M/Sgt W. F. Pudlik

T/S Pudlik

PHILCO

TECHREP DIVISION BULLETIN

1. Waveguides
2.

AUGUST
and
SEPTEMBER
1951

Technical Information Requests

A new procedure for the handling of technical information requests from the field is now in effect at TechRep Division Headquarters.

In the past, such requests either were directed to the Correspondence Department, or were included in letters to other departments. This has sometimes resulted in considerable delay in answering because of the rerouting necessary. Also, personnel in other divisions of Philco have been considerably inconvenienced by field engineers who have written directly to the Industrial, Engineering, or Research Divisions for technical data which could have been furnished directly from our own Headquarters.

In the future, all field requests for technical information (other than publications) should be addressed to the "Technical Information Section, Technical Department," at Headquarters. This specifically includes requests for information on microwave communication problems. Under no circumstances are field personnel to address such requests to the Industrial, Engineering, or Research Divisions.

All field requests for publications, including training manuals, trouble-shooting manuals, military publications, and Philco commercial equipment bulletins, should be addressed to the Correspondence Department, as in the past.

④ PHILCO

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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.

Letters to the Editor

In the correspondence we are now receiving, there is an increasing number of constructive suggestions concerning further improvement of the BULLETIN. We sincerely welcome these suggestions, and hope that many more readers will express their opinions in a like manner. The following are excerpts from letters received since the last issue went to press:

●

"My copies are as dog-eared as a comic book on Bougainville, which is testimonial enough for a technical publication."

Ralph R. Saylor,
Philco Field Engineer, ADC.

●

"If suggestions are in order, a touch of humor now and then would make the BULLETIN an even better publication."

Francis E. Gates,
Philco Field Engineer, WADF.

(Really good humorous items are difficult to obtain, Francis, but wait until you see the October issue. Ed.)

●

"Your faces must be very red after mixing figures 2 and 3 on pages 13 and 15, in H. W. Merrihew's article, 'Nomography,' in the June BULLETIN. I am certain you will hear about that from more remote corners than this one in Korea. However, it's a little unjust to criticize on this small point without stating that, in just three issues, the BULLETIN has proved itself very valuable. All the electronics people in this outfit (and there are quite a few) have read it with both interest and profit. Each department head has found at least one article which has been applicable to a problem within his department."

John H. Rogers, Captain, USMCR,
Marine Ground Control Intercept
Squadron 1.

(They were, and we did. Ed.)

●

"In view of the recent emphasis on 'systems engineering,' I recommend that the BULLETIN include, or even emphasize, articles on that subject."

Carl Eastman,
Philco Field Engineer, AACS.

(We agree, Carl; we are looking for good articles on systems engineering. Ed.)

●

"The BULLETIN is an outstanding publication, and every page is crammed full of valuable information. I will be looking forward to every issue."

1st Lt. Wilfred W. Helm,
1936th AACS Squadron
(APO 406).

●

"With reference to your fine article, 'Voltage Regulators — Vest-Pocket Style,' in the July issue—page 22, second paragraph, eighth line, should read: 'the OB2 and OA2 miniatures,' etc. This change would match each tube type to its correct regulating voltage."

T/Sgt. Lyle, Hq. Squadron,
EADF, Stewart AFB,
Newburgh, N. Y.

(You are right, Sgt. Lyle — thank you. Ed.)

●

"I wish to add my heartiest endorsement to the many letters of appreciation you have received for the fine articles and work of the Philco TechRep Division BULLETIN. We number eighteen Tech Reps here at Erding Air Depot, and we are engaged in a multitude of diversified assignments and projects; every article is utilized to full advantage to accomplish our mission here in Germany."

John L. Herre,
Philco Team Leader.

Editorial

PUBLICATION OF CLASSIFIED MATERIAL

By Robert L. Gish, Managing Editor

Since the production of the BULLETIN began, we have received a number of suggestions that we publish classified material.

We realize that certain advantages would accrue if we did publish classified articles. For example, a number of individual field activities would certainly benefit, if we published a comprehensive article on the operational characteristics of specific heavy-ground-radar equipments under various siting conditions. Likewise, a detailed analysis of the circuitry involved in specific guided-missile control devices would be of extreme interest to many BULLETIN readers.

On the other hand, however, to publish such articles would require that we restrict our mailing list to only those persons for whom appropriate security clearances are on file in our Headquarters offices, and whose duties require that they be given such information. Furthermore, the copies which we could mail under those restrictions would have to be handled in accordance with all other requirements of security regulations, including storage in a locked safe. Obviously, such a limited distribution would defeat the purposes for which the BULLETIN was created: namely, to increase the coordination and exchange of technical information between Philco Field Engineers and their military and civilian associates, and to achieve the maximum possible distribution to industry, of the technical knowledge developed through the field experience of thousands of Philco men.

There is an almost inexhaustible fund of technical data which is not classified, on which BULLETIN contributors can draw for articles; therefore, limitation to this subject material will not impose any serious restrictions on our contributors.

After considering the problem from all angles, we have established a policy of printing in the BULLETIN only those articles which require no classification. To those potential contributors who have written or planned to write articles which deal with classified equipments or techniques, we suggest that the material be discussed with the appropriate security personnel before it is submitted to this magazine. In many cases, it may be possible to rewrite the material in such a way that it discloses no classified information, or to obtain specific permission to publish it. In those cases where a decision cannot be obtained locally, we will initiate the necessary correspondence from TechRep Division Headquarters.

An Interference-Blanking System for MTI-Modified Radars

By James B. Hangstefer
Former Philco Field Engineer

Theory and circuit data for a device which eliminates delayed "main-bang" interference caused in MTI-modified radars by associated radar equipment operating in the same area.

(Editor's Note: The system described in this article was developed and tested by the author about a year ago, in Europe. The prototype test was so successful that Headquarters, USAFE, forwarded the idea to Air Materiel Command, for possible use throughout the Air Force. This is another example of the excellent field-engineering services Philco men are providing to the U. S. Armed Forces.)

This system has been developed to allow auxiliary radar equipment to operate at the same site location as an MTI-modified search radar. No equipment modifications are required.

This article describes a prototype system. The initial field tests, with an MTI-modified AN/CPS-1 (MX-563C) operating side by side with an AN/CPS-4 height finder, were entirely successful.

Since the operating characteristics (PRF) of the MTI equipment determine the operating characteristics of the blanking system, the exact circuit-component values used in the system described here are not necessarily applicable to MTI equipments having different characteristics (PRF). However, the method of blanking is applicable, and only slight modification would be required for its use on other MTI systems.

