted to remain. How this is effectively accomplished in the magnetron may be seen from the following explanation:

First, consider the path of an electron which is aiding oscillations and has

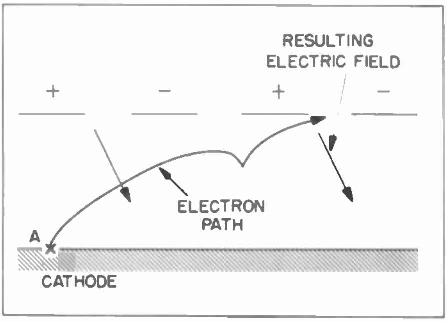


Figure 9. Path of Electron Which Is Aiding Oscillation

started from point A in figure 9. The electron will follow a cycloidal path and progress toward the anode in a direction at right angles to the existing field. Since the electron chosen started its motion under an anode segment that is positive, it is moving in the correct direction to cause it to aid oscillations. The time the electron takes to complete half of its cycloidal cycle, and the distance it progresses in this time, are adjusted by the proper choice of magnetic and electric fields to be such that when the electron is approaching the next anode segment, that segment is swinging positive. Hence, this anode segment is driven more positive, and the oscillations are sustained. This aiding condition occurs each time the cavity completes a half cycle of oscillation in the same time it takes an electron to move from one segment to the next. The electron ends one cycle of its cycloidal motion at the third anode segment, simultaneously returning to zero velocity. It again starts its motion toward the third anode segment while that segment is positive, continuing to aid oscillations until it finally reaches an anode, whereupon it has no further effect on the oscillations. The kinetic energy given up by an electron when it strikes an anode segment constitutes a loss which is supplied by the d-c source. Thus, for oscillations to be sustained, the average progression (from left to right in figure 9) of an electron must be such that the electron passes two anode segments in the time required for the cavity to complete one cycle of oscillation.

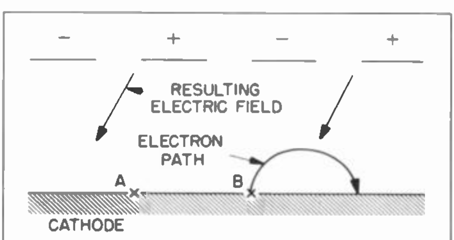


Figure 10. Path of Electron Which Is Opposing Oscillation

Next, consider the path of an electron which would oppose oscillation. Such an electron would leave a point on the cathode (point B in figure 10) at a time when the polarities are as indicated. The electron is starting its motion in such a direction as to oppose oscillation since it is approaching an anode which is swinging negative. However, the resulting electric field is in a direction which does not allow the electron to progress toward the anode segments, but instead causes it to retrogress toward the cathode, since the electron motion will be at right angles to the electric field. Thus, this electron quickly returns to the cathode and has little effect on stopping the oscillations. The kinetic energy given up by the electron when it strikes the cathode also constitutes a loss which is supplied by the d-c source, and is in the form of heat and secondary emission. The net effect, then, is that oscillations are sustained because those electrons which are aiding oscillations are more effective than those which are opposing them.

It can be seen from the foregoing considerations that one of the factors involved in causing the magnetron to function properly is the choice of electric and magnetic field strengths that will cause the electrons to follow optimum paths. However, these considerations are remote from the objectives of this article and will not be considered. The fact that a field configuration can ist, which causes the electrons to fol- .. optimum paths, was assumed rather than developed; however, such a configuration is obtained in actual practice.

MODES OF OSCILLATION IN A MAGNETRON

In the preceding explanation, it was assumed that the cavities of a magnetron operate as tuned circuits. The coupling between cavities by magnetic and electric fields was neglected, so as not to complicate the development of how a magnetron works. However, the simplified equivalent circuit is not useful in enabling one to predict the operation of a magnetron under any but the special conditions for which the assumptions were made. Thus, while the equivalent circuit does not indicate such a possibility, it turns out that a magnetron can oscillate at more than one frequency. The different frequencies of oscillation are spoken of as modes of oscillation, and are designated by mode numbers. The mode number is determined by dividing 2π into the number of radians of phase shift that exist around the inside surface of the anode block. The determination of the mode number for different modes of oscillation will be exemplified in the following discussion.

The simplest and most commonly used mode of oscillation occurs when alternate segments of a magnetron are at equal and opposite r-f potentials, as shown in figure 11. This condition was the one assumed in the explanation of magnetron operation. It was shown that for oscillation to be sustained, it is necessary for the electron to pass two anode segments in the time it takes the cavity to complete one cycle. To determine the mode number for this case, it is necessary to determine first the phase shift in radians existing around the inside surface of the anode block. Since there is a phase shift of π radians (180 degrees) from one segment to the next, there is a total phase shift of 6π radians completely around the block. Therefore, the mode number is equal to 6π divided by 2π, which results in a mode number of 3. This mode of oscillation is also commonly referred to as the "π mode," because of the shift of π radians between segments. It should be noted that the mode number depends not only upon the phase shift between segments but also upon the number of cavities contained in the magnetron. For example, if a magnetron has eight cavities and is operating in the π mode, a total phase shift of 8π radians exists around the anode, giving a value of 4 for the mode number as compared to 3 for the six-cavity magnetron. Correspondingly, a 10-cavity magnetron operating in the π mode has a mode number of 5.

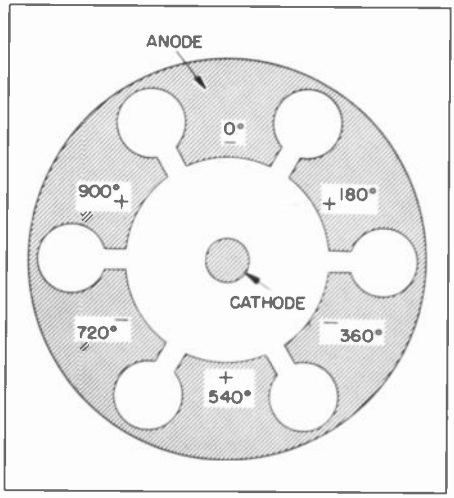


Figure 11. Six-Cavity Magnetron Operating in π Mode, or Mode 3

The phase shift between adjacent segments does not have to be π radians but can be any value, provided only that the total phase shift around the anode block is a multiple of 360 degrees. Thus, in a six-cavity magnetron, oscillation can be sustained when there is a phase

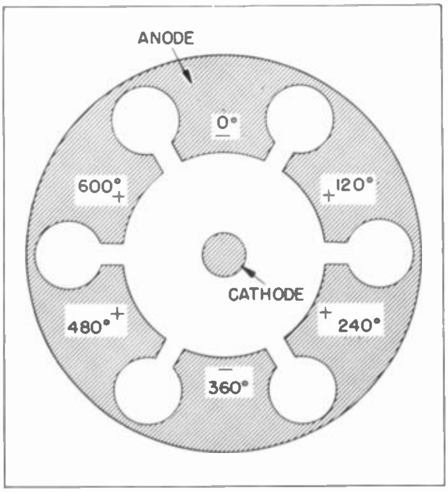


Figure 12. Six-Cavity Magnetron Operating in Mode 2

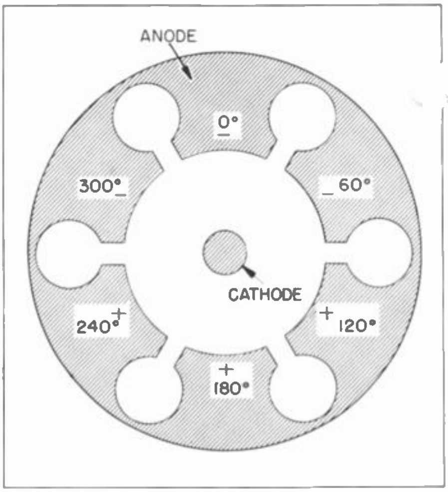


Figure 13. Six-Cavity Magnetron Operating in Mode 1

shift of 120 degrees between adjacent segments. Such a condition exists when the electron passes three anode segments per cycle of oscillation. The mode number for this case is found by dividing 4π by 2π, which gives a value of 2. The magnetron is therefore said to be operating in mode 2 (figure 12).

If the phase shift between segments is 60 degrees, the electrons in the interaction space pass six anode segments per cycle of anode oscillation. The mode number for this case is found by dividing 2π by 2π, giving a value of 1 for the mode number. This condition is shown in figure 13.

MAGNETRON STRAPPING

The fact that a magnetron can oscillate at more than one frequency is a disadvantage in an application such as radar, since the equipment is usually designed to operate at only one frequency. Hence, changes in frequency and simultaneous operation on more than one mode must be prevented. This is accomplished by the use of strapping, which not only decreases coexisting modes both in number and in magnitude of oscillation, but also increases the efficiency of operation. In strapping, alternate segments are connected with short conductors, as shown in figure 14. It can be seen that making these connections will favor the π mode, since the short circuits through the straps will place alternate segments at the same potential (corresponding to a phase shift of 2π radians). Oscillation in mode 2 is not favored by the connections because this mode requires a phase shift of $4/3\pi$

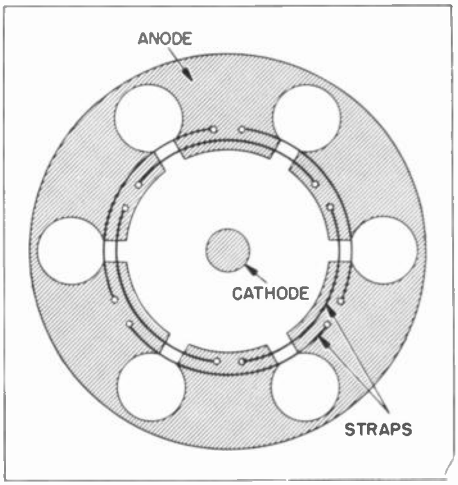


Figure 14. Strapping Arrangement in a Six-Cavity Magnetron (Double-Ring Strapping)

radians between alternate segments—an impossible condition when the segments are short-circuited.

However, the straps cannot be considered as simple short circuits if one is interested in any but the simplest explanation. Each strap, being a conductor, has inductance, and, since it passes near segments between those which it connects, capacitance also is present. Thus, the use of straps will alter the frequency at which a particular mode arises. In the π mode, both ends of the strap are at the same potential and carry no current, except for a small capacitive component. Therefore, the strap inductance is negligible. The capacitive component, however, changes the frequency of the π mode.

If the magnetron is excited in a different mode, the straps will carry more current because of the potential difference across them. This increased current increases the inductive and capacitive effect of each strap, and thereby causes a greater change in the operating frequency of the other modes. Thus, mode separation is accomplished by the addition of straps, and the possibility of coexisting modes is minimized.

The suppression of all but the desired mode (normally the π mode) markedly increases the efficiency of the magnetron. This fact becomes evident if one considers that the load into which a magnetron works must be matched to the magnetron for most efficient power transfer. This matched condition would be different for different frequencies of oscillation. In addition, the unwanted modes usually produce frequencies that are not within the normal response of the radar receiver, and thus constitute a total loss.

At first, symmetrical strapping arrangements, as shown in figure 14, were used to discriminate against unwanted modes. However, it was found that a

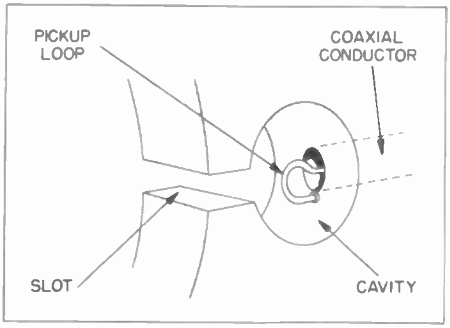


Figure 15. Common Method of Coupling Energy from Magnetrons Operating at Wavelengths Longer than 3 Cm.

certain measure of asymmetry of the strapping arrangement could be employed to discriminate still further against unwanted modes. This led to the omission of certain straps, particularly the straps across the resonant element in which the output coupling is situated, and those immediately adjacent to the input cathode. In certain cases, other straps are also omitted, but the ultimate strapping pattern is determined, at present, by experiment.

COUPLING

Energy is usually taken from a magnetron by means of a coupling loop. For magnetrons operated below about 3 cm., the coupling loop is formed by bending the inner conductor of a coaxial line into a loop and soldering the end to the outer conductor, as shown in figure 15.

At wavelengths shorter than 3 cm., it is not practicable to make loops small enough to be located within the cavity. Instead, the coupling loop is located at one end of the cavity, as shown in figure 16. In both cases, the loop provides inductive coupling. Some other coupling methods which are frequently used for higher-frequency magnetrons are shown in figure 17. Figure 17A shows the output conductor connected to the upper surface of one of the segments of the anode block, and figure 17B shows the output conductor attached to one of the

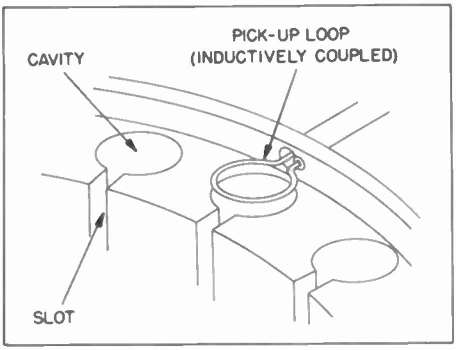


Figure 16. Coupling Method for Magnetrons Operating at Wavelengths Shorter than 3 Cm.

ring straps commonly used in short-centimeter radar.

CAVITY TYPES

Up to this point, the operation of only the strapped type of magnetron has been considered. However, there is another type of magnetron known as the rising-sun magnetron. The colorful name given to this type is due to the suggestive designs of the anode blocks, two of which are shown in figure 18.

The objective in the design of the anode block of a rising-sun magnetron is the same as that for a strapped magnetron; namely, to obtain as large a frequency separation between the π mode and neighboring modes as is necessary for proper operation. This objective is accomplished without employing straps by making alternate resonators alike and adjacent resonators unlike. As the sizes of the two sets of resonators are changed, the mode frequencies diverge into two groups as though each group corresponded to one set of the resonators (see figure 19). Therefore, the two groups of mode frequencies are well separated from the π mode, making this type of magnetron suitable for π-mode operation without the use of straps.

Rising-sun magnetrons are principally employed at wavelengths of 3 cm. and less. At these very short wavelengths, it is impracticable to employ straps because of the small dimensions of the resonators. However, for wavelengths of 3 cm. and above, it is usually desirable to keep the magnetron block as small as possible, thus favoring the strapped magnetron.

PERFORMANCE CONSIDERATIONS

Two methods have been found convenient and are in common use for presenting magnetron performance: the magnetron characteristic curves and the Rieke diagram. Magnetron characteristic curves are very useful in determining the optimum operating conditions for a magnetron. This is necessary in the

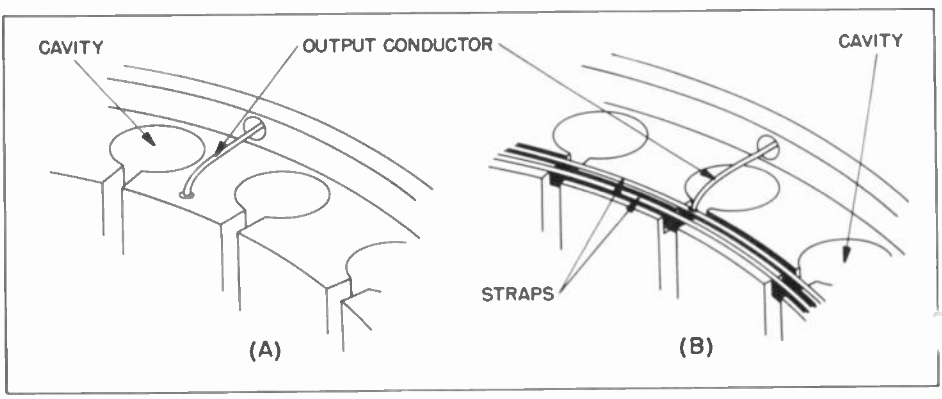


Figure 17. Other Common Methods of Coupling Energy from a Magnetron

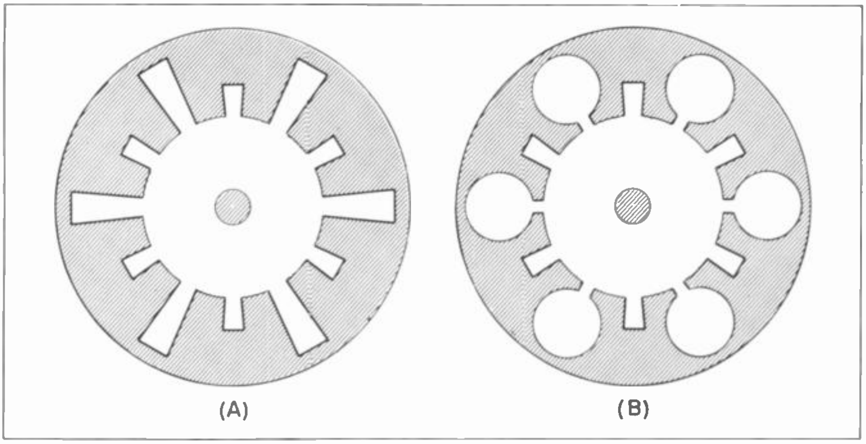


Figure 18. Typical Designs of Anode Blocks Used in Rising-Sun Magnetrons
A. Slot type
B. Slot Type with Modified Resonators

design of a radar in order to permit the choice of the proper modulating voltage and magnetic field strength. The Rieke diagram is useful in determining the effect on magnetron operation of variations in r-f loading from the optimum value. Knowledge of this effect is necessary in order to deduce the operation of a magnetron in a particular application.