THE BASIC INTERFERENCE PROBLEM

Transmitter radiations from an associated radar set (AN/CPS-4) are received in the AN/CPS-1, and are presented on its indicators as unsynchronized pulse interference. This "main bang" interference, which greatly reduces the operating efficiency of the

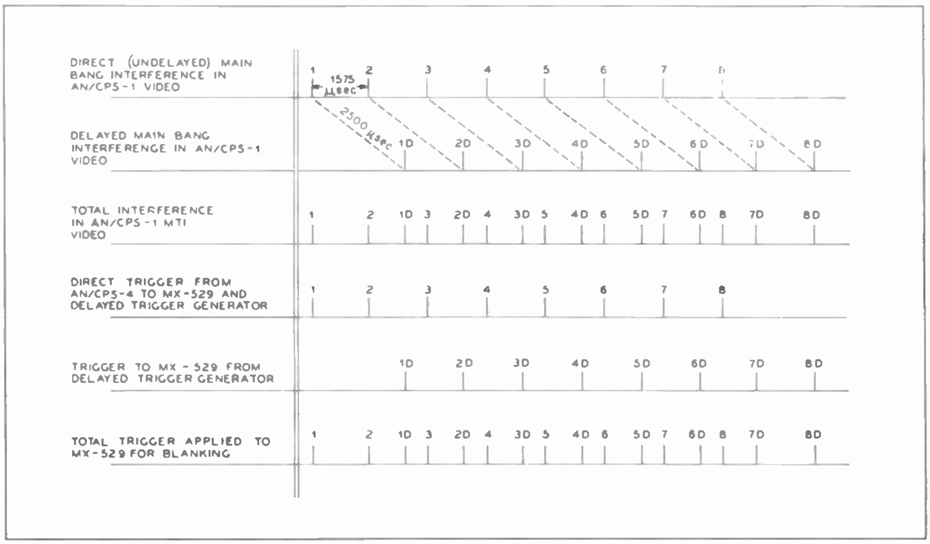

Figure 1. Waveform Analysis of Interference-Blanking System

AN/CPS-1 is the result of: (1) the operation of the two radars within the same frequency band. (2) the extremely high transmitter pulse power of both sets, and (3) the operation of the two units close together (within a few hundred feet).

In a conventional AN/CPS-1 installation (without MTI), this main-bang interference is removed by Interference Blanker MX-529/U. The interference blanker is placed in series with the AN/CPS-1 receiver output, between the receiver and the indicators. A trigger from the interfering radar (AN/CPS-4) is coupled into the blanker. The blanker then prevents video signals from passing to the AN/CPS-1 indicators during the period of the trigger. Since the trigger pulse is coincident with the main bang causing the interference, no interference pulse will be passed on to the indicators. (Ref. T.O. 16-35MX-529-3).

In an MTI-modified AN/CPS-1 (MX-563C), the interference problem becomes more complex. In addition to the main-bang interference described above, "delayed-main-bang" interference is also present. This delayed interference is brought about by the passage of MTI video through a 2500-microsecond supersonic delay line (see figure 1). All of these interference pulses must be removed in order to obtain satisfactory operation of the AN/CPS-1 (MX-563C).

MTI-INTERFERENCE BLANKING

Interference-blanking in the MTI equipment is accomplished with the aid of two units: (1) Interference Blanker MX-529/U, and (2) a delayed-trigger generator (see figures 3 and 4). The first unit is a standard Air Force component, while the second one is a proposed new unit, essential for MTI-interference blanking.

A trigger from the interfering radar is coupled to both the interference blanker and the delayed-trigger generator. Since this trigger is coincident with the direct main-bang interference, the blanker will function to prevent the direct main-bang interference from reaching the indicators. The delayed-trigger generator produces an output trigger that is delayed from the originating trigger by 2500 microseconds. This trigger output is coincident with the delayed main-bang interference, and, since it is also coupled to the interference blanker, the delayed main-bang interference cannot pass to the indicators. The removal of these two series of interference pulses allows completely normal operation of the AN/CPS-1 (MX-563C). (See figures 1 and 2.)

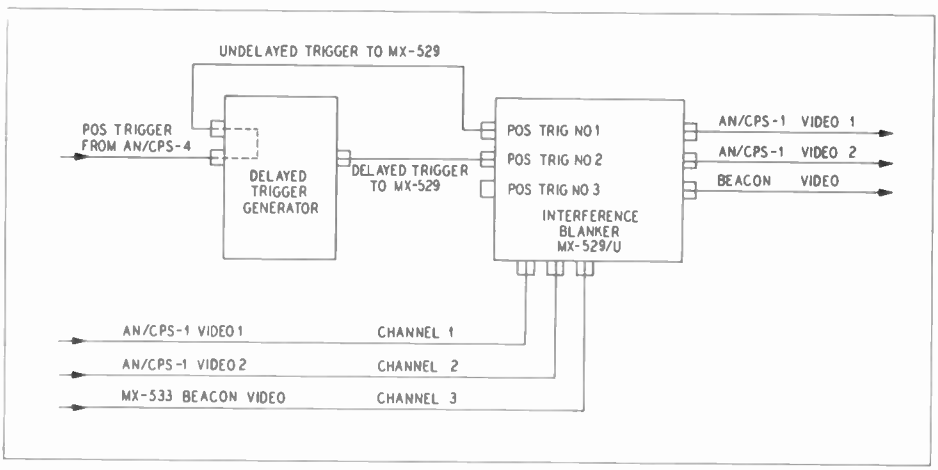

Figure 2. Cabling Diagram of Interference-Blanking System

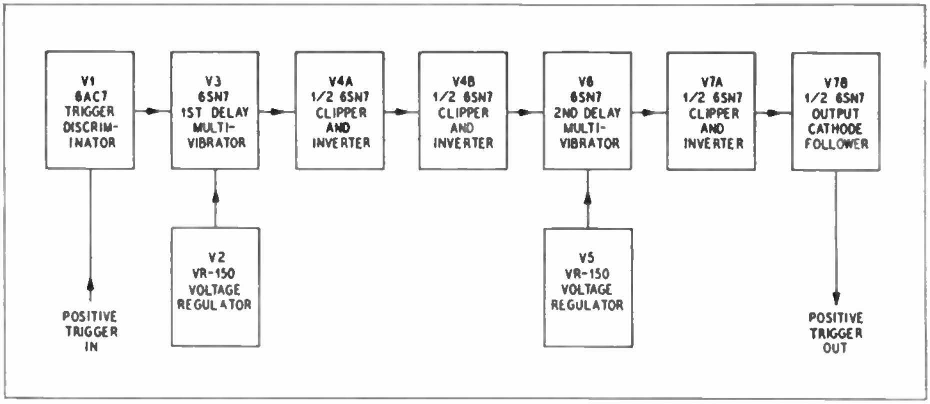

Figure 3. Block Diagram of Delayed-Trigger Generator

No appreciable loss of sensitivity in the MTI equipment is encountered, because of the random nature of the interference pulses with respect to the MTI equipment.

DELAYED-TRIGGER GENERATOR

Two delay multivibrators are used in the delayed-trigger generator, in order that input triggers with repetition rates up to 750 p.p.s. may be used. A positive trigger with an amplitude of approximately 20 volts is applied to the input-trigger-discriminator stage, V1. (See figures 4 and 5.) This tube is negatively biased to prevent spurious voltages from triggering the system. The positive trigger is inverted to a negative pulse in the plate circuit of this stage. This negative pulse is coupled directly to the first delay multivibrator, V3.