Magnetron characteristics are drawn in Cartesian coordinates with the modulating voltage being plotted against the current during the pulse and the load being always adjusted for maximum power. Contours of constant flux density and r-f power are drawn. and additional contours of efficiency and input impedance are usually included. See figure 20. In some cases. frequency contours are also shown. A number of facts about magnetron operation can be seen from the characteristics. First. in order to obtain high efficiency. it is necessary to have a high value of modulating voltage and magnetic flux density. Second. for a constant magnetic flux density there is an optimum value of modulating voltage. Also. it is necessary that the modulating voltage remain essentially constant during the pulse. otherwise the power output will drop off. accompanied by a variation in frequency which might seriously affect the receiver performance.

Frequency change produced by changing input conditions is known as "pushing." The pushing figure is usually expressed in terms of megacycles per second per ampere. In most cases, push-

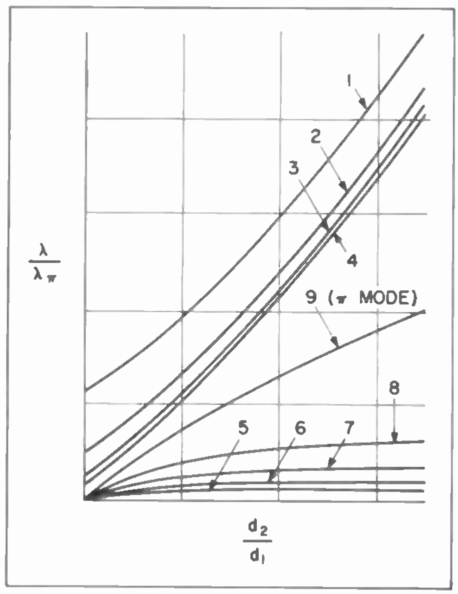


Figure 19. Typical Mode Spectrum of a Rising-Sun Magnetron (λ/λ_π is ratio of the mode wavelength to the π-mode wavelength, and d_2/d_1 is the ratio of the large resonator depth to the small resonator depth.)

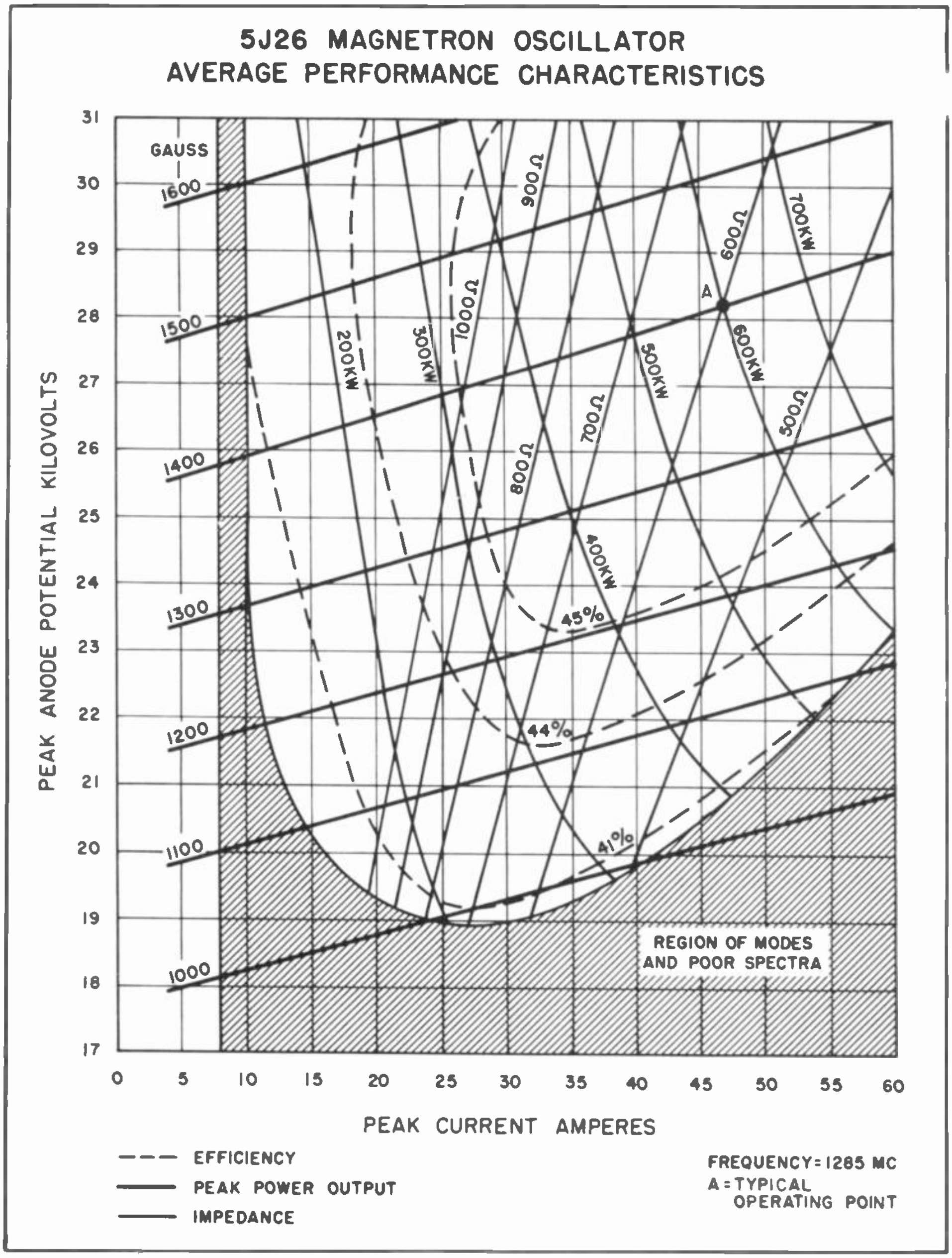


Figure 20. Typical Magnetron Characteristic Curves

ing is slight and can be neglected except at the shorter wavelengths.

The Rieke diagram is a polar diagram that shows the magnetron performance for fixed modulating voltage (or current) and flux density with variations in load impedance. See figure 21. Contours of constant frequency, constant power output, and efficiency are usually included. An important property of the

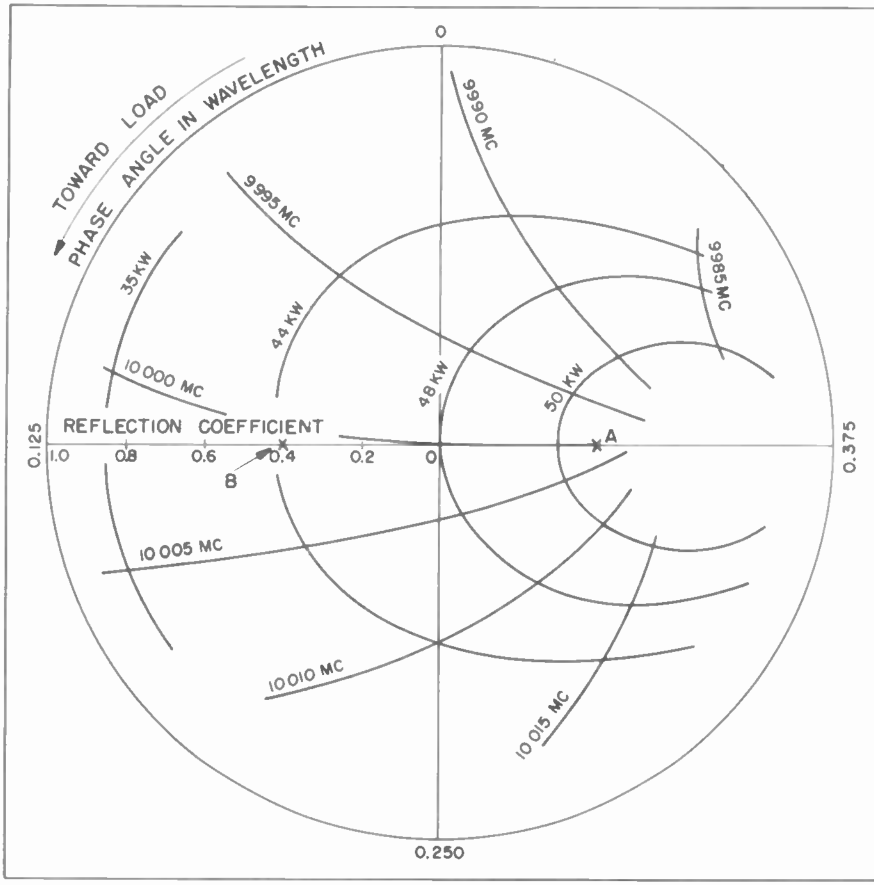


Figure 21. Typical Magnetron Rieke Diagram

Rieke diagram is that when the transmission line is terminated in a load represented by point A. the magnitude of the reflection coefficient (K) is represented by OA and the phase of the reflection by O. The reflection coefficient (K) is related to the standing-wave ratio (ρ) by the expression:

$$K = \frac{1 - \rho}{1 + \rho}$$

As an example of some of the facts that can be deduced about magnetron operation from the Rieke diagram, consider the following problem: It is desired to operate a magnetron with high output power and good frequency stability. Point A lies in the region of highest output power. However, following the constant K circle from A to the labeled axis reveals that the reflection coefficient is not zero, and, hence, that the standing-wave ratio is not unity. Thus, the line or waveguide is not matched. Also, frequency stability at point A is poor, since a slight change in load (due possibly to a faulty rotating joint) causes a comparatively great change in frequency, because the frequency contours converge in this region. The change in magnetron frequency due to a change in load impedance is known as "pulling." Thus, good frequency stability is impossible when a magnetron is operated for high output power. However, if the length of the line to the mismatch is changed until point B is reached, it can be seen that the frequency stability is improved, since the "pulling" is less. At point B, for the same change in loading, the frequency change is less. The operating point of a magnetron is usually a compromise between high output power and good frequency stability, and is usually chosen at the center of the Rieke diagram, which corresponds to a matched load.

Magnetron performance is often summed up in what is called the "pulling figure." This figure is the total change in frequency (usually in mc.) which occurs when the load is adjusted to produce a voltage-standing-wave ratio of 1.5:1 and the phase of the reflection is varied through 360 degrees.

PRETUNING A MAGNETRON

The straps in a magnetron provide a means for pretuning. Before the end plate is placed on the magnetron, a c-w signal from a klystron oscillator at the desired frequency is injected into the magnetron through the coupling loop, and the straps are bent until the magnetron resonates at the required frequency. Resonance is determined by an indicator in the coaxial line between the klystron and the magnetron. Allowance for the presence of the end plate and the space charge is made by the use of empirical rules which enable the required accuracy to be obtained.

MAGNETRON TUNING

Two factors that are commonly made use of in tuning a magnetron are the anode length and the end-plate-to-anode spacing, both of which influence the magnetron frequency. One form of a tunable magnetron has a thin, flexible end plate which may be depressed at the center to change the frequency. Changes of about 150 mc. at 3000 mc. can be obtained in this way. Another method that is used to alter the effective anode length and anode-to-end plate separation makes use of metal rods or pins which move in and out of the anode block.

PRACTICAL CONSIDERATIONS

Arcing

Arcing, or sparking, in small magnetrons, or magnetrons operated at high power is very common. One of the causes is the presence of traces of gases, which are evolved during long periods of idleness. Each time this condition occurs, the magnetron must be reseasoned, as described in the following section. Arcing shortens the useful life of a magnetron by damaging the cathode; however, the cathode can withstand considerable arcing for short periods of time without any immediate damage. Other causes of arcing are the presence of sharp surfaces, mode shifting, and overworking the cathode.

Seasoning

New magnetrons require an initial break-in period because arcing has been found to be particularly violent when they are first put in operation. The breaking-in, or seasoning, of a magnetron consists of slowly increasing the voltage applied until arcing occurs several times a second. The voltage is left at this value until the arcing has died out. Then the voltage is raised until arcing again occurs and is left at that value until the arcing dies out again. This procedure is continued until the normal operating voltage of the magnetron is reached. Many times the arcing will become very violent, often resembling a continuous arc. Since this condition is particularly dangerous to the cathode, the applied voltage should be quickly reduced to permit the magnetron to recover. The above procedure for seasoning new magnetrons is also used on magnetrons that have been left idle for any extended period of time.

Handling Magnetrons

Although a magnetron is very rugged

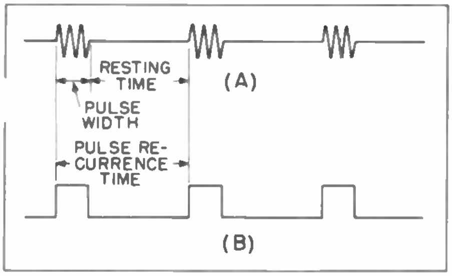


Figure 22. Magnetron Test Waveforms

electrically, it should be handled with a great deal of care. The glass-to-copper seals are very fragile and are easily damaged. Even if the seals are not broken, careless handling of the leads may alter the alignment of the cathode and seriously change the characteristics of the magnetron.

Cathode Features

When a magnetron is in operation, all of the electrons that are not aiding the oscillation are forced back to the cathode and cause additional heat to be dissipated by the cathode. The energy received by the cathode in this way is great enough, in many applications, to permit the cathode heater voltage to be reduced, or even be switched off, after the initial warm-up period. Failure to reduce the heater voltage, in many cases, would result in cathode disintegration.

MAGNETRON TESTING

When a magnetron or an entire transmitter is tested, it is common practice either to detect the r-f envelope and deduce the quality of performance from

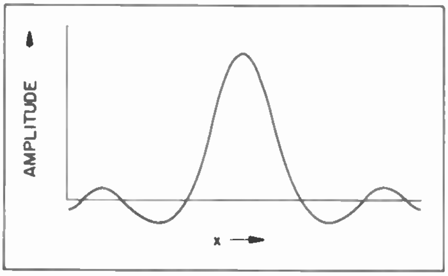


Figure 23. Curve of (sin x)/x

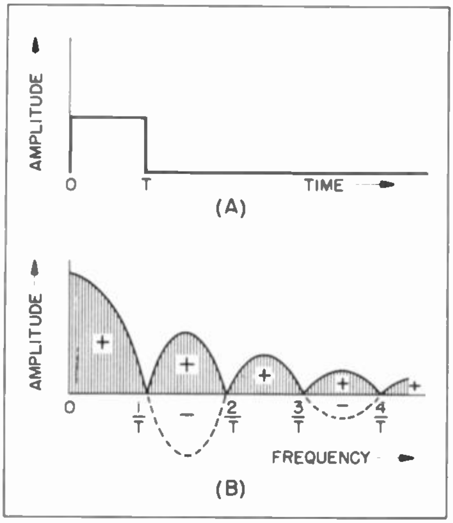


Figure 24. Graph of D-C Pulse with Corresponding Frequency Spectrum

its shape or to analyze the r-f pulse directly. For proper operation, the r-f pulses and the corresponding detected d-c pulses should appear as in figure 22.