V3 is a conventional flip-flop multivibrator, with the grid of V3b returned to its cathode, and operating at zero bias. The plate current drawn by V3b through the common cathode resistor is sufficient to produce cutoff bias on V3a. Until an external voltage is applied to the stage, V3a remains cutoff and V3b conducts heavily. When the negative trigger from V1 is applied to the grid of V3b, the plate current of V3b begins to decrease. This reduced plate current causes a reduction in the bias on V3a; therefore, V3a begins to conduct. The plate current through V3a causes a negative signal to be passed on to the grid of V3b, reinforcing the original trigger; the regenerative action is very rapid, almost instantly causing V3a to reach heavy conduction, and V3b to cut off. The condition of cutoff exists until the bias on V3b is reduced to a point above cutoff. The time duration of this condition is determined by the RC time constant in the grid circuit of V3b (1225 microseconds). At the end of this period, V3b begins to conduct, and the regenerative action of the circuit causes the tubes to revert to their steady-state condition almost immediately. The output of this multivibrator is a positive square wave with a duration of 1225 microseconds. V2, the voltage-regulator tube in the plate supply of the multivibrator, prevents plate-voltage changes from altering the delay time of the multivibrator.

The output square wave of V3 is coupled to V4a through a differentiating circuit. The grid of V4a is returned to +250 volts through a 1-megohm resistor, and the positive bias developed causes the tube plate current to be at saturation. Positive pulses present on the grid of this tube cause only a slight change in plate current, while negative pulses cause a large change. This effect provides a clipping action for positive pulses, but allows negative pulses to pass freely to the plate circuit, where they are inverted to positive pulses. The output of this stage is coupled to V4b, the grid of which is biased to cutoff so that negative pulses applied to this stage are not passed. Positive pulses, however, are passed quite readily, and are inverted to negative pulses in the plate circuit. As can be seen by this description, the only pulses that pass through the clipping system are the differentiated trailing edges of the square-wave output of the first delay multivibrator, V3. These pulses are delayed from the input trigger by 1225 microseconds.

The negative-pulse output of V4b is directly coupled to the second delay multivibrator, V6. The operation of this stage is identical to that of V3, with the exception of the time duration of the output square wave, which is variable from approximately 1225 microseconds to 1325 microseconds. Variable control over the time duration allows precise adjustment of the delay provided by the second delay multivibrator (to approximately 1275 microseconds). This time delay, in addition to the 1225-microsecond time delay of the first delay multivibrator, gives a total delay of 2500 microseconds throughout the unit.

The output square wave of V6 is coupled to V7a through a differentiating circuit. V7a clips and inverts the differentiated waveform in exactly the same

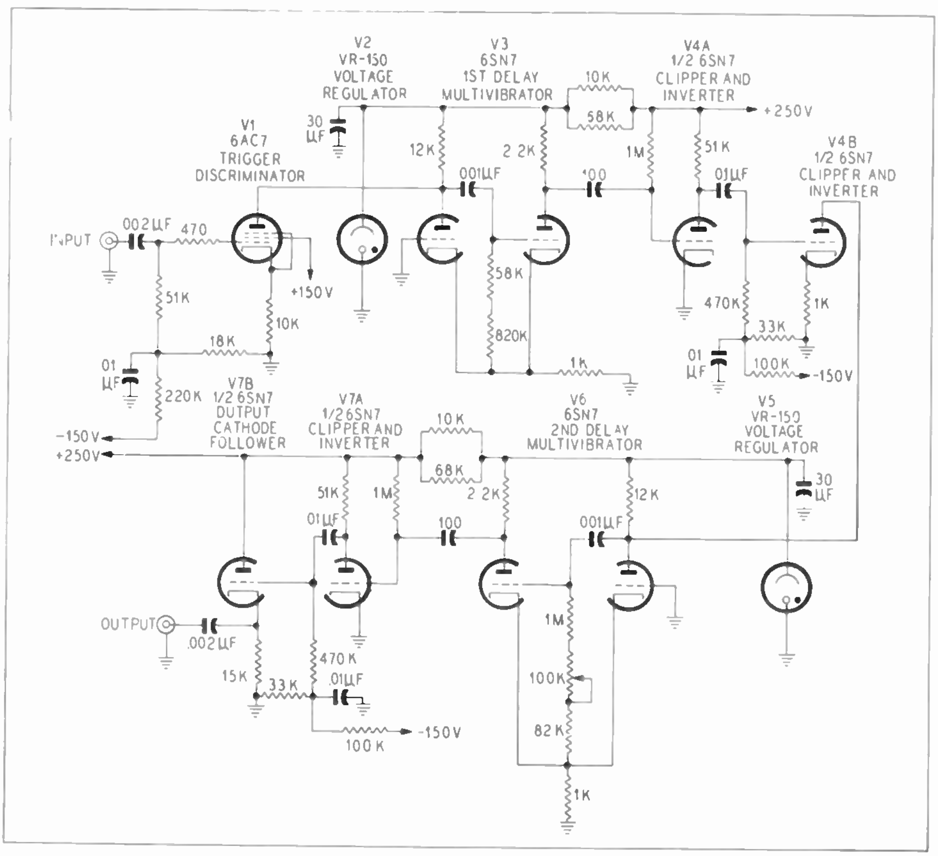

Figure 4. Schematic Diagram of Delayed-Trigger Generator

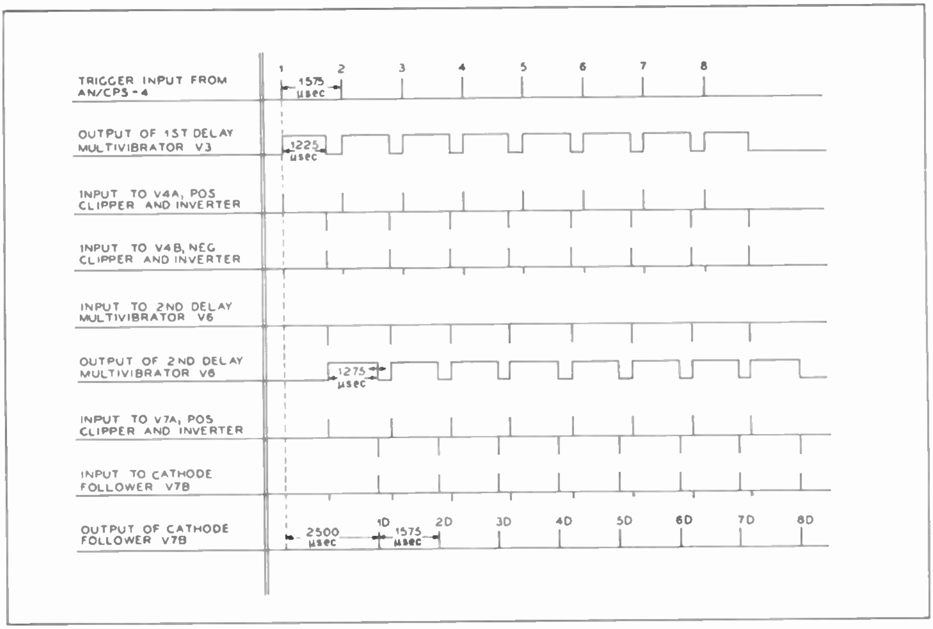

Figure 5. Waveform Analysis of Delayed-Trigger Generator

manner as V4a, and its output is coupled to V7b, the output cathode follower. This stage, which is biased to cutoff, clips the small negative pulses present at the input, but passes the large positive pulses (see figure 5). The positive pulses appear at the cathode of the stage as positive output trigger pulses of short duration, with an amplitude of approximately 20 volts. These output pulses are delayed from the input originating trigger pulses by 2500 microseconds.