As a background for understanding the methods of analyzing pulses, consider some of the characteristics of rectangular pulses. The characteristics of interest are the frequency components and their relative amplitudes. This information can be found by the use of Fourier integral analysis. The results of such an analysis show that the envelope of the amplitudes of the frequency spectrum will be of the form (sin x)/x. The curve for this expression is shown in figure 23, and is characterized by symmetry about a main maximum amplitude with adjacent minima.

The frequency spectrum of a d-c pulse is shown in figure 24. Since the average value of the pulse is not zero, a d-c component is present and is indicated in figure 24B by the zero frequency amplitude. Also, from the curve, the frequency components n/T (where n is an integer) are not present since they are at the zero points of the curve. Two differences between this curve and the

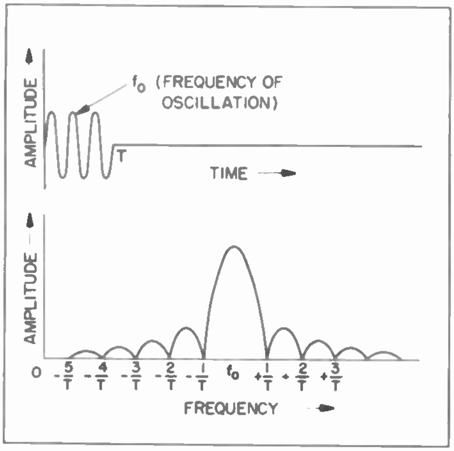


Figure 25. Frequency Spectrum of R-F Pulse

(sin x)/x curve are the absence of symmetry about the main maximum and the 180-degree reversal of the negative swings indicated by the dashed lines in figure 24B. The absence of symmetry is due to the fact the negative frequency components are not shown, since they have no physical meaning. The 180-degree reversal occurs because the curve

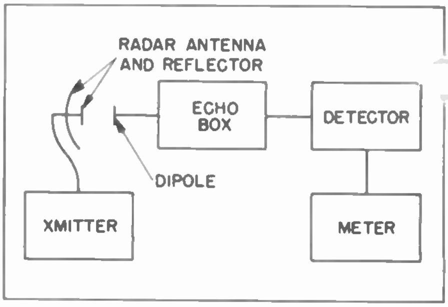
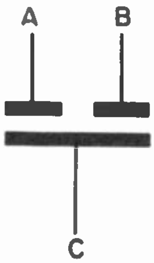


Figure 26. Typical Test Setup Used in Spectrum Analysis

is normally not plotted with negative amplitudes (since they only indicate a 180-degree phase change), and, in practice, the magnitudes are much more readily measured. Thus, the dashed curve is replaced by the solid portion.

The frequency spectrum of an r-f pulse is shown in figure 25. Since the average value of the r-f pulse is zero, no d-c component is present. Also, the main maximum occurs at the frequency of the pulse, with the zero-amplitude components occurring at $f_0 \pm n/T$ (where n is an integer).

What's Your Answer?



Dual ceramic capacitors are becoming more and more common in electronic equipment, and it is important that these units have certain characteristics. Examination of specifications showed that a dual .004-μf. capacitor (as shown in the drawing) should have no more than 20 $\mu\mu$f. of stray capacitance between points A and B. However, just looking at the circuit shows that between points A and B, a capacitance of 2000 $\mu\mu$f. should exist because the two sections effectively appear in series. A capacitor-analyzer test confirmed this view.

Are the specifications at fault? Explain.

(Solution next month)

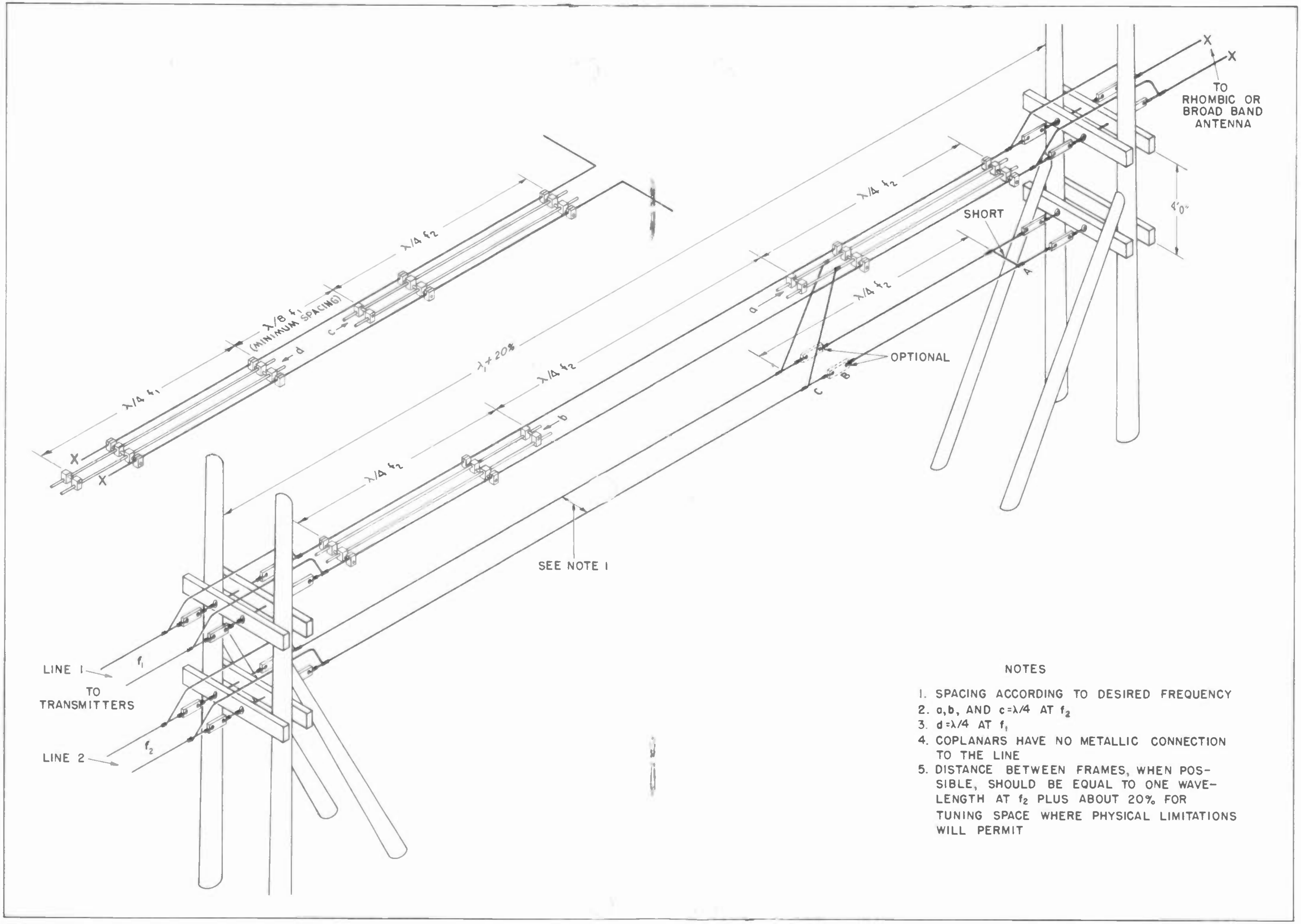


Figure 1. Typical Simultaneous Transmission System Incorporating Inductively Coupled Quarter-Wavelength Coupling Sections (Coplanars) to Match Antenna to Line, and Similar Sections (Used as Traps) to Prevent Interaction between Transmitters

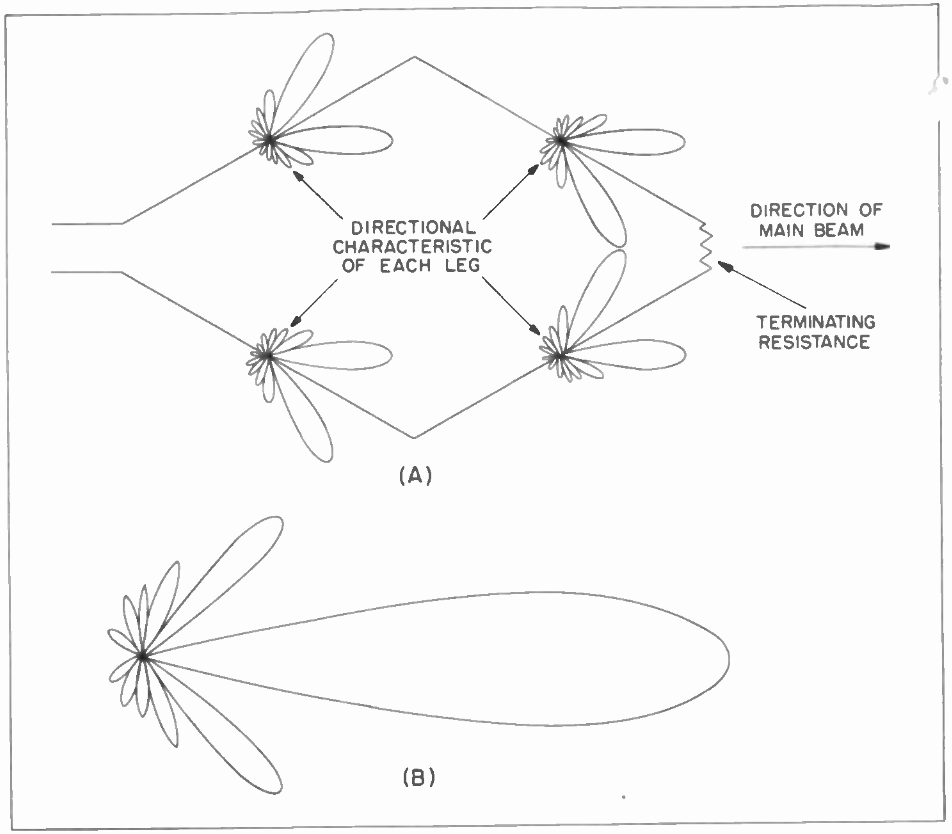


Figure 2. Radiation Pattern of Typical Terminated Horizontal Rhombic, Showing (A) Radiation from Individual Legs, and (B) Resultant Pattern

is a flat, 600-ohm line terminated by the 600-ohm rhombic. The second signal, f_2, is fed into line 2, and then coupled into line 1 by means of coupling section a. At the left end of the drawing, a quarter-wavelength trap (b) cut to f_2 is installed to prevent f_2 energy from being transmitted back along line 1 to the f_1 transmitter. At the upper left of the drawing, quarter-wavelength impedance-matching sections cut to f_1 and f_2 are used to match the antenna to the transmission line in order to "flatten" the line at the desired frequencies. Additional similar elements could be added to the circuit, if desired, in order to extend the system to three frequencies. Each element of the system shown in figure 1 will be discussed in detail in this article, as will a number of alternate circuit elements which might also be used.

ANTENNAS

There are several types of antennas that can be used for the transmission of high-powered directional communications signals; these types include Sterba-curtain arrays, and the long-wire, Beverage, and rhombic (or diamond) antennas.

The multi-wire, horizontal rhombic antenna is the one most commonly employed for long-distance, sky-wave transmission. The Army Communications Service has standardized a number of types of rhombics (with the design of each based on required range, frequency, etc.), each of which are fur-

SIMULTANEOUS TRANSMISSION SYSTEMS

Part I of Two Parts

by F. R. Sherman

Headquarters Technical Staff

A general discussion of the requirements of simultaneous transmission systems, including types of antennas, characteristics of transmission lines, and methods of detecting unbalance in lines; followed by a detailed coverage of several methods used for matching the input impedance of a multi-wire rhombic or other broad-band antenna to the impedance of an r-f transmission line at two or more frequencies.

(Editor's Note: Portions of this article are based on various Radio Engineering Reports published by the Electronics Establishment Branch of the Civil Aeronautics Administration, U. S. Department of Commerce. Grateful acknowledgement is made to that agency for its assistance in supplying technical data and for permission to extract material from its reports on simultaneous transmission.)

At times, the space available for an antenna farm may be small, thus reducing the number of antennas that may be erected below the number actually required for efficient operation. However, if it is possible to use one antenna simultaneously or successively for the transmission of signals at different frequencies, a large saving in initial installation expense can be realized, in addition to a considerable reduction in maintenance costs.

There are, of course, many types of broad-band arrays that are adaptable to multi-band operation at harmonically related frequencies throughout the high-frequency spectrum. But these arrays require resonant r-f transmission lines, which, in turn, are critically dependent upon line length. On the other hand, it is possible to transmit simultaneously from one antenna several non-harmonically-related frequencies that are separated by as little as 6% in frequency. This is accomplished by inserting into the r-f transmission line, near the antenna input, matching sections which will match the antenna impedance to the line impedance at each frequency desired. The result will be that the r-f transmission line will be non-resonant (flat) at each of these frequencies, and the networks coupling energy from secondary transmitters into the main r-f transmission line will not affect the circuit except at the frequencies at which they act as coupling devices. This system, when set up for three frequencies, will allow the efficient transmission of from one to three signals simultaneously from one antenna.

TYPICAL SIMULTANEOUS TRANSMISSION SYSTEM

Figure 1 shows a typical simultaneous-transmission installation in which a rhombic antenna is fed the output of two transmitters (at frequencies f_1 and f_2), each coupled into the same transmission line. It can be seen that the f_1 signal is fed directly into line 1, which

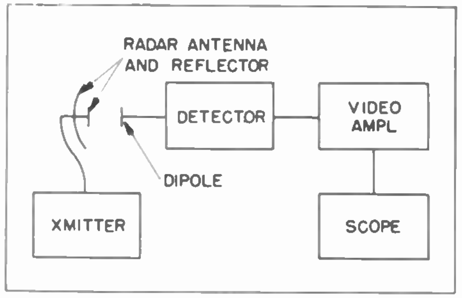


Figure 28. Typical R-F Envelope Indicator Test Setup

A second method of testing a magnetron makes use of direct observation of the rectified pulse on an oscilloscope screen. The results are compared with the ideal rectified output shown in figure 22. A typical test setup, known as an r-f envelope indicator, is shown in figure 28.

The antenna used in this setup also has a low Q so that it will not be frequency-sensitive. The detected output is amplified by a wide-band amplifier, and is applied to the vertical-deflection plates of the oscilloscope. The sweep voltage of the oscilloscope is linear, and the sweep circuit is triggered by the same pulse that triggers the transmitter. Some typical waveforms obtained with an r-f envelope indicator, showing ideal, good, and poor magnetron operation, are contained in figure 29.