Power Supply PP-229/CPS-1 was used as the source of power for the prototype unit; however, any power supply with suitable characteristics may be used.

In Coming Issues

The October issue will contain a feature not previously found in the BULLETIN: a book-review section. Many excellent new technical books are being published by John Wiley & Sons, D. Van Nostrand, McGraw-Hill, and other publishers; we plan to review certain of these books in each future issue.

In addition, the October issue will contain another article by Gail W. Woodward. The new article, titled "S-Band Radar Propagation," is an analysis of the effects of atmospheric conditions on radar propagation, with particular emphasis on the causes and effects of various types of M-curves.

MICROWAVE WAVE GUIDES AND COMPONENTS

By Gail W. Woodward
Technical Publications Department

Theory and characteristics of radio-frequency wave guides and other circuit components used in the transmission and control of microwave energy. Mechanical details and RTMA symbols are shown for all new or unusual components.

(Editor's Note: This article is the author's generalized version of a lecture he originally wrote for one of the new Philco training manuals on microwave equipment. Since the material is of wide interest to BULLETIN readers, we decided to publish it in this form.)

WAVE guides have been developed to meet the need for an efficient microwave transmission system. At microwave frequencies, an open-wire line is useless because of the high losses encountered. A coaxial line is more efficient than an open-wire line, but the need for supporting the center conductor presents a problem. If dielectric materials are used to support the conductor, the losses at microwave frequencies are relatively high, and if stub supports are used, the coaxial line is frequency-sensitive; that is, only a limited range of frequencies can be transmitted without serious losses. Another disadvantage of the coaxial line is that it may arc over if used with a high-powered transmitter.

A wave guide is simply a hollow metal pipe (usually rectangular) that has the property of conducting high-frequency energy with very little loss. The main requirement for wave-guide conduction is that one dimension be equal to or greater than one-half of a free-space wavelength of the energy being conducted. To meet this requirement at low frequencies, a wave guide would be excessively large, but at microwave frequencies it may be relatively small, and, therefore, very practical to use.

Since wave-guide circuits are quite different from those commonly encountered, a comparison of wave-guide technique with standard technique is made in this article. The new RTMA series of symbols for wave-guide components is used throughout.

REFLECTION

A wave guide operates on the principle of surface reflection. Upon entering

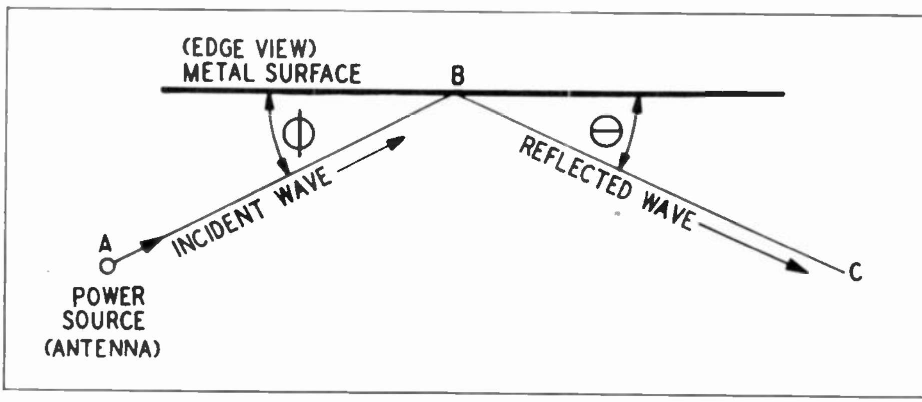

Figure 1. Reflection from a Plane Surface

a wave guide, the energy does not pass straight through, but is reflected from side to side. In the fundamental mode of operation, the reflection occurs between the walls with the greatest separation. To better understand this operation, refer to figure 1. Point A in this figure represents the location of an antenna. Energy radiated by the antenna and reaching point B strikes the metal surface, and is reflected to Point C.

After reflection occurs, there are two signals to consider — the incident (direct) wave and the reflected wave. These two signals combine to form a resultant wave, which in effect, travels parallel to the metal surface. Figure 2 shows the structure of the incident wave in terms of voltage. The lines at right angles to the direction of power flow represent wave fronts, or regions of constant voltage, and, for the purpose of explanation, are assumed to be straight and parallel. The solid lines represent zero voltage points, the minus (—) lines represent negative voltage peaks, and the plus (+) lines represent positive voltage peaks. One wavelength is shown as the distance between two consecutive positive peaks. When these wave fronts strike the metal surface, they are reflected off at the same angle at which they struck. (According to the law of reflection, angle θ is equal to angle (ϕ).)

Since a metal surface acts as a short circuit, the voltage is zero at the surface; this means that each positive wave front must become a negative wave front after reflection. (If the wave front remained positive after reflection, the sum of the incident and reflected voltages at the surface would not agree with the zero-voltage condition that exists there.) The combined waves are shown in figure 3. Notice that, at the reflecting surface, each + is cancelled by a —. Along the line D-D$_1$, each + incident wave front meets a + reflected wave front, and each — incident wave front meets a — reflected wave front. The incident and reflected waves, therefore, combine in phase along this line, and produce a resultant wave