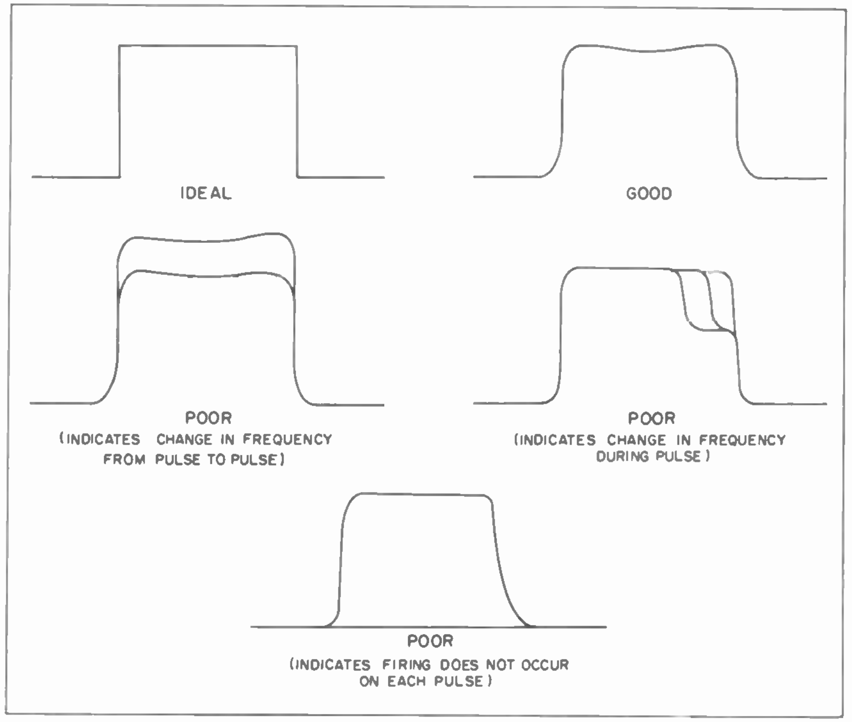


Figure 29. Typical Waveforms Obtained with R-F Envelope Indicator, Showing Ideal, Good, and Poor Magnetron Operation

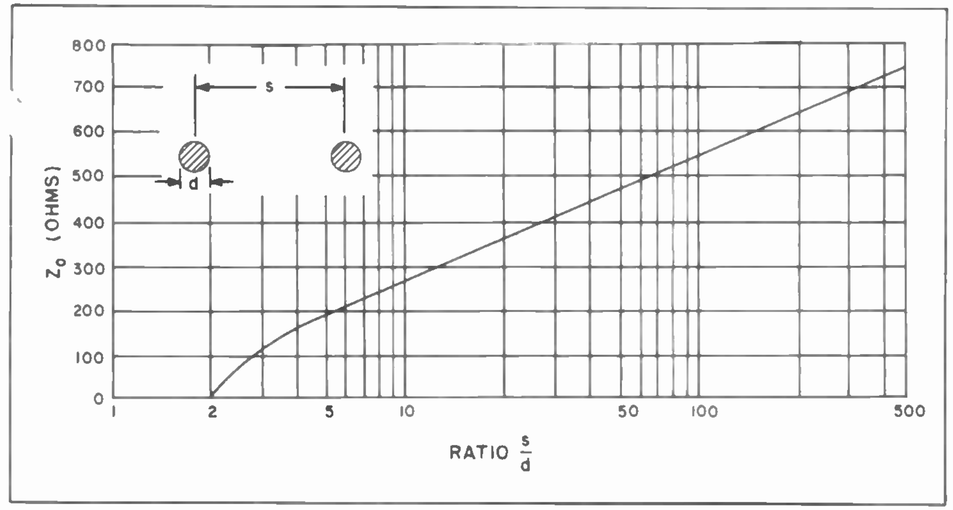


Figure 3. Graph of Characteristic Impedance (Z_o) of Two-Wire Line

nished complete with telegraph poles or steel towers. These antenna kits are ordered according to the particular requirements of each site. When used for transmitting, they generally incorporate a multi-wire (curtain) arrangement on each leg, which serves to reduce the antenna input impedance to about 600 ohms, and provides a more uniform impedance over the operating-frequency band. In fact, by adjusting the terminating resistance to an optimum value, it is possible to obtain an input impedance which, in a typical case, has resistive and reactive components neither of which deviates more than 20% from a mean value of 800 to 600 ohms for a frequency range of 4 to 20 mc., respectively. This, combined with the high directivity of the rhombic (see figure 2), more than offsets the 50% loss of power (due to power dissipated in the terminating resistor) in a terminated rhombic antenna.

Although the directivity of the rhombic decreases if the frequency is lowered while the length of the antenna legs is held constant, the variations are not severe. Thus, while optimum performance is achieved at one frequency only, the input impedance and directivity of the rhombic vary so little over a wide range of frequencies, that the antenna lends itself readily to simultaneous transmission. In such cases, it is necessary that the antenna input impedance approximate the characteristic impedance of the r-f transmission line at all frequencies to be used. Furthermore, the antenna must be matched to a properly *balanced* transmission line at all assigned frequencies. Ordinarily, unbalance is not produced by the antenna system, because such systems are usually electrically symmetrical throughout.

R-F TRANSMISSION LINES

The principal types of r-f lines are the twisted two-wire line, the spaced

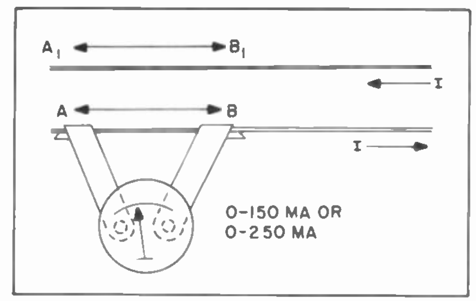


Figure 4. Shunt-Milliameter Method for Checking Balance of Transmission Line

TABLE I. PARTS LIST FOR TROLLEY METER SHOWN IN FIGURES 5 AND 6

QUANTITY	SYMBOL	DESCRIPTION
1 ea.	L_1, L_2, L_3, and L_4	Each two turns #18, solid, insulated conductor (Belden #8945), all coils laced in one cable and placed in shielded pickup loop
1 ea.	R	Resistor, 7500-ohm, 1/2-watt
1 ea.	S_1	Consists of two Centralab Type-S ceramic wafers, and one K-121 index head
1 ea.	S_2	Switch, toggle, s-p-s-t, Bud Radio #SW-1003
1 ea.	M	Milliameter, 0-1 ma., d-c, Weston #301, bakelite case
1 ea.	B	Battery, 1-1/2 volt, #2 flashlight cell
1 ea.	C	Capacitor, variable, 25-$\mu\mu$f., Bud Radio #MC-909
1 ea.		Plate, dial, Crowe #263
2 ea.		Knob, bar, Crowe #286
1 ea.		Frame, calibration chart, Allied Radio #A1461
As req.		Miscellaneous hardware, solder, etc.

two-wire line, the shielded two-wire line, the flexible coaxial cable, and the rigid coaxial cable.

The most commonly used type is the spaced, two-wire line consisting of two parallel conductors which are maintained at a fixed distance by insulating spacers, or spreaders. This type of line is used because of its ease of construction, its economy, and its efficiency, and is the type which will be discussed here.

Z_0 of a Parallel-Conductor Line

When the operating frequency makes the reactive components of impedance exhibited by a line large enough so that the resistance is negligible, the characteristic (or surge) impedance of a transmission line (in ohms) is equal to the square root of the ratio of the inductance (in henries) to the capacitance of the line (in farads). Thus:

$$Z_0 = \sqrt{L/C}$$

In the case of a two-wire, parallel-conductor line,

$$L = K \log_{10} (s/d), \text{ and}$$

$$C = K \frac{1}{\log_{10} (s/d)}$$

where K is a constant, s the separation, or spacing, of the conductors, and d the diameter of the conductor expressed in the same units as s. An increase in the separation (s) of the wires will increase the inductance of the line and decrease its capacitance. This increases the LC ratio and hence the characteristic impedance. Reducing the diameter (d) of the wires increases the inductance, decreases the capacitance, and again increases the characteristic impedance.

The relationship between the characteristic impedance and the dimensions of a two-wire line is expressed mathematically by the equation:

$$Z_0 \text{ (in ohms)} = 276 \log_{10} (2s/d),$$

where s is the spacing between the wires (center-to-center), and d the diameter of the conductor. Example—for a two-wire line constructed of #6 wire (diameter = 162.0 mils) spaced 12 inches apart:

$$\begin{aligned} Z_0 &= 276 \log_{10} (2s/d) \\ &= 276 \log_{10} (2 \times 12)/.162 \\ &= 276 \log_{10} 148 \\ &= 276 \times 2.1703 \\ &= 600 \text{ ohms} \end{aligned}$$

Figure 3 shows the relationship between the characteristic impedance and the physical dimensions of the two-wire transmission line (the latter being expressed as a ratio of the conductor spacing to conductor diameter). By the use of this chart, the characteristic impedance of a two-wire, parallel-conductor

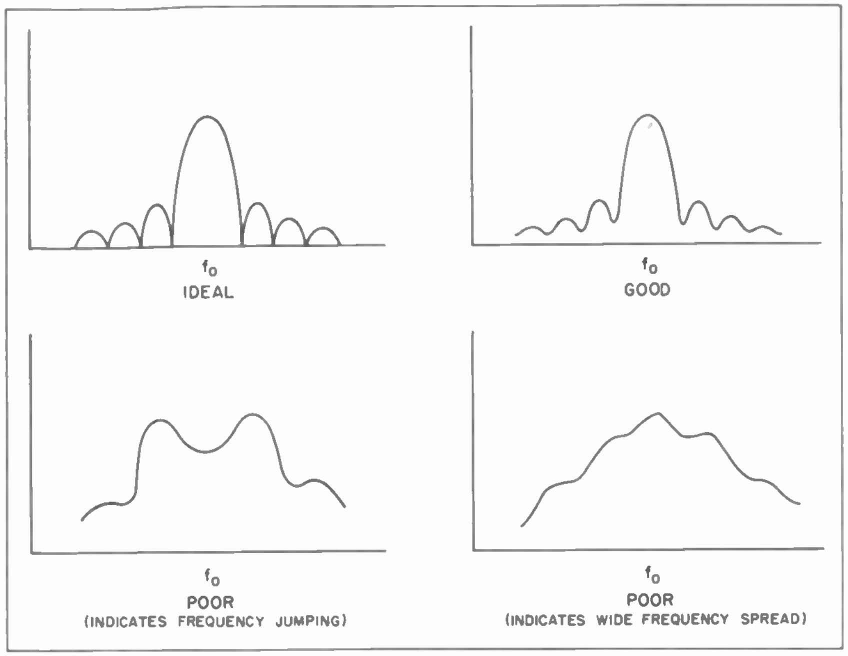


Figure 27. Typical Spectra of R-F Pulse

It is interesting to note that the pulse power (maximum-amplitude components) is principally concentrated in a band of frequencies from $f_0 - 1/T$ to $f_0 + 1/T$. It will be found that most radar receivers are designed to pass most of this band of frequencies. Thus, the optimum bandwidth of a radar receiver is equal to 2/T. For example, if a radar has an r-f pulse duration of 0.8 microsecond, the optimum bandwidth of the receiver would be $2/0.8 \times 10^{-6}$ (or 2.5 mc.). A typical radar employing this pulse duration has a receiver bandwidth of 2 mc. Should the occurrence of a trouble make the pulse duration shorter, much of the pulse energy would be lost because it would not be amplified properly by the receiver. The shorter the pulse duration, the greater the required bandwidth, and vice versa.

One method of testing a magnetron is termed "spectrum analysis." The method involves measuring the relative amplitudes of the frequency components present in the output of a transmitter and comparing the results with the ideal output previously discussed. A number of different test setups can be employed to determine the spectrum; however, the echo box is most commonly used in the field because of its portability and simplicity. A typical test setup employing an echo box for spectrum analysis is shown in figure 26.

The output from the antenna is picked up by a dipole that has a low Q to permit all the frequency components of the r-f pulse (shown in figure 25) to be applied to the echo box. The echo box, which is very frequency-sensitive, is tuned over a range on each side of the magnetron frequency, and the results are tabulated. Some typical spectrums, indicating ideal, good, and poor results, are shown in figure 27. The echo-box Q must be sufficiently high to result in a bandwidth that is small compared to 2/T (a Q of 30,000 will result in a bandwidth of 100 kc. at 3000 mc.).

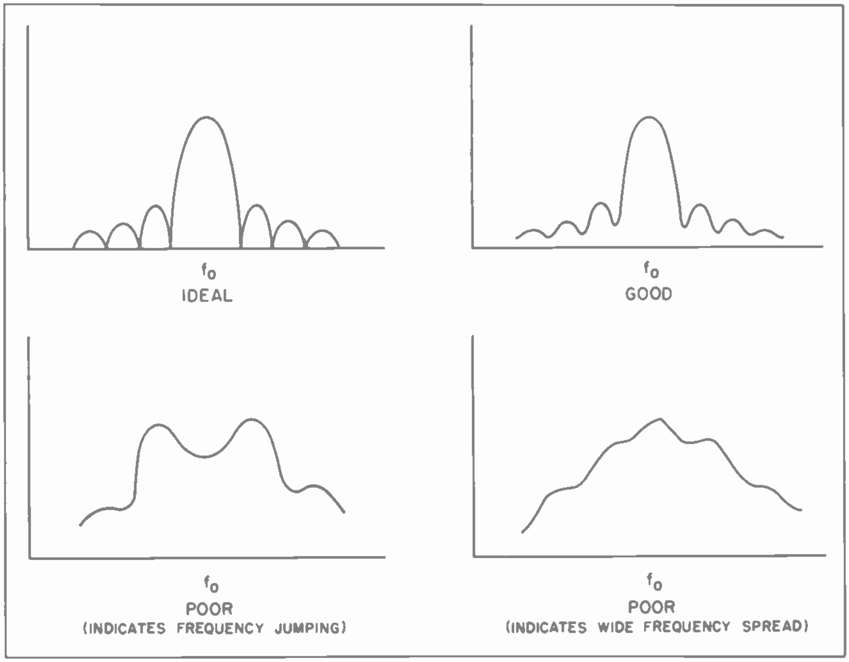


Figure 27. Typical Spectra of R-F Pulse

It is interesting to note that the pulse power (maximum-amplitude components) is principally concentrated in a band of frequencies from $f_0 - 1/T$ to $f_0 + 1/T$. It will be found that most radar receivers are designed to pass most of this band of frequencies. Thus, the optimum bandwidth of a radar receiver is equal to 2/T. For example, if a radar has an r-f pulse duration of 0.8 microsecond, the optimum bandwidth of the receiver would be $2/0.8 \times 10^{-6}$ (or 2.5 mc.). A typical radar employing this pulse duration has a receiver bandwidth of 2 mc. Should the occurrence of a trouble make the pulse duration shorter, much of the pulse energy would be lost because it would not be amplified properly by the receiver. The shorter the pulse duration, the greater the required bandwidth, and vice versa.

One method of testing a magnetron is termed "spectrum analysis." The method involves measuring the relative amplitudes of the frequency components present in the output of a transmitter and comparing the results with the ideal output previously discussed. A number of different test setups can be employed to determine the spectrum; however, the echo box is most commonly used in the field because of its portability and simplicity. A typical test setup employing an echo box for spectrum analysis is shown in figure 26.

The output from the antenna is picked up by a dipole that has a low Q to permit all the frequency components of the r-f pulse (shown in figure 25) to be applied to the echo box. The echo box, which is very frequency-sensitive, is tuned over a range on each side of the magnetron frequency, and the results are tabulated. Some typical spectrums, indicating ideal, good, and poor results, are shown in figure 27. The echo-box Q must be sufficiently high to result in a bandwidth that is small compared to 2/T (a Q of 30,000 will result in a bandwidth of 100 kc. at 3000 mc.).

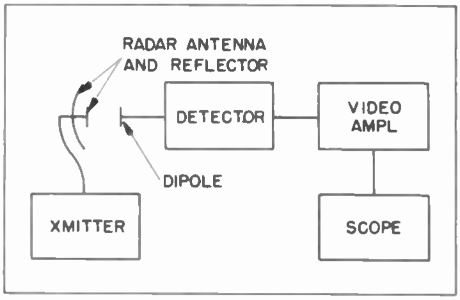


Figure 28. Typical R-F Envelope Indicator Test Setup

A second method of testing a magnetron makes use of direct observation of the rectified pulse on an oscilloscope screen. The results are compared with the ideal rectified output shown in figure 22. A typical test setup, known as an r-f envelope indicator, is shown in figure 28.

The antenna used in this setup also has a low Q so that it will not be frequency-sensitive. The detected output is amplified by a wide-band amplifier, and is applied to the vertical-deflection plates of the oscilloscope. The sweep voltage of the oscilloscope is linear, and the sweep circuit is triggered by the same pulse that triggers the transmitter. Some typical waveforms obtained with an r-f envelope indicator, showing ideal, good, and poor magnetron operation, are contained in figure 29.

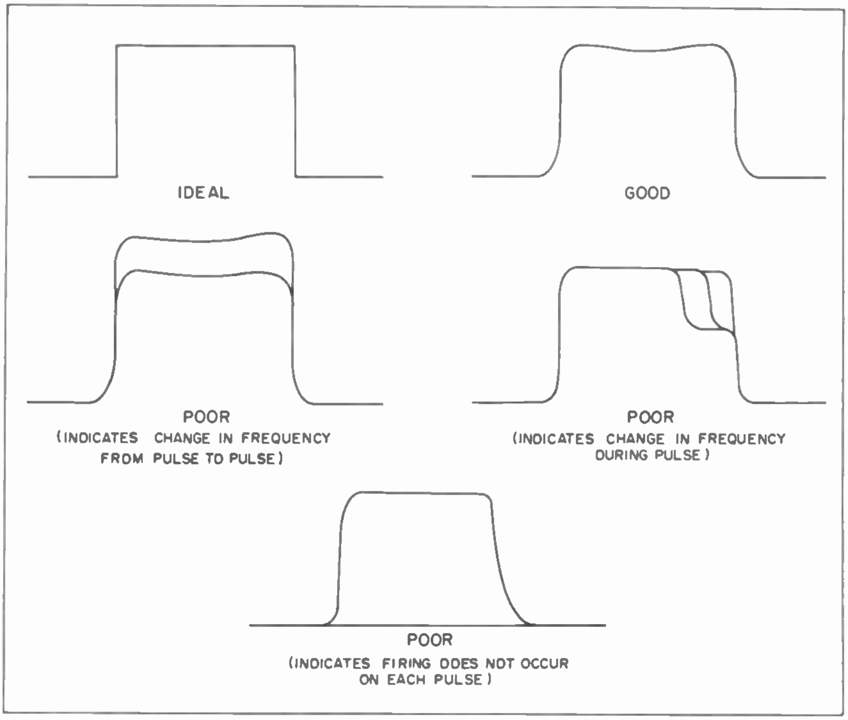


Figure 29. Typical Waveforms Obtained with R-F Envelope Indicator, Showing Ideal, Good, and Poor Magnetron Operation

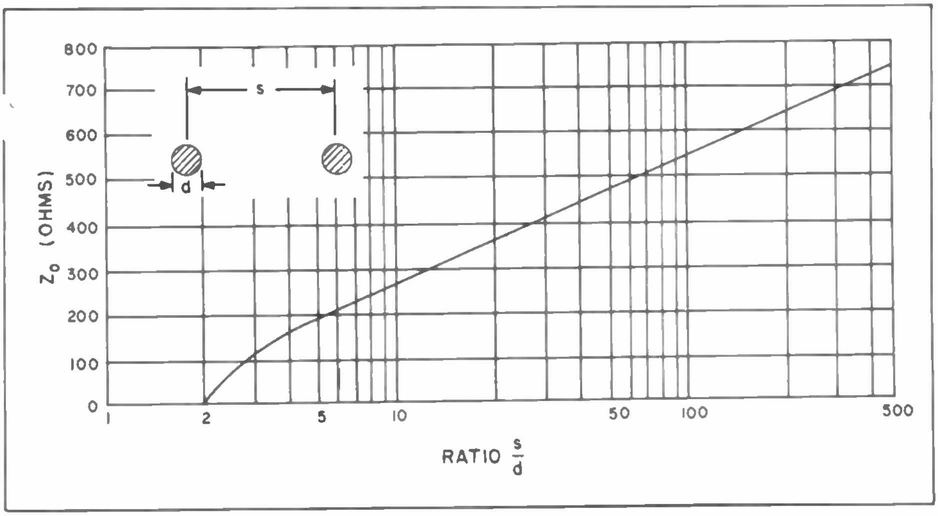


Figure 3. Graph of Characteristic Impedance (Z_o) of Two-Wire Line

nished complete with telegraph poles or steel towers. These antenna kits are ordered according to the particular requirements of each site. When used for transmitting, they generally incorporate a multi-wire (curtain) arrangement on each leg, which serves to reduce the antenna input impedance to about 600 ohms, and provides a more uniform impedance over the operating-frequency band. In fact, by adjusting the terminating resistance to an optimum value, it is possible to obtain an input impedance which, in a typical case, has resistive and reactive components neither of which deviates more than 20% from a mean value of 800 to 600 ohms for a frequency range of 4 to 20 mc., respectively. This, combined with the high directivity of the rhombic (see figure 2), more than offsets the 50% loss of power (due to power dissipated in the terminating resistor) in a terminated rhombic antenna.

Although the directivity of the rhombic decreases if the frequency is lowered while the length of the antenna legs is held constant, the variations are not severe. Thus, while optimum performance is achieved at one frequency only, the input impedance and directivity of the rhombic vary so little over a wide range of frequencies, that the antenna lends itself readily to simultaneous transmission. In such cases, it is necessary that the antenna input impedance approximate the characteristic impedance of the r-f transmission line at all frequencies to be used. Furthermore, the antenna must be matched to a properly *balanced* transmission line at all assigned frequencies. Ordinarily, unbalance is not produced by the antenna system, because such systems are usually electrically symmetrical throughout.

R-F TRANSMISSION LINES

The principal types of r-f lines are the twisted two-wire line, the spaced

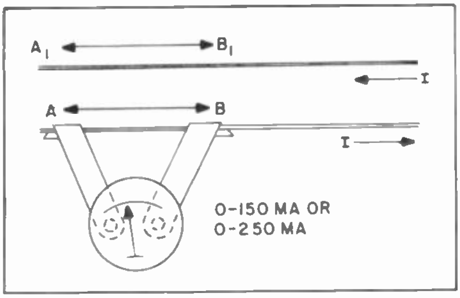


Figure 4. Shunt-Milliameter Method for Checking Balance of Transmission Line

TABLE I. PARTS LIST FOR TROLLEY METER SHOWN IN FIGURES 5 AND 6

QUANTITY	SYMBOL	DESCRIPTION
1 ea.	L_1, L_2, L_3, and L_4	Each two turns #18, solid, insulated conductor (Belden #8945), all coils laced in one cable and placed in shielded pickup loop
1 ea.	R	Resistor, 7500-ohm, 1/2-watt
1 ea.	S_1	Consists of two Centralab Type-S ceramic wafers, and one K-121 index head
1 ea.	S_2	Switch, toggle, s-p-s-t, Bud Radio #SW-1003
1 ea.	M	Milliameter, 0-1 ma., d-c, Weston #301, bakelite case
1 ea.	B	Battery, 1-1/2 volt, #2 flashlight cell
1 ea.	C	Capacitor, variable, 25-$\mu\mu$f., Bud Radio #MC-909
1 ea.		Plate, dial, Crowe #263
2 ea.		Knob, bar, Crowe #286
1 ea.		Frame, calibration chart, Allied Radio #A1461
As req.		Miscellaneous hardware, solder, etc.

two-wire line, the shielded two-wire line, the flexible coaxial cable, and the rigid coaxial cable.

The most commonly used type is the spaced, two-wire line consisting of two parallel conductors which are maintained at a fixed distance by insulating spacers, or spreaders. This type of line is used because of its ease of construction, its economy, and its efficiency, and is the type which will be discussed here.

Z_0 of a Parallel-Conductor Line

When the operating frequency makes the reactive components of impedance exhibited by a line large enough so that the resistance is negligible, the characteristic (or surge) impedance of a transmission line (in ohms) is equal to the square root of the ratio of the inductance (in henries) to the capacitance of the line (in farads). Thus:

$$Z_0 = \sqrt{L/C}$$

In the case of a two-wire, parallel-conductor line,

$$L = K \log_{10} (s/d), \text{ and}$$

$$C = K \frac{1}{\log_{10} (s/d)}$$

where K is a constant, s the separation, or spacing, of the conductors, and d the diameter of the conductor expressed in the same units as s. An increase in the separation (s) of the wires will increase the inductance of the line and decrease its capacitance. This increases the LC ratio and hence the characteristic impedance. Reducing the diameter (d) of the wires increases the inductance, decreases the capacitance, and again increases the characteristic impedance.

The relationship between the characteristic impedance and the dimensions of a two-wire line is expressed mathematically by the equation:

$$Z_0 \text{ (in ohms)} = 276 \log_{10} (2s/d),$$

where s is the spacing between the wires (center-to-center), and d the diameter of the conductor. Example—for a two-wire line constructed of #6 wire (diameter = 162.0 mils) spaced 12 inches apart:

$$\begin{aligned} Z_0 &= 276 \log_{10} (2s/d) \\ &= 276 \log_{10} (2 \times 12)/.162 \\ &= 276 \log_{10} 148 \\ &= 276 \times 2.1703 \\ &= 600 \text{ ohms} \end{aligned}$$

Figure 3 shows the relationship between the characteristic impedance and the physical dimensions of the two-wire transmission line (the latter being expressed as a ratio of the conductor spacing to conductor diameter). By the use of this chart, the characteristic impedance of a two-wire, parallel-conductor

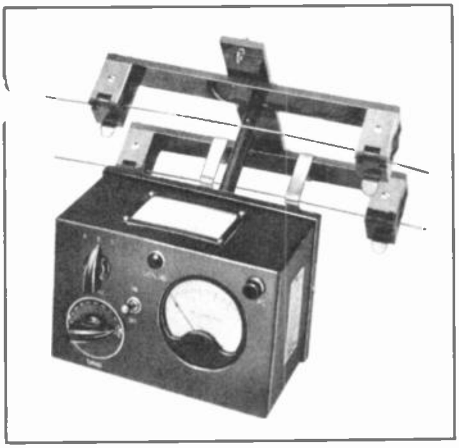


Figure 5. CAA Trolley Meter CA-543

line can be determined from its physical dimensions. Conversely, the physical dimensions for a line of desired characteristic impedance can be determined from the chart.

Direct Measurement of Z_0

It is usually more convenient to calculate the characteristic impedance of a given line from formulas or charts like the one in figure 3 rather than to obtain it by measurement. However, a fairly accurate and simple method of determining the characteristic impedance by direct measurement is to terminate the line with a variable, non-inductive resistance, which is varied until the standing waves are eliminated, at which time the line is properly terminated and the value of the resistance is equal to the characteristic impedance of the line. The ohmic value of the characteristic impedance can then be obtained by measuring the value of the terminating resistance with an ordinary ohmmeter.

Balance in Two-Wire Parallel-Conductor Lines

The ease with which simultaneous transmission networks can be incorporated into an antenna system depends largely upon the initial installation of both the antenna and the two-wire r-f transmission line. These lines must be carefully engineered and installed with all factors affecting balance being taken into consideration. Symmetry between wires and surrounding objects is essential in order that there be equal capacity all along the r-f transmission line, since matching the antenna to the main r-f transmission line at the frequencies to be used will be impossible to achieve if there is too much unbalance in the system. If, for some reason, the current in one conductor is higher than in the other, or if the currents in the two wires are not exactly 180 degrees out of phase, the electromagnetic fields will not cancel completely and a considerable amount of power may be radiated from the line.

Maintaining a good balance requires, first of all, a balanced load at the termination of the line. For this reason, the antenna should be fed, whenever possible, at a point where each conductor "sees" exactly the same thing (the looking-in impedance is the same for both conductors). Usually this requires that the antenna system be fed at its electrical center. Even though the antenna appears to be physically symmetrical, it can be unbalanced electrically if a part connected to one of the antenna conductors is inadvertently coupled to something (such as house wiring or a metal pole or roof) that is not duplicated on the other side of the antenna. Every effort should be made to keep the antenna as far as possible from other wiring or sizable metallic objects. The transmission line itself will cause some unbalance if it is not brought away from the antenna at right angles to it for a distance of at least a quarter-wavelength. The minimum separation between either conductor and all other wiring should be at least four or five times the conductor spacing. The shunt capacitance introduced by close proximity to metallic objects can drain off enough current to ground to unbalance the line currents, and result in increased line radiation. A shunt capacitance of

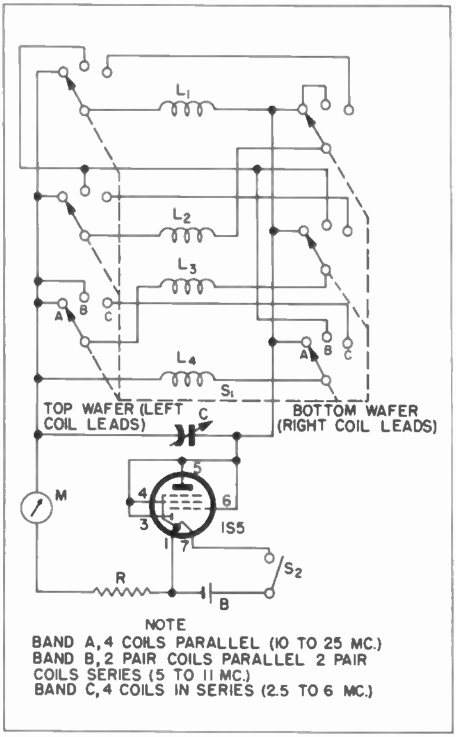


Figure 6. Schematic Diagram of Trolley Meter Shown in Figure 5

this sort also constitutes a reactive load on the line, which causes an impedance "bump" that will prevent making the line actually flat.

Checking Balance

The simplest method of detecting unbalance is the use of an r-f, thermocouple-type milliameter (see figure 4). Short, heavy wires or straps are fastened to the meter which is then hooked directly on one conductor. Since the section of line between the straps acts as a shunt for the meter, adjusting the separation between the straps will vary the range of the meter. When this type of meter is used, a reading is taken on one conductor (points A and B), then placed on the other conductor (points A_1 and B_1) directly opposite the positions where the first readings were made. If the readings are within 10% of each other, the line is sufficiently balanced.

Such instruments as the CAA trolley meter CA-543 (shown in figures 5 and 6) may be used in a similar manner.

The most accurate method of checking balance is by means of a temporary building-out section (stub) between 1/8 and 1/4 wavelength long, shorted at the end, and grounded at the center of the shorting bar (figure 7). Since unbalance is usually caused by the transmitter, it is best to check for unbalance as near to the transmitter house as possible. The trolley meter is held at point A, near the short, and tuned for maximum indication. The meter is then held the same distance from the grounding wire at point B. If the second reading is more than 1/10 of the first, the unbalance is too great.

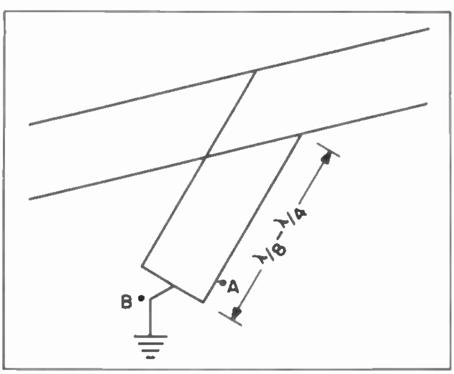


Figure 7. Temporary Building-Out Section (Stub)

If unbalance is indicated, a thorough check must be made to see that the pairs of lines are separated by at least four times the spacing of one pair; this also includes switching devices, since these are a common cause of unbalance and intercoupling. Acute turns, lines passing closely at oblique angles, and unequal-length jumpers all may cause unbalance. However, it is possible that there may be a loose or poor connection in the antenna circuit or transmission line that may be causing the unbalance. Even though there may be other factors contributing to the unbalance, it is almost

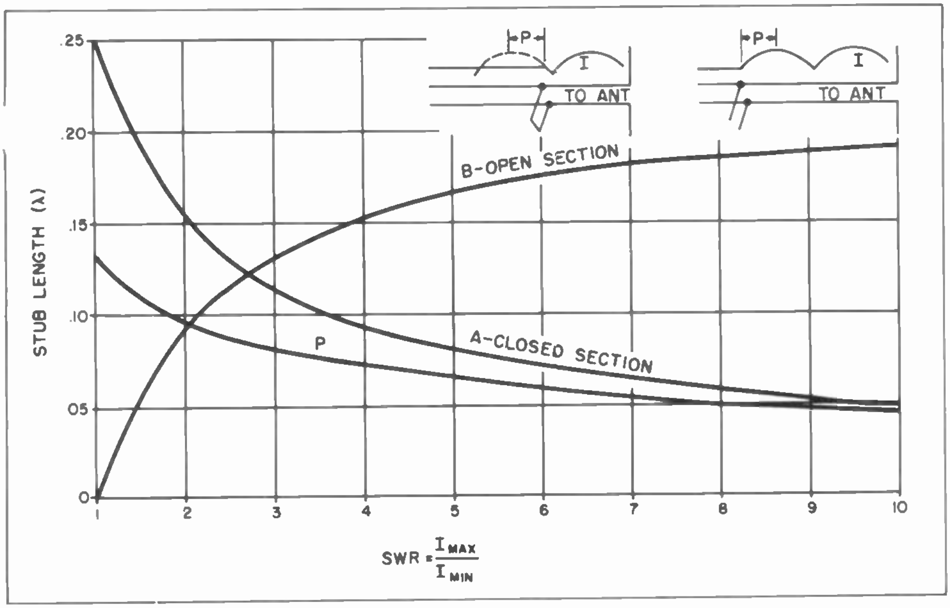


Figure 8. Chart for Determining Length and Position of Stub When SWR is Known

invariably true that the most serious cause of unbalance is at the transmitter or the transmission-line connections in and at the transmitter building.