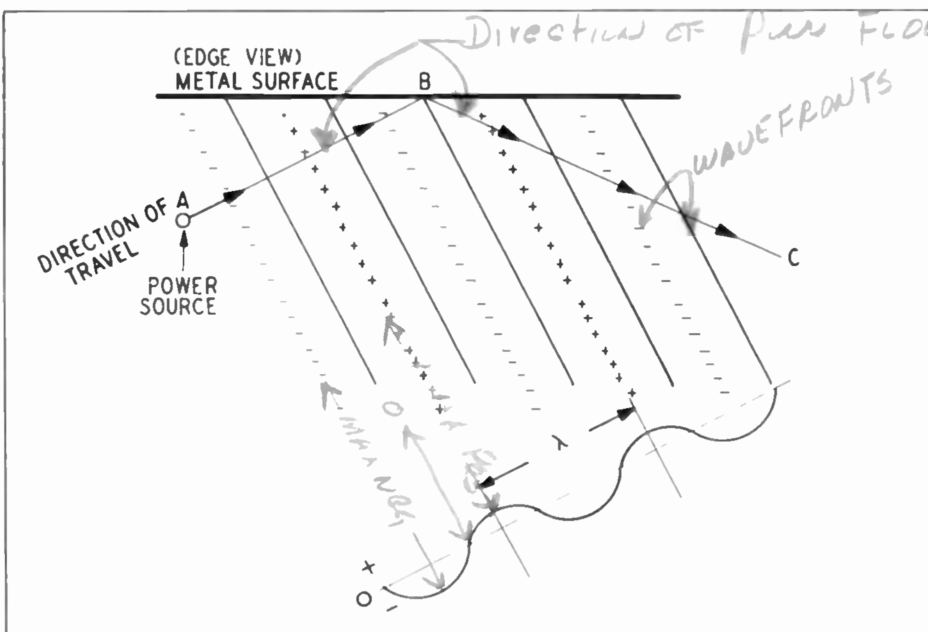

Figure 2. Wave Fronts Incident Upon a Plane Surface

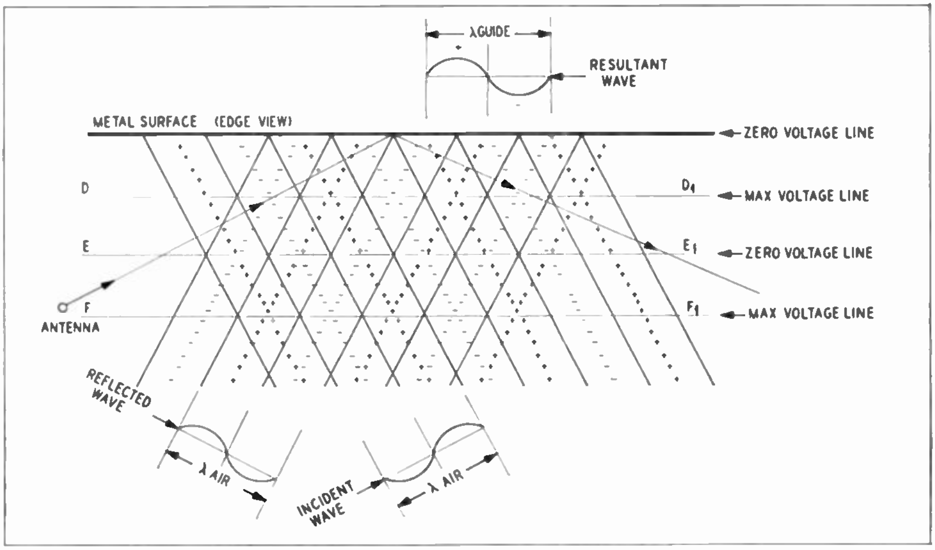

Figure 3. Combination of Incident and Reflected Wave Fronts

that is longer than the applied wave, and which is parallel to the reflecting surface. Along the line E-E_1, each + incident wave front meets a — reflected wave front, and each — incident wave front meets a + reflected wave front; thus a zero-voltage condition exists all along the line. Line F-F_1 shows in-phase conditions like those along line D-D_1, and, therefore, displays identical results. If this process were continued, it would be found that the next would repeat the conditions found in line E-E_1. Thus it is clear that a series of zero and maximum-voltage lines exists parallel to the metal surface.

Since a metal surface represents zero-voltage conditions, it follows that wherever a line of zero voltage occurs (see figure 3), a second metal surface could be placed parallel to the first, with no distortion of the wave-front pattern. In standard wave guides, such as are most often found in SHF equipment, a second metal surface is placed at the line of zero voltage nearest the first metal surface—in this case, line E-E_1. It is apparent that many possible locations are available for this second surface; however, each one results in a different method, or mode, of wave-guide operation. Since a conventional wave guide employs the minimum usable spacing, it operates in the fundamental mode, which is the only mode considered in the following discussion. It is interesting to note that the fundamental mode is the most efficient of all the modes encountered in wave-guide operation.

When a second metal surface is used parallel to the first, energy may be propagated, along a path parallel to the surfaces, by a process of reflection from one surface to the other. Since the energy is confined between the two metal surfaces, very little loss results. To complete the wave guide, two more metal surfaces are placed at right angles to the existing ones, to form a rectangular metal pipe (see figure 4). These two metal surfaces are spaced too close together to allow them to function in the same manner as the first pair. Examination of figure 3 shows that the zero-voltage line (E-E_1) can never approach closer than one-half wavelength to the metal surface.

Figure 4 shows how energy passes through a wave guide by being reflected back and forth between the wide-spaced

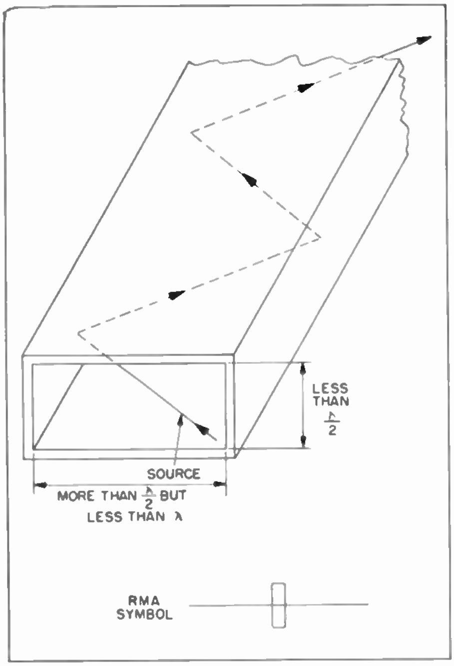

Figure 4. Propagation Through a Wave Guide

walls. For operation in the fundamental mode, a wave guide must be proportioned as follows:

1. The narrow dimension must be less than one-half wavelength, to prevent propagation from occurring between the close-spaced walls.

2. The wide dimension must be greater than one-half wavelength, but less than one wavelength, to insure operation in the fundamental mode.

The inside dimensions determine the frequency range of a wave guide for operation in the fundamental mode. In RG-52/U wave guide, the inside dimensions of the wave guide are .9″ by .4″, and the frequency range is from 6600 mc. to about 13,200 mc. (for the fundamental mode). The cutoff frequency for the wave guide is 6600 mc.; frequencies below this point are attenuated to a very great degree. If frequencies above 13,200 mc. are applied to the wave guide, the energy will be transmitted, but the wave guide may operate in modes other than the fundamental. Such a condition would result in losses or unwanted resonance conditions. To maintain operation within these limits, an operating range of from 8100 mc. to 12,500 mc. is recommended. Brass RG-52/U wave guide has a loss factor of less than .1 db per foot (and even less when silver-plated), with a power-carrying capacity of about one megawatt. These figures alone indicate the superiority of wave guide over conventional transmission lines.

Wave-guide operation is further clarified by figure 5, which represents a view of the wide dimension of a wave guide operating at three different frequencies. The heavy lines show the direction of travel, and the light lines indicate wave fronts. The distance between wave fronts is one-half of a free-space wavelength; the distance between points of intersection of wave fronts within the wave guide is defined as a wave-guide half-wavelength (denoted $\frac{\lambda \text{ guide}}{2}$). At frequencies near cut-off (A of figure 5), the energy travels many times back and forth across the wave guide, resulting in a very long wave-guide wavelength. This long wavelength might lead one to believe that the energy is being propagated through the wave guide at a very high velocity (several times free-space velocity), but this is only an apparent condition called *phase* velocity. Since the energy must be reflected from one wall to the other a great many times to pass through the wave guide, the actual velocity (called *group* velocity) is much slower than the velocity in free space. Parts B and C of figure 5 show the action within a wave guide at progressively higher frequencies. It can be seen that as the frequency increases, wave-guide wavelength more nearly approaches free-space wavelength, while fewer reflections are required for propagation; this also results in increased group velocity.