Thus, a balanced transmission line must consist of conductors of equal size and length, evenly spaced throughout their run, and possessing equal capacitances to ground and to surrounding conducting objects. Since each trans-

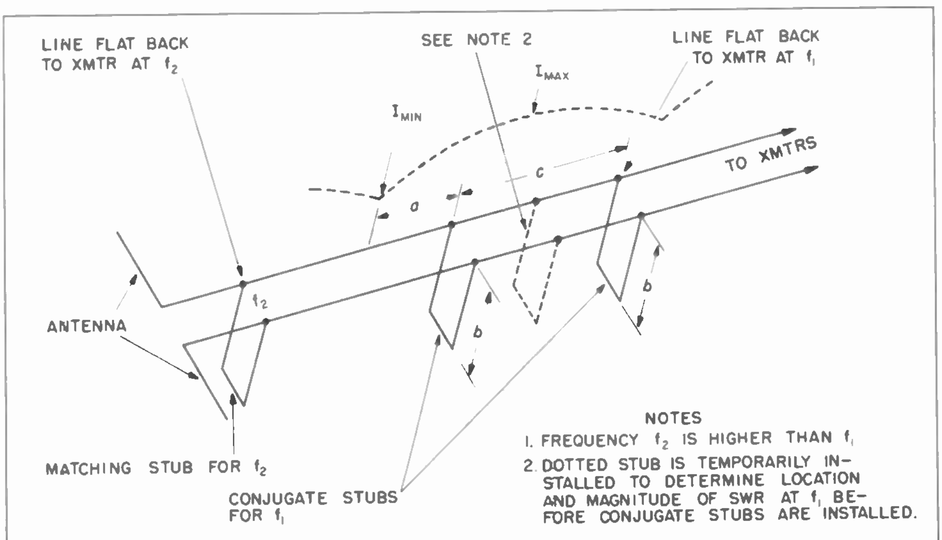


Figure 9. Sketch of System for Matching Antenna to Transmission Line at Two Frequencies (A simple, closed stub is used to flatten the line at highest frequency, f_2, and a pair of conjugate stubs to flatten at lowest frequency, f_1, without affecting the line at f_2. See figures 10, 11, and 12 for dimensions a, b, and c.)

mission line must be balanced before the antenna is tuned, the load may be simulated by a network of noninductive resistors (of sufficient power rating) connected in series-parallel to equal about 600 ohms.

MATCHING THE ANTENNA TO THE MAIN R-F TRANSMISSION LINE

No matter what its physical form, every antenna system will have a definite value of impedance at the point where the line is to be connected. Since the input impedance varies with frequency, the problem is to transform the antenna input impedance to the proper value to match the line at all frequencies to be used. In other words, the desired condition is one in which all power sent down the transmission line will be absorbed by the antenna, and none reflected back along the transmission line. This will occur only when the antenna input impedance is made to appear equal to the characteristic impedance (Z_0) of the line.

In a transmitting circuit, it is the antenna and only the antenna that determines the standing-wave ratio (SWR) on the line. No adjustments that can be made at the input end of the line can change the SWR, nor is it affected by changing the line length.

The SWR is the measure of the mismatch between the antenna and the line, and is the ratio of instantaneous maximum current or voltage to instantaneous minimum current or voltage, respectively. Thus:

$$SWR = i_{max}/i_{min} = e_{max}/e_{min}.$$

When the SWR is equal to one (unity), the line is perfectly matched, or flat; that is, there is no variation in current or voltage along the line.

There are many varieties of matching networks from which to choose the type that will fill the particular requirements of each installation. They include building-out sections (stubs), coplanar sections, conjugate networks, and reentrant networks, each of which will be discussed separately.

The impedance-matching element used must be located on the main r-f transmission line so that its own impedance in parallel with the line's impedance at that point results in a purely resistive impedance equal to the characteristic impedance of the line. Reflections and standing waves are thereby *eliminated* from the main line between the matching element and the *r-f source*. Since this does *not* eliminate standing waves from the section of the main line between the matching network and the load, the network should be located as close to the load as possible.

Building-Out Sections

When only one frequency is to be transmitted, it is usually best to use building-out sections (stubs) to match the antenna to the transmission line. These stubs are usually constructed from lengths of transmission line, and may be either closed or open. For mechanical reasons, and to reduce radiation losses, a closed stub is generally used.

Figure 8 is included for the purpose of determining the length and placement of the stubs in the main r-f transmission line when measuring from the I_{max} point nearest the load. If I_{max} is used as the reference point, an open stub would be placed in the line towards the transmitter from I_{max}, while a shorted stub would be placed towards the load from I_{max}.

Conjugate Sections

When two frequencies, f_1 and f_2, are to be simultaneously transmitted, two simple building-out sections cannot be used to flatten the line for both frequencies, because of the interaction be-

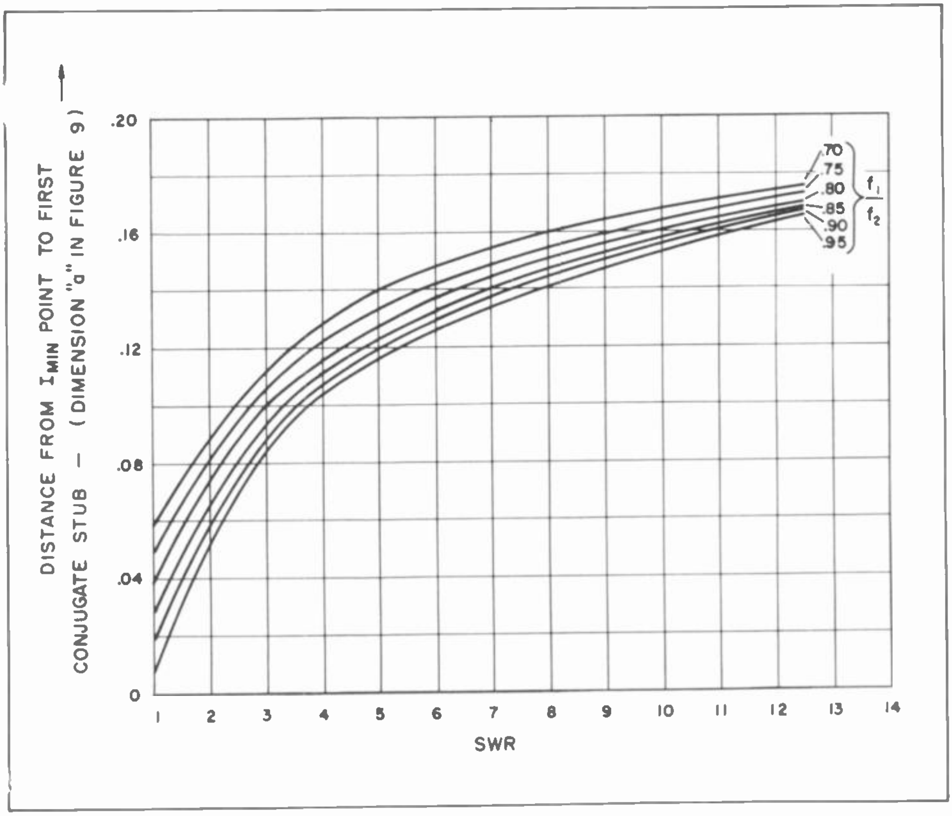


Figure 10. Chart for Determining Distance from I min Point to First Conjugate Stub (See figure 9.)

tween the two stubs at the two frequencies. In such a situation, however, it *is* practical to use a single matching section to flatten the line at the highest frequency, and to add an additional *pair* of stubs, spaced $\lambda/4$ wavelength apart (at the highest frequency), and so cut and positioned that their combined effect flattens the line at the second (lowest) frequency while apparently not existing at the highest frequency.

Such a pair of stubs are said to be *conjugate* to each other, meaning that they have equal and opposite reactive effects which cancel at one frequency (because of their $\lambda/4$ spacing at that frequency), while offering a net reactive effect just sufficient to flatten the line at a second frequency.

The mathematical design of conjugate building-out sections is quite complex, but a practical approach to the following cut-and-try method will result in an efficient matching network:

The antenna is matched to the transmission line at the higher of the two frequencies (f_2) using figure 8 to determine the size and placement of the stub.

The most accurate method for designing and installing the conjugate sections (after the line has been flattened at the higher frequency, f_2, as above) involves the attachment of a temporary building-out section to the line. The section is cut to the lower frequency, f_1, using dimensions obtained from the chart in figure 8, and placed on the transmitter side of the stub used to match the system for the higher frequency. The f_1 transmitter

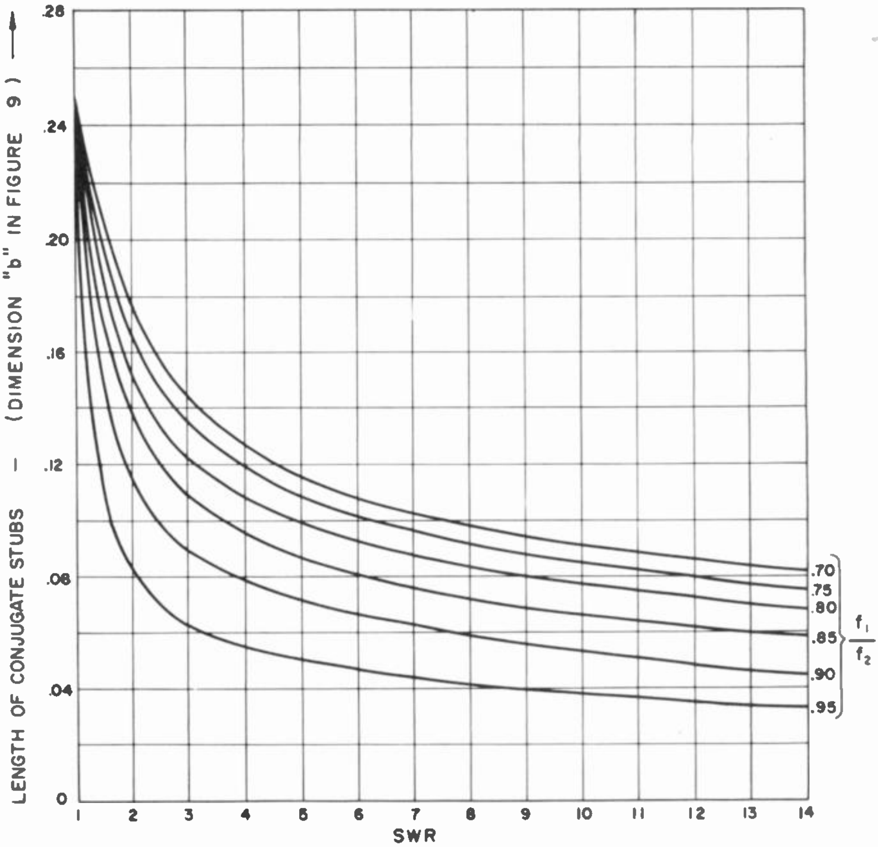


Figure 11. Chart for Determining Length of Conjugate Stubs (See figure 9.)

is then energized, and the temporary f_1 stub is adjusted until the SWR at f_1 is reduced to the minimum value.

The length of the temporary stub (including one-half the length of the short) is then measured with a tape, and the value converted to wavelength. For example: If $f_1 = 10$ mc. ($\lambda_1 = 100'$), and the measured length of the temporary stub = 6′ 7″, then, the stub length = 6′ 7″/100′ = $0.066\lambda_1$ at 10 mc.

Using the value ($0.066\lambda_1$), and referring to figure 8, it can be seen that the circuit is operating at a SWR of 7/1, and that the distance (P) of I_{max} from the point of attachment is $0.058\lambda_1$.

The temporary stub may now be removed, and its point of attachment used as a reference point for location of the I_{max} and I_{min} points. As found above, I_{max} is $0.058\lambda_1$ (5′ 8-1/2″) toward the transmitter from the point of attachment, and I_{min} approximately 25′ either side of I_{max}.

Continuing the example given above and assuming f_2 to be 13.33 mc., then the ratio f_1/f_2 = 10/13.33 = 0.75. This value determines which curve in the family of curves should be used in

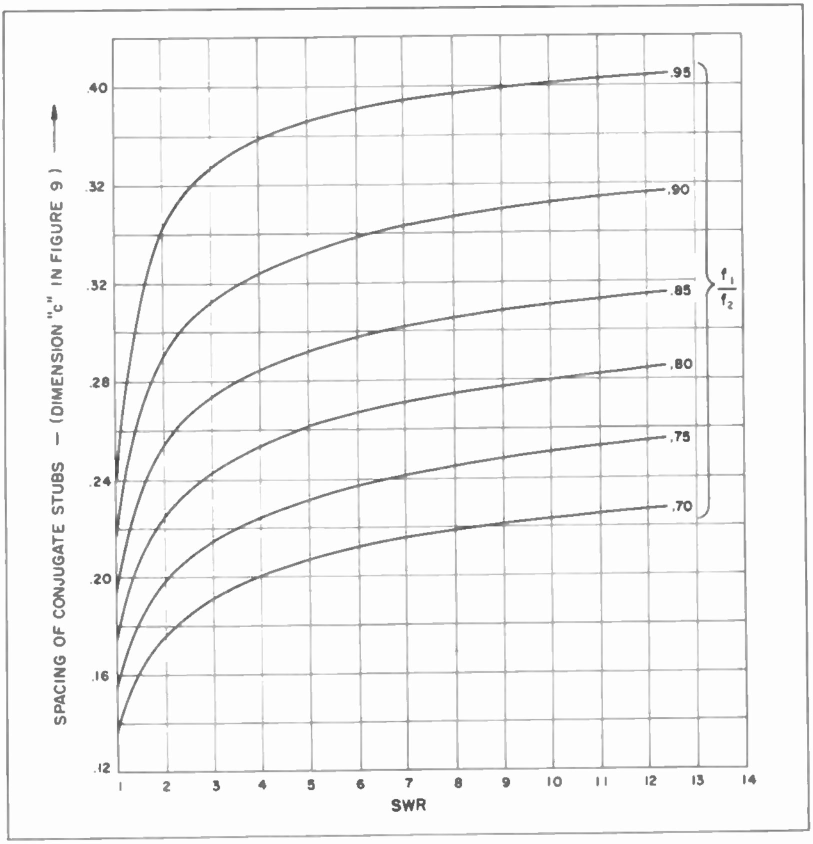


Figure 12. Chart for Determining Spacing between Conjugate Stubs (See figure 9.)

the charts (figures 10, 11, and 12) for constructing and positioning the conjugate sections. Completing the example, where $f_1/f_2 = 0.75$, and the SWR = 7, figure 10 shows that the stub nearest the antenna is placed $0.149\lambda_1$ (14′ 11″) from I_{min}. Figure 11 shows that the length of each section will be $0.096\lambda_1$ (9′ 7″ at 10 mc.) minus one-half the length of the short. Figure 12 shows that the second section should be installed $0.24\lambda_1$ (24′) on the transmitter side of the first section. (Figure 9 illustrates the dimensions referred to in figures 10, 11, and 12.)

After completing the installation, the SWR's should once more be measured, and any required fine adjustments made until the desired conditions are achieved. If the SWR is too great, a careful check should be made of all values used.

A less accurate but somewhat easier method for designing and positioning the conjugate sections is as follows:

The antenna is matched to the line at

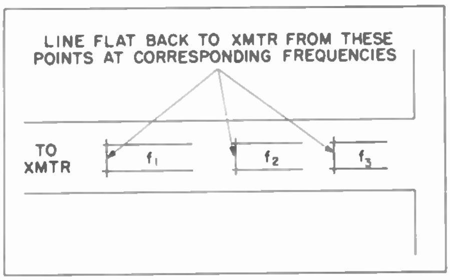


Figure 13. Schematic Diagram of Circuit in Which Inductively Coupled Matching Sections (Coplanars) are used to Match an Antenna to a Transmission Line at Three Frequencies

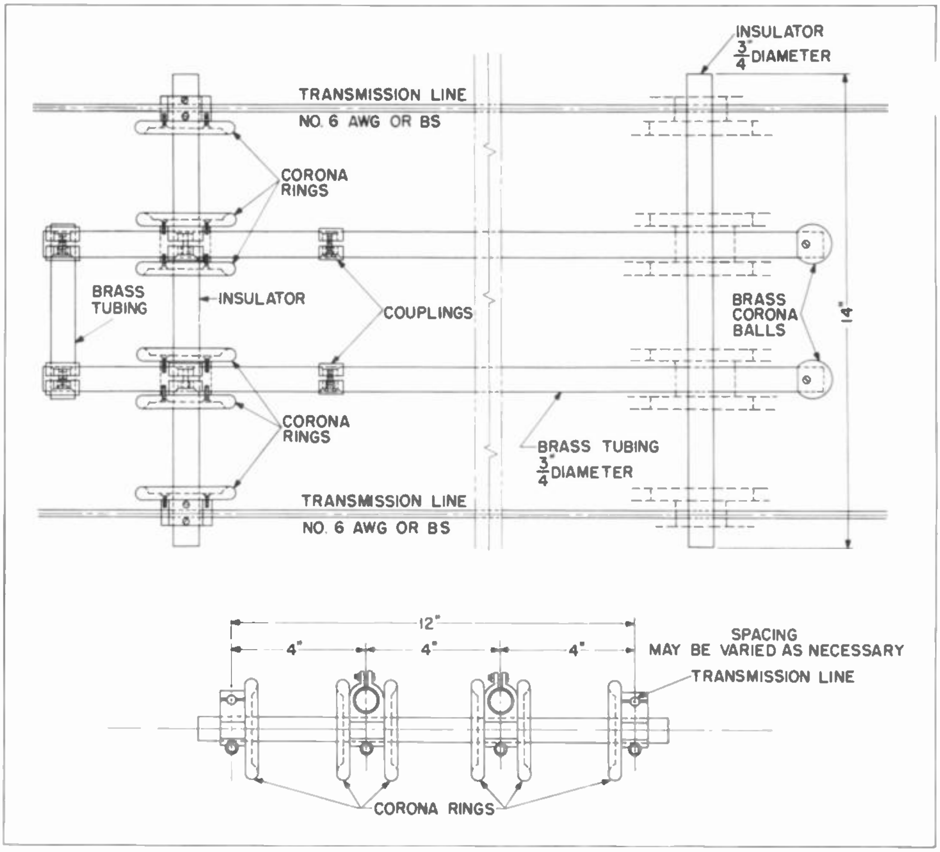


Figure 14. Construction Details of Coplanar Section (Top View of Mounted Section and Cross-Sectional View of One Support)

f_2 by means of a stub, as before (using figure 8).

The antenna is then energized at f_1 (the lower frequency), and the following data obtained:

1. Ratio of low frequency to high frequency (f_1/f_2)

2. SWR (I_{max}/I_{min})

3. Position of I_{max} and I_{min}

Then, using the values obtained and the charts shown in figures 10, 11, and 12, the dimensions, spacing, and positioning of the required conjugate sections are determined. Final minor adjustments for minimum SWR complete the installation.

Theory of Coplanar Coupling Section

Although the above-described conjugate networks are an effective means of obtaining simultaneous matching of an antenna and a transmission line at two frequencies, it is preferable, where feasible, to use resonant circuits, inductively coupled to the line, to accomplish the matching. The sections may be constructed either from transmission line or from tubing, and are variously called linear transformers or coplanars.

The feature which makes coplanars preferable to conjugate networks is the fact that no electrical connection exists between the line and the coplanar except at the frequency for which the length of the coplanar represents one-quarter wavelength. (The effect is similar to that of an absorption-type wavemeter, which extracts energy from an r-f field only when tuned to resonance.) Figure 13 shows schematically an antenna and a line matched for three frequencies by means of coplanars.

If the coplanar section is attached to an energized transmission line (very low power levels should be used) and a trolley meter is placed near and parallel to the short, a current will be found to be flowing in the short—a current which increases as the short approaches a resonant point (I_{max} point) on the line and decreases as the short is moved away from the I_{max} point. Furthermore, when the short is on the opposite side of the I_{max} point from the open end of the coplanar, the reactive effect on the transmission line is exactly the same as that of a capacitive stub. Similarly, if the short is moved to the other side of the I_{max} point (toward the open end of the coplanar) the reactive effect on the line is equivalent to that of an inductive stub.

Thus, it can be seen that the coplanar can be used in the same manner as a stub for matching an antenna to a line (tuning out the reactive components which produce standing waves). Likewise, the method for determining the position on the line for a coplanar is identical with that for building-out sections (stubs), i.e., the coplanar should be placed on the line with its short in almost exactly the same location as the attachment point for a stub used for the same purpose. (Figure 8 is used as in determining the position of a stub.)

Construction of Coplanar Sections

A typical coplanar section supplied in kit-form for use in CAA installations consists of 3/4-inch, thin-wall, copper tubing, together with the associated brass fittings and insulators necessary

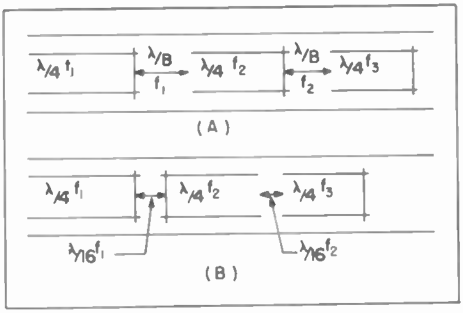


Figure 15. Minimum End-to-End Spacing for Adjacent Coplanar Coupling Sections

to permit the device to be properly spaced and hung in the proper plane in the main r-f transmission line.

If such kits are not available, the sections can be fabricated as shown in figure 14. The tubing can be aluminum, copper, or brass, and should be of lengths based on the desired frequency. The support insulators are the 12-inch spacers found in conventional transmission-line kits, or they may be other types of porcelain insulators. The supports for the tubing can be machined from copper, brass, or aluminum bar stock, and the supports for holding the insulators to the transmission line are machined from the same stock as the tubing supports. It is recommended that small pulley wheels be provided in the recesses which ride on the wire so that the whole unit can easily be moved along the line to facilitate tuning.

The element spacing of the coplanar shown in figure 14 is about 33% of the spacing of the transmission line conduc-

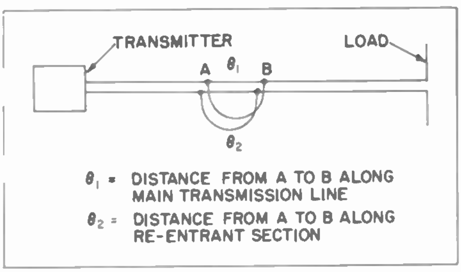


Figure 16. Simple Re-Entrant Network

tors—this figure is acceptable as long as the frequencies are in the region of within 10% to 15% of each other. However, if the frequencies have a separation of 25% or more, the elements of the coplanar should be spaced about 60% of the spacing between the transmission line conductors.

The length of the sections may be determined by the formula:

length (ft.) = 246K/F (mc.)

where K is a constant which depends on the type of material used for the coplanar. If parallel wire is used, K is approximately 0.975, and for tubing, 0.95. (By cutting the sections slightly longer than above, they can be adjusted to cover a variety of closely adjacent frequencies, since the position of the short determines the effective length of the section.)

Adjustment of Coplanar Sections

If, after a coplanar whose short is on the opposite side of I_{max} from the open ends has been installed on the line, a current *maximum* is found to exist a quarter wavelength toward the transmitter. the coplanar should be moved a few inches closer to the transmitter and the short moved toward the resonant point a slight amount (approximately 1/4 inch), in preparation for the next trial. On the other hand, if a current *minimum* is found to exist a quarter wavelength toward the transmitter from the location of the coplanar, the opposite procedure should be followed.

Slight adjustments should then be made until the SWR of the line is reduced to 1.1/1. or less.

After the matching coplanar has been installed for the highest frequency to be used, the next lower frequency should be applied to the line and the tuning procedure repeated. The second coplanar should be placed on the transmitter side of the first, not too close to the first, preferably not less than 1/2 the length of the longer section. (See figure 15.) If this distance cannot be obtained, the section being added should be oriented so that like ends of adjacent sections face each other. Even then, they should not be nearer to each other than 1/4 of the length of the longer section. It is of interest that it makes no difference which way the sections face in the line, so long as the minimum-space requirements stated above are met.

Re-Entrant Network

A fourth method of achieving a match between the impedance of an antenna and that of a transmission line involves the use of the re-entrant network. This device, as its name implies, is simply a section of transmission line that taps off a transmission line at one point and then re-enters the line at a second point, as shown in figure 16.

Since this type of network is used ordinarily only to prevent interaction between two or more transmitters feeding a common transmission line, a thorough discussion of its properties is reserved until the second half of this article, which deals more completely with systems used for that purpose. (Actually, the only situation in which the re-entrant network would ever be used instead of stubs or coplanars for matching an antenna to a line, would be in very-high-powered systems where the extremely high r-f voltages and resultant corona which would be found at the open ends of those elements which might prove undesirable.)

The re-entrant network is very easily understood if its operation is compared to an equivalent shorted or open stub. In fact, its effect on the line can be directly compared with that of the stub.

Consider first the circuit of figure 16, in which the current from the transmitter divides equally at point A. O_1 and O_2 are equal in length, so the currents arrive at B in phase; therefore, the attachment of O_2 to the line produces no effect on the line other than a slight standing wave due to the mismatch created by the branching of the line. On the other hand, if O_1 or O_2 are transposed, or if O_1 and O_2 are made to differ in length by exactly 1/2 wavelength, then the signals cancel at B, effectively placing a short circuit across the line at B.

It can be seen that these two conditions compare with attaching, at point B, shorted stubs of lengths $\lambda/4$ (or $3\lambda/4$) and $\lambda/2$, respectively. It can further be seen that all values of impedance and reactance which can be achieved by the use of stubs can also be achieved with properly proportioned re-entrant networks.

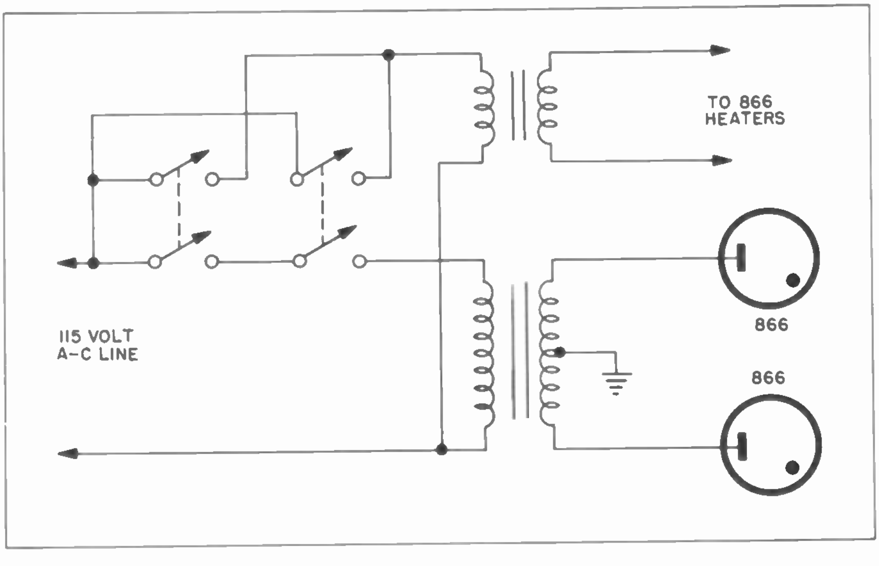
Likewise, all the data presented earlier in this article pertaining to the use of matching stubs, either single or in conjugate pairs, can be applied to the use of re-entrant networks for the same purposes. All necessary additional information on the use of re-entrant networks, such as determination of the positioning, length, and spacing of conjugate pairs, will be found in the second half of this article.

(TO BE CONCLUDED)

Solution to . . .

Last Month's "What's Your Answer?"

Using two conventional d-p-d-t switches, the primary of the circuit is wired as shown.



TRANSMITTER MONITOR

by Harold Gullstad
Philco Field Engineer

A practical, field-tested device for simplifying transmitter maintenance and testing. All necessary construction data and operating instructions are provided.

THE INSTRUMENT described in this article was designed and constructed by the author as the result of a need for a simple, quick means of trouble-shooting speech-modulated transmitters. An important advantage of this instrument over conventional methods is that it allows checking the output from the transmitter antenna rather than at some point preceding the antenna. It is thus possible to determine accurately what is being transmitted.

The important transmitter checks which can be made with this instrument are: (1) approximate frequency output, (2) relative power output, (3) modulation percentage, and (4) aural checking of the modulating signal. In addition, the circuit provides a means for checking microphones.

CONSTRUCTION AND THEORY

This instrument is quite simple in construction, and utilizes conventional parts which can be readily obtained. In the test model, the components are mounted on a 4″ x 7″ chassis, and the entire unit is housed in a cabinet measuring 5″ x 6″ x 9″. Figure 1 shows a front view of the cabinet, and illustrates the arrangement of meters, controls, and jacks. Several binding posts (use described later) are mounted on the rear panel in a convenient manner. Figure 2 shows the schematic diagram of the circuit used in the test model. The values are shown for all commercially available parts, while construction data is provided for the r-f coils.

The transmitter signal is coupled to the unit through T_1, the secondary of which is tunable to the incoming frequency by C_1, a front-panel adjustment. The amount of coupling between the transmitter antenna and the unit is controlled by the length and location of the

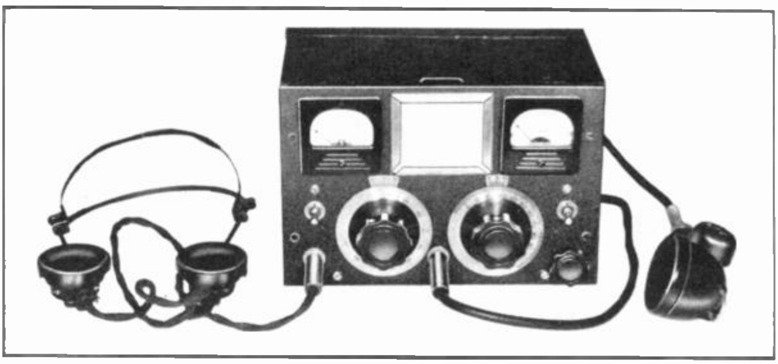


Figure 1. Front-Panel View of Test-Model Transmitter Monitor (The power-input cable and the three binding posts used for various transmitter tests are on the back of the unit.)