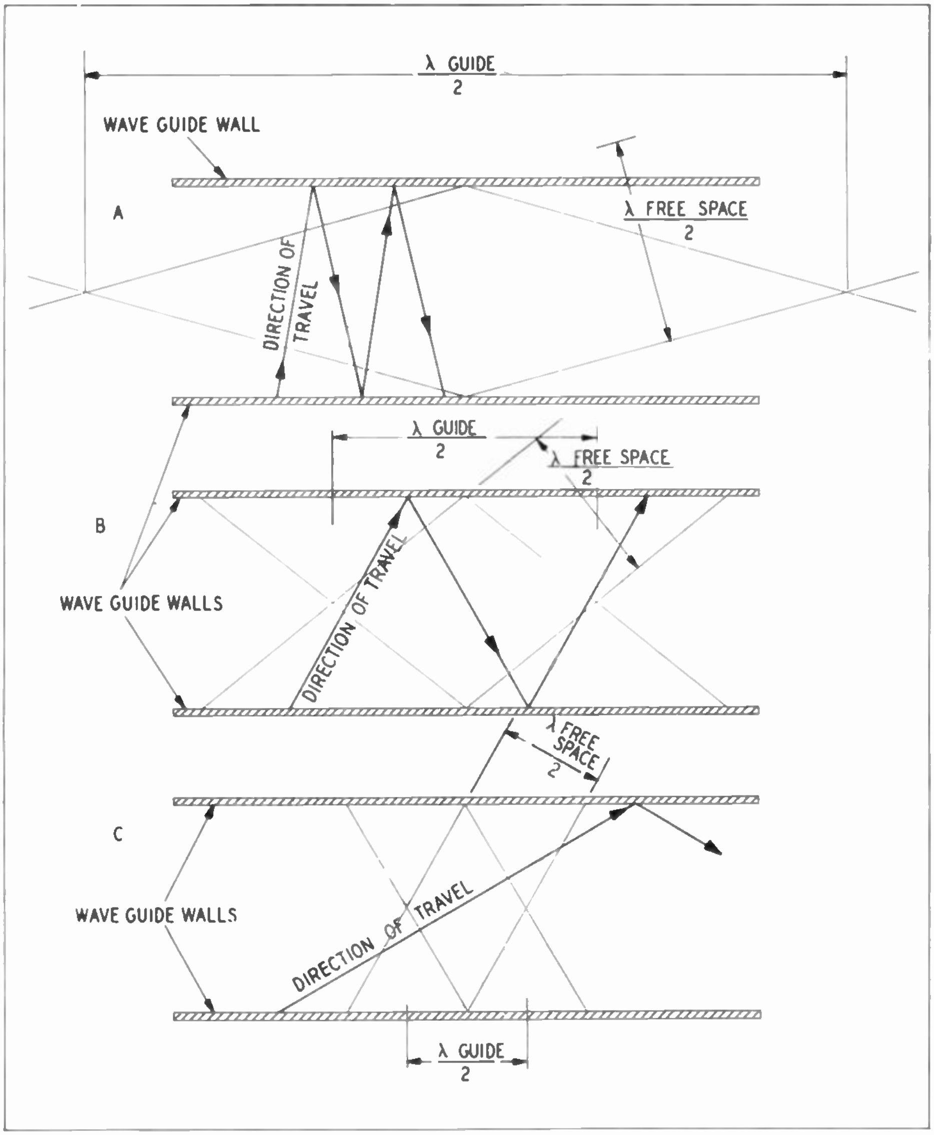

Figure 5. Propagation at Three Frequencies Through a Wave Guide

In the propagation of electromagnetic energy, two fields are always in existence—the electrostatic field and the electromagnetic field. These two fields are at right angles to each other, so that if the electrostatic field appears across the wave guide, the magnetic field must extend down the wave guide. For simplicity of presentation, only the electrostatic field is considered in this discussion. In a wave guide operating in the fundamental mode, the electrostatic field exists across the narrow dimension, and is said to be *transverse electric* (TE). The transverse electrostatic field pattern is described symbolically by using two subscripts with the symbol TE, such as $TE_{0,1}$, where the first subscript indicates the number of half-wave patterns across the narrow dimension of the wave guide, and the second subscript indicates the number of half-wave patterns across the wide dimension (see figure 6). Thus $TE_{0,1}$ describes the fundamental mode, and $TE_{0,2}$ describes the next higher mode, which occurs in a wave guide twice as wide as the one for the fundamental mode, or is a wave guide the same size as for the fundamental mode, but operated at a higher frequency. It is possible that a wave guide which is large enough to operate in the $TE_{0,2}$ mode may also be operated in the $TE_{0,1}$ mode, but the mode of operation may change during propagation. Figure 6 shows the distribution of electrostatic lines of force for the TE type of propagation. The length of each line represents the intensity of the field at that point. These lines result from the combination of incident and reflected wave fronts, as indicated in figure 3. Note that operation in the fundamental mode results in maximum intensity in the exact center of the wave guide, with zero intensity at the surface of the wide-spaced walls.

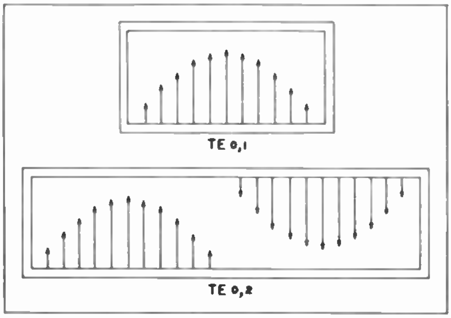

Figure 6. Electrostatic Field in a TE Wave Guide

IMPEDANCE

In a discussion of transmission lines, an important consideration is characteristic impedance. A wave guide has impedance, but the impedance is not easy to determine. Wave-guide impedance depends upon: (1) the mode of operation, (2) the width of the wave guide, (3) the height of the wave guide, and (4) the frequency of the signal. Thus, for a given mode of operation, any change in the height or width of the wave guide, or in the signal frequency, changes the impedance of the wave guide. For example, lowering the frequency or increasing the narrow dimension of a wave guide operating in the $TE_{0,1}$ mode raises the impedance, while increasing either the wide dimension or the signal frequency lowers the impedance. (At the cutoff frequency, wave-guide impedance is considered infinite.) Since so many factors influence the impedance, this property of wave guides is seldom discussed as such. Also, since it is very difficult to consider microwave circuit elements in terms of inductance, resistance, and capacitance, because of the extremely small values involved, it is much more accurate to discuss wave guides in terms of *matching* impedance. Instead of saying that a 200-ohm wave guide is feeding a 200-ohm load, it is better to say that the wave guide is feeding a matched load for a given frequency or range of frequencies.