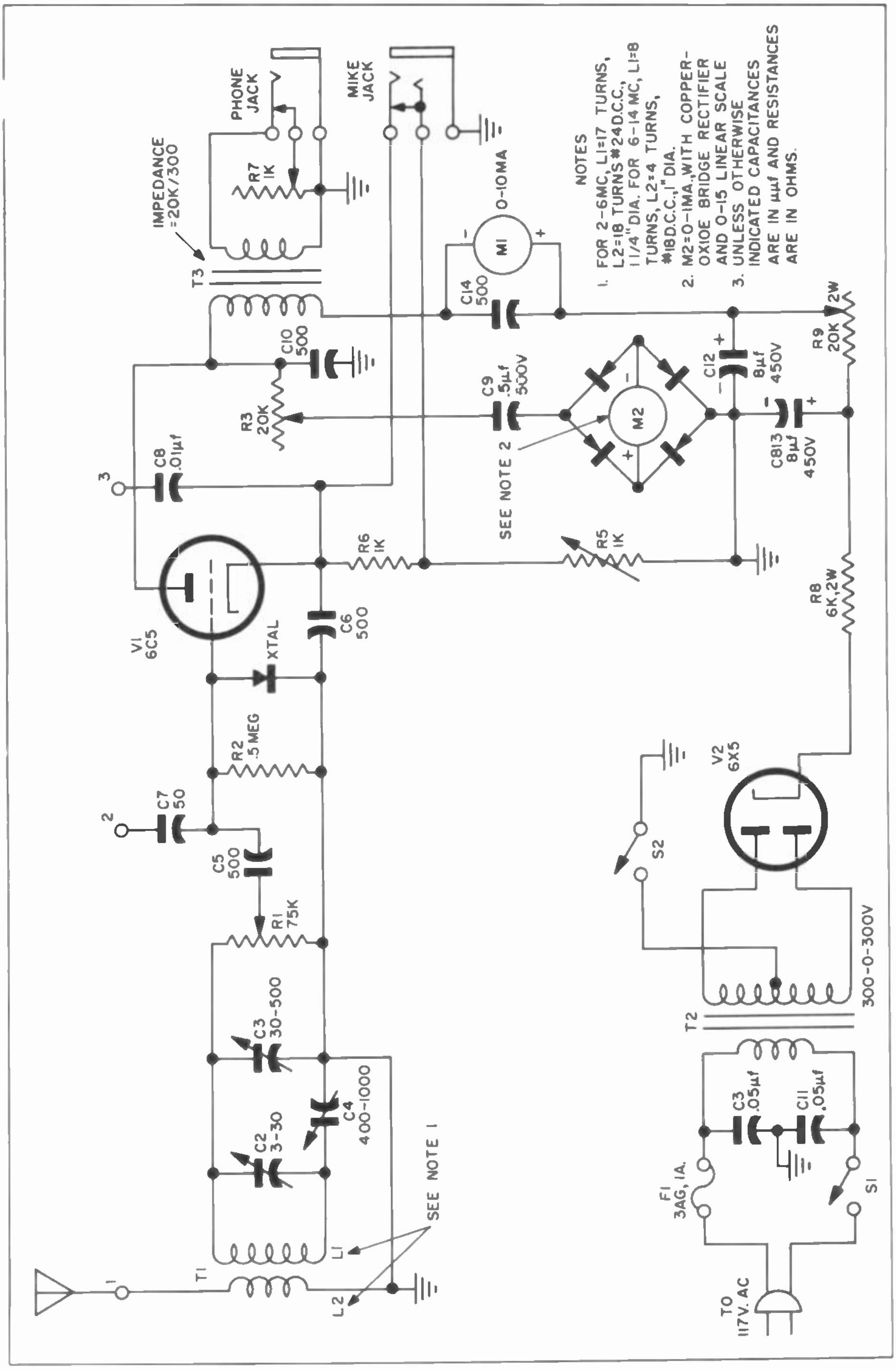


Figure 2. Schematic Diagram of Transmitter Monitor (The numbers of turns indicated for L_1 are approximate. It is suggested that L_1 be wound with 20% additional turns, and then trimmed as necessary.)

coupling antenna connected to the unit. An antenna (binding post 1) is provided on the rear of the cabinet. The frequency range of the input circuit is 2 to 6 mc., with the coil used; however, other coils may be designed to provide operation on other frequencies. Trimmer capacitor C_2 is used to align the input circuit at the high-frequency end of the operating band, while padder C_4 adjusts the input circuit at the low end. R_1, another front-panel adjustment, determines the amplitude of r-f signal that is applied to the crystal detector. The modulated r-f signal is rectified by the crystal detector, and the rectified output is amplified by V_1. A pair of headphones, which are plugged into the jack across the secondary of the output transformer T_3, provide aural monitoring of the modulation signal. Milliammeter M_1, in the plate circuit of the amplifier, provides an indication of the amount of r-f power applied to the detector. The amount of r-f power applied is controlled by the setting of R_1, which is calibrated. Thus, by reference to a calibration chart, the resonant frequency of the input circuit can be roughly determined for various settings of C_1. (With an r-f signal applied, C_1 is adjusted for a dip on meter M_1.) An a-c voltmeter, which consists of a variable multiplier, a bridge rectifier, and an 0-to-1 ma. meter, is effectively connected across the primary of output transformer T_3, and is used to indicate modulation percentage. With the proper adjustments, an indication of 10 on M_2 corresponds to 100-percent modulation (when the correct amplitude of r.f. is applied to the crystal detector, as determined by the indication on M_1).

Binding posts 2 and 3 provide connections to an oscilloscope for visual observation of the modulation trapezoid. Binding post 2 can also be used to provide a signal for visual observation of the r-f envelope pattern, as well as for application of an external r-f signal to zero beat with the applied transmitter signal for more accurate determinati of the transmitter frequency. The m jack is provided for microphone checking, both for condition and for fidelity of reproduction. The detailed procedure for making the various checks is given later in this article.

CALIBRATION

Since various adjustments and calibrations are necessary before the instrument can be properly used, a brief description follows—be sure to make the adjustments in the order given.

Calibration of C_1

Approximate frequency checks can be made if the resonant frequency of the input circuit is known for any setting of C_1. The frequency range of the test model (with the coil used) is from 2 to 6 mc. The use of a grid-dip meter is recommended for making frequency checks of the input circuit.

1. With C_1 set for maximum capacity, adjust padder C_4 so that the input circuit is resonant to the lowest desired frequency (2 mc.).
2. With C_1 set for minimum capacity, adjust trimmer C_2 to resonate the input circuit to the highest frequency desired (6 mc.).
3. Plot the resonant frequencies of the input circuit for a number of intermediate settings of C_1, and draw a calibration curve.
4. When making successive frequency checks, this curve is referred to in order to establish the approximate frequency to which the circuit is tuned.

Adjustment of R_5

For making microphone checks, R_5

must be adjusted for correct resistance.

1. Apply power to the unit and allow a reasonable warm-up period.
2. Plug a dummy microphone plug (which is shorted between ring and sleeve) into the mike jack.
3. Adjust R_4 for an indication of 10 ma. on M_1.
4. Plug a microphone (known to be in good condition) into the mike jack.
5. Adjust R_5 for an indication of 5 ma. (mid-scale) on M_1.
6. Recheck R_4 and then R_5. This final setting of R_5 should remain permanently fixed.

Calibration of R_1

Approximate-power-input checks can be made if R_1 is properly calibrated and the coupling between the transmitter antenna and the monitor remains constant.

1. Set R_1 to zero, and adjust R_4 for an indication of 10 ma. on M_1.
2. Set R_1 to its mid-scale value (reading of 50 on the dial).
3. Tune C_1 for a dip on M_1.
4. Adjust the coupling between the transmitter antenna and the monitor unit (vary length of monitor coupling antenna) to obtain an indication of 7 ma. on M_1.
5. Decrease the transmitter output in definite steps. measure its power output (with a reliable power-measuring device), and record the power and the settings of R_1 when an indication of 7 ma. occurs on M_1.
6. Increase the transmitter output and proceed as in step 4.
7. Draw a power curve by plotting the power output from the transmitter for various settings of R_1 (when an indication of 7 ma. occurred on M_1).

Adjustment of R_7

This resistor must be adjusted to reflect the same a-c impedance into the triode plate circuit as the head phones that are to be used with the monitor.

1. Plug a pair of head phones (of the type to be used with the unit) into the phone jack, and with a modulated r-f signal applied to the unit, observe the indication on M_2.
2. Adjust R_7 (with head phones removed from unit) to produce the same indication on M_2 as in step 1.

Adjustment of R_3

This adjustment is necessary to produce a standard indication on M_2 for a definite percentage of modulation of the r-f input signal.

1. Set R_1 to zero, and adjust R_4 for an indication of 10 ma. on M_1.
2. Apply a 100-percent-modulated, sine-wave, r-f signal to the unit.
3. Set R_1 to mid-scale.
4. Adjust C_1 for a dip on M_1.
5. Adjust R_1 to obtain a reading of 7 ma. on M_1.
6. Adjust R_1 to obtain an indication of 10 on M_2.

OPERATION

The unit is now properly calibrated and adjusted. The remainder of the article is devoted to a discussion of the procedures for making the various checks that are possible with this instrument.

Checking Modulation Percentage, Approximate Frequency, and Power Output

CAUTION: Monitor must have the same degree of coupling to the transmitter antenna as during the calibration procedure.

1. Adjust R_4 until M_1 indicates 10 ma.
2. Set R_1 to mid-scale.
3. Key the transmitter.
4. Tune C_1 for a dip on M_1.
5. Adjust R_1 until M_1 indicates 7 ma.
6. Compare the settings of R_1 and C_1 with the calibration charts. The approximate frequency and power output are thus determined.
7. Speak into the microphone and read modulation percentage on M_2. 100-percent modulation is indicated by a reading of 10 on this meter.

Accurate Frequency Check

1. Adjust monitor as in approximate-frequency check, above.
2. Connect r-f signal generator to binding post 2 of monitor. Adjust r-f generator to approximate frequency indicated by setting of C_1.
3. Vary r-f generator frequency to obtain zero-beat in head phones.
4. Read transmitter frequency on r-f generator dial. (The accuracy of this measurement is roughly equal to the generator accuracy.)

Observation of Trapezoidal Pattern

With unit adjusted as in frequency checks, connect the output of binding post 2 to the vertical input of an oscilloscope and the output of binding post to the horizontal input of the osci scope.

Observation of Modulation Envelope

Connect output of binding post 2 to the vertical input of the oscilloscope, and use internal horizontal sweep.

Microphone Check

1. Set R_1 to zero.
2. Adjust R_4 to obtain a reading of 8 ma. on M_1 (with the microphone out of the circuit).
3. Plug microphone into the circuit; M_1 should indicate 5 ma. if microphone is normal. (R_5 was previously adjusted for this indication during initial calibration of the unit.)
4. Press push-to-talk button; M_1 should indicate between 8 and 9 ma. if microphone is normal. (If microphone element is shorted, M_1 indicates 10 ma. with push-to-talk button depressed. If element is open. M_1 indicates 6 ma. with push-to-talk button depressed.)
5. Microphone reproduction can be monitored by use of the headphones used with the unit.

CONCLUSION

From the foregoing discussion, it can be seen that an instrument of this type provides (after it is properly calibrated) a simple and quick method of checking the condition of the transmitted signal. Its use should cut on-the-air time required for radio checks to a minimum, and maintenance personnel will find this unit a valuable addition to the transmitter site.

AN EMERGENCY TELETYPE RELAY SYSTEM

by Perry S. Gaye
Philco Field Engineer

A simple switching system for converting a primary teletype station into a relay between two other stations.

THE SYSTEM described in this article makes it possible to continue normal communication between two teletype stations in the event of failure of the land-line facilities between the stations, by conversion of a third station (which is still in communication with both) into a relay. Figure 1 shows a block diagram of a system of three inter-connected terminal stations. Without the system described here, failure of the land-lines between any two of the stations would necessitate resort to back-up facilities, such as h-f or f-m radio links, until normal operation could be resumed. With this circuit however, communication can be restored between the two stations in less than one second (the time required to throw one bank of ganged switches).

It can be seen in figure 1 that if circuits 3 and 4 are broken, for example, all communication between stations B and C is lost. Figure 2A is the basic block diagram for normal operation at station A, while figure 2B is the basic block diagram for that station when operated as a relay. Examination of the circuit of figure 2B shows that the two power supplies are placed in series in each circuit, but that in each case the series-circuit resistance is also doubled by the addition of the second machine. Further examination shows that any message received from station B on receive-machine A is also sent on to station C by send-machine C, and that any message received from station C on receive-machine D is sent on to station B by send-machine B.

The one disadvantage of this system is that when stations B and C are both sending, all four machines at station A are tied up; however, this situation sel-

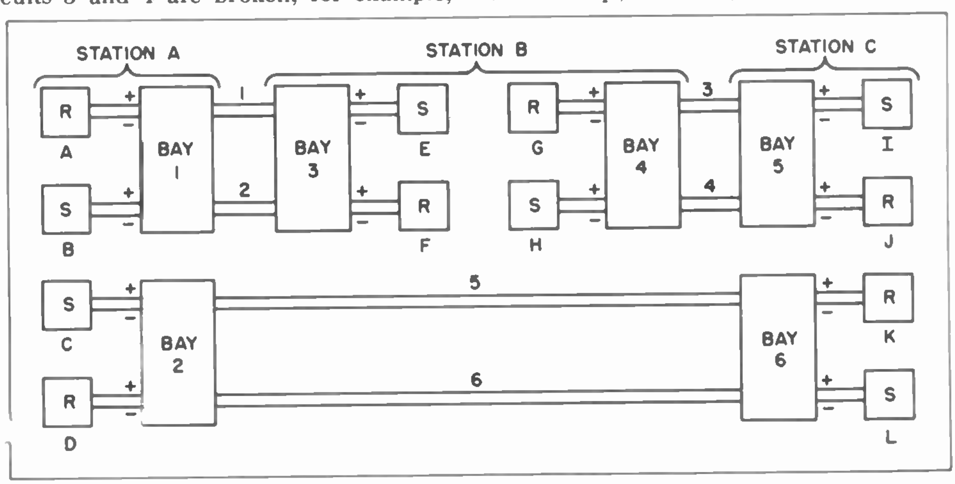


Figure 1. Block Diagram of Typical Three-Station Teletype System

dom occurs. Figure 3 shows the method of wiring each station for this emergency system. Five d-p-d-t switches (or one five-pole-two-position rotary switch) are used at each station—no other circuit components are required. If five separate switches are used, they should be installed so that they can be thrown simultaneously, in order to meet the time requirements specified by AACS regulations, and to reduce possible errors by operating personnel.

Correct polarities as indicated in figure 2B must be observed; otherwise, the machines will run open. The switches required for this modification are available from USAF supply channels, and conventional pushback or spaghetti-type wire (18-20 gauge) may be used for all interconnections.

Tracing the circuit of figure 3 and referring again to figure 2B, it can be seen that with the switches in the "down" position, normal operation is provided, while with the switches in the "up" position, two sets of series circuits are provided, each consisting of two machines and two batteries, and connected in such a way that the two remote stations are in full communication.

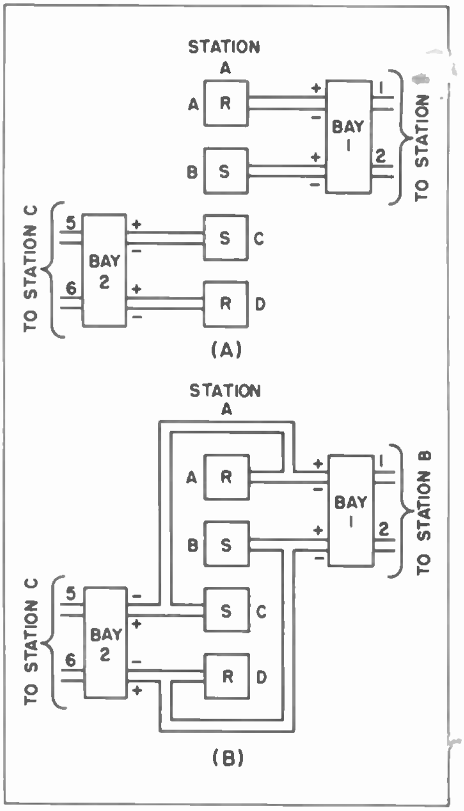


Figure 2. (A) Block Diagram for Normal Operation at Station A (B) Block Diagram for Relay Operation at Station A

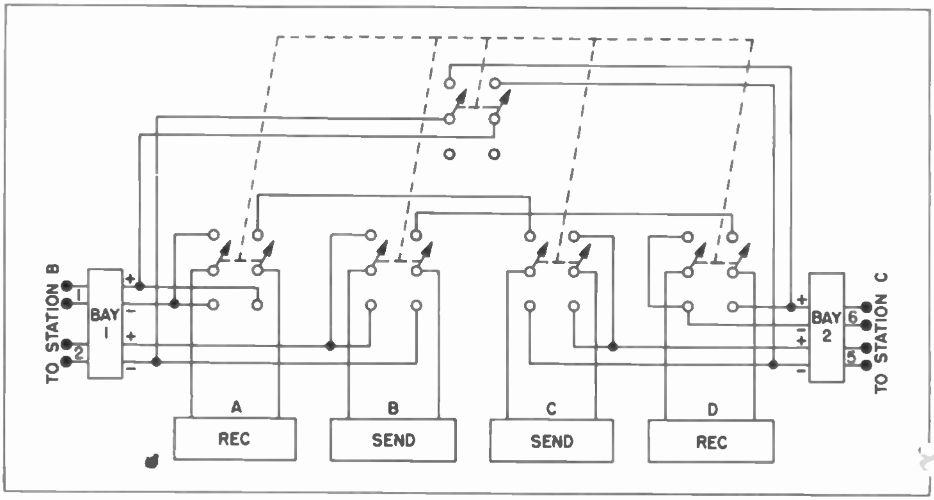


Figure 3. Schematic Diagram of Switching Circuit Installed at Each Station (Circuit shown is for station A; however, wiring system is identical at all three.)

PHILCO

TECHREP

WORLD WIDE

SERVICE

AN-299

Printed in U.S.A.