ACTION OF DIELECTRIC

It is sometimes desirable to place a movable dielectric material inside a

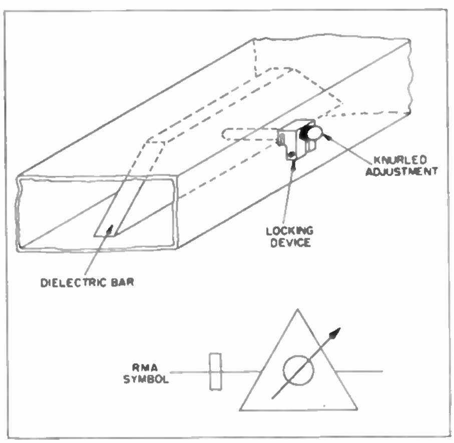

Figure 7. Dielectric Bar (Phaser) in a Wave Guide

wave guide. A dielectric material has the effect of concentrating electrostatic lines of force by providing a better path for them. The material used for this purpose must be a good insulator (as are all good dielectrics), because any conduction would cause resistive power losses and inductive effects. In the case of a wave guide, the dielectric has the effect of making the wave guide look longer electrically than physically. The dielectric element, which may either completely or partially fill the wave guide, is mainly used in impedance-matching or in varying the phase of the r-f energy being transmitted through the section. The dielectric element may take many forms. Figure 7 shows a bar-type adjustable dielectric section; this section can be moved laterally across the wave guide, from the center to one side, so that the electrical length of the wave guide can be changed. The end of the dielectric section is tapered, to prevent a *sudden* discontinuity in the wave guide. Such a discontinuity would cause an abrupt change in the wave-guide impedance, and would produce undesirable reflections. The dielectric has maximum effect in the center of the wave guide, because of the high intensity of the electric field in this region, and has minimum effect at the side of the wave guide. The bar-type dielectric shown in figure 7 is usually called a phaser section, and is adjusted by means of a knurled knob, which is locked in position after adjustment.

MATCHING

Wave-guide impedance-matching can be accomplished by several different methods, one of which makes use of a quarter-wave resonant section of line, as shown in figure 8. The impedance of the resonant section is adjusted to a value between the two impedance values to be matched. The action of the resonant section, which is similar to that of a "Q-bar" section used in open-wire lines, is as follows: Assume that a signal is traveling through the low-impedance section toward the matching section. When the signal reaches point A, a certain percentage of the signal is reflected back toward the signal source, and the remainder proceeds toward the high-impedance section. Upon arrival of the signal at point B, another reflection occurs, and, if the matching section is correctly adjusted, the amount of the energy reflected is equal to that at A. Since the matching section is a quarter-wave long, the signal reflected from B, upon reaching A, will be 180° out of phase with the signal reflected from A. As a result, the two reflected signals cancel. Therefore, the quarter-wave sec-

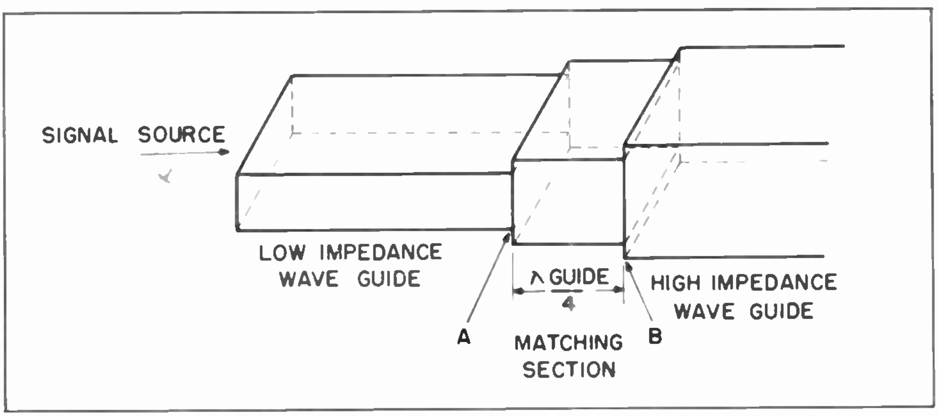

Figure 8. Quarter-Wave Matching Section

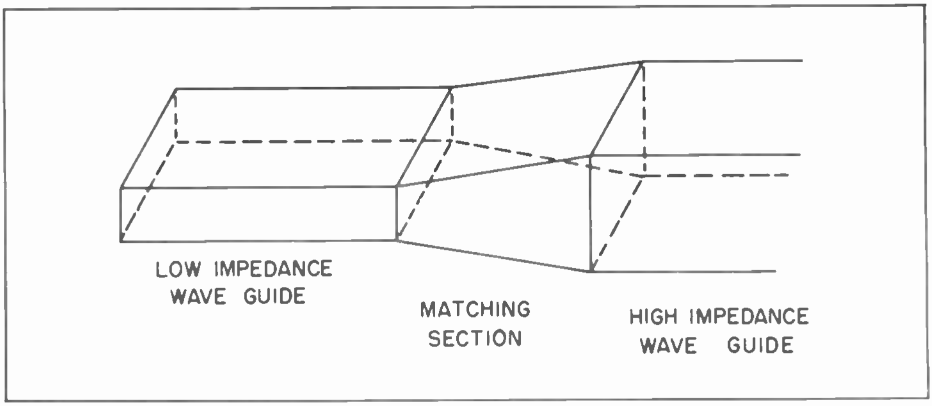

Figure 9. Taper Section

tion matches the impedance between the two different sections.

It can be seen that the action just described is effective at only that frequency for which the matching section is a quarter-wave element. Therefore, the matching section is said to be frequency-sensitive.

A second method of matching is shown in figure 9, where a *tapered* section is used to join two wave guides of different impedance. As a signal passes through the tapered section, it encounters very little discontinuity if the taper is gradual. For this reason, if the matching section is long compared to a wavelength, very little reflection occurs at any one point, and a reasonably well-matched condition exists. The tapered section has the advantage of being a broad-band device, since its operation does not depend upon frequency.

TERMINATION

It is often necessary to terminate a transmission line so that no reflection is produced. This is done by connecting to the line a resistive load which has a resistance equal to the line impedance. A proper termination, or load, must satisfy two requirements: it must match the impedance of the transmission line, and it must dissipate the entire power output of the transmission line.

Figure 10 shows a commonly used type of wave-guide load. A tapered strip of resistive material is placed in the center of the wave guide, at the point of maximum voltage. The resistive material can be any one of several poor conductors; carbon-impregnated linen bakelite is commonly used. Since the resistive element is across the maximum voltage points of the wave guide, a current flow is set up in the resistive material, and heat is generated. If the load matches the wave guide, all of the incident power is absorbed, and is dissipated in the form of heat. A load of the type shown in figure 10 will dissipate only a few watts of r-f power. In one common type of high-power load, the end of the wave guide is filled with resistive material; the load is cooled by the use of cooling fins or circulating water.

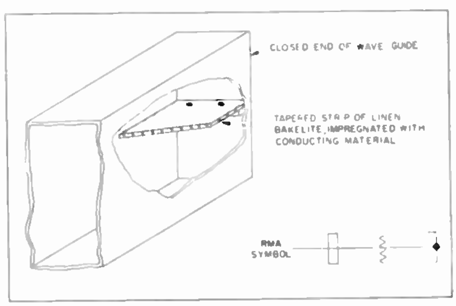

Figure 10. Wave-Guide Load

The action of the tapered end of the resistive element in preventing reflections, is similar to that of the dielectric bar. By the time the signal reaches the closed (shorted) end of the wave guide, its amplitude is reduced to such a value that any reflection that occurs will cause no serious standing-wave condition.

VARIABLE ATTENUATOR

The principle of the wave-guide load is used in the construction of a wave-guide variable attenuator. The variable attenuator is used, during measurements, to adjust the power output from a wave guide. Figure 11 shows the construction of a typical wave-guide attenuator. In this attenuator, the resistive section may be moved to any position from the edge of the wave guide to the center. When it is at the center, the output may be reduced by as much as 30 db (1000:1), and, when it is at the edge, the effect of the attenuator can be neglected, since the voltage is zero at that point. The variable attenuator just discussed is classed as the dissipative type, because it removes power from the wave guide in the form of heat, instead of merely rejecting the signal. As in the dielectric bar, the taper principle is used to prevent power reflection from the surface of the element. However, in this attenuator, both ends of the element are tapered, because the attenuator must work equally well in two directions. A second type of wave-guide attenuator, classed as non-dissipative (power flow is rejected), is in the form of a section of cutoff wave guide (too small for $TE_{0,1}$ operation) whose length can be varied. This type of attenuator is commonly found in earlier equipments, and can easily achieve a range of 100 db of attenuation.

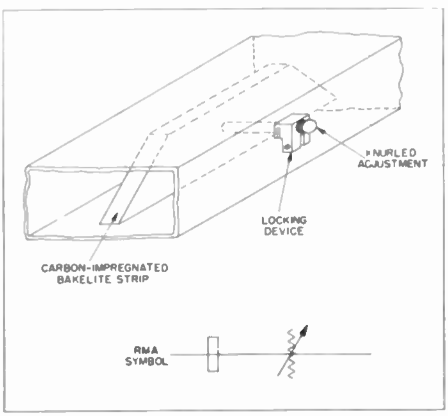

Figure 11. Wave-Guide Attenuator

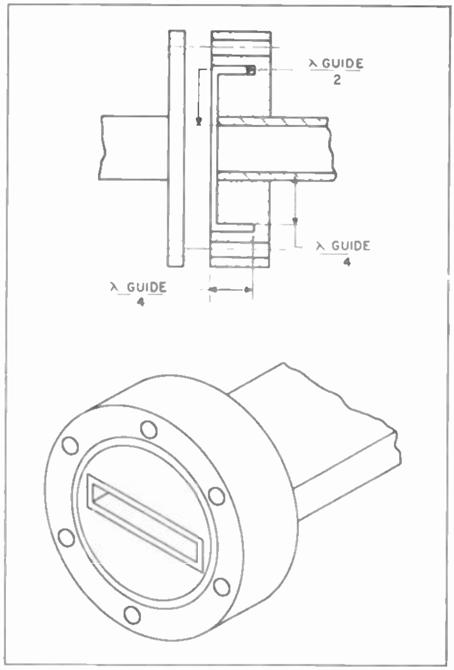

Figure 12. Wave-Guide Choke Joint

JOINTS

The rigidity of a wave-guide system makes it difficult to connect adjacent sections. For the best connection, the ends of the sections to be joined should be cut perfectly square, smoothed and polished carefully, aligned properly, and then clamped together under high pressure. This method of connection is, of course, applicable only in laboratory work. Choke joints that are very simple to use have been developed from practical work. This type of joint consists of a choke section facing a plain flange.

(See figure 12.) For the choke section, a circular groove, one-quarter wavelength deep and one-quarter wavelength away from the points of maximum voltage inside the wave guide, is milled around the end of the wave guide. The bottom of the groove forms a short circuit, which is reflected one-half wavelength away (at the junction of the two wave-guide sections) as a short circuit. This reflected short circuit electrically joins the two sections together. The flange serves only as a path for the reflection currents that produce the short circuit. The choke joint may be simply bolted together, because a slight separation between the choke and flange, or a slight off-setting of the two sections, does not cause appreciable reflection. If a choke section is joined directly to another choke section—instead of a flange—the two quarter-wave slots may form a half-wave resonant section, and cause abnormally large reflections at certain frequencies. Therefore, when two choke joints must be joined together, the flange action must be supplied by means of a metal adapter plate inserted between the choke sections; the adapter plate thus acts as a flange to both choke sections.

In cases where it is desired to pressurize a wave-guide system (to exclude moisture and foreign material), a second shallow slot is milled just inside the outer rim of the choke section. In this slot is placed a neoprene gasket, which is compressed between the choke and flange sections when they are bolted together. This gasket, which does not affect the electrical functioning of the joint, insures a good pressure seal.

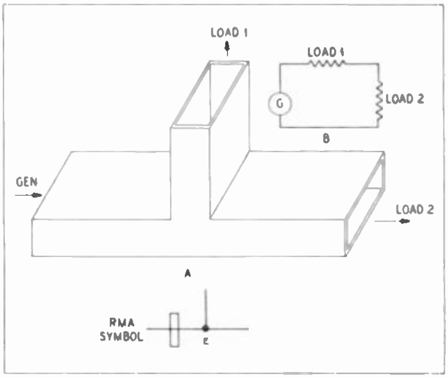

Figure 13. E-Type T Junction

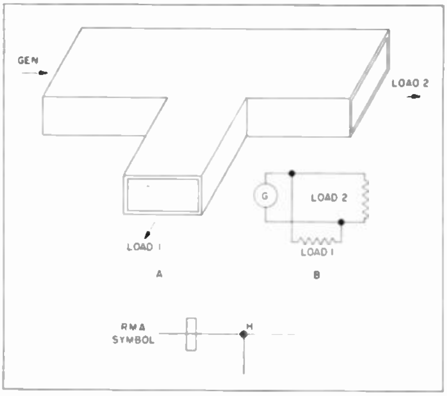

Figure 14. H-Type T Junction

FLEXIBLE WAVE GUIDE

A flexible wave-guide section consisting of a series of choke and flange sections held in place by means of a rubber sleeve has been developed for special applications. The rubber sleeve permits bending, and the choke joints maintain electrical contact. Flexible sections are useful for connecting test equipment, or r-f units that are independently shock-mounted. In the latter case, since the units vibrate independently, a rigid coupling device would be subjected to a flexing motion, and would soon break. Another type of flexible section is in the form of a spiral ribbon with interlocking edges somewhat similar to the armor used in BX cable. The ribbon is formed to wave-guide dimensions, and surrounded with a neoprene or rubber sheath. This type has a somewhat higher loss than that previously described, and is not as easily flexed, but it is less bulky.

Where wave guides are to be permanently joined together, a T junction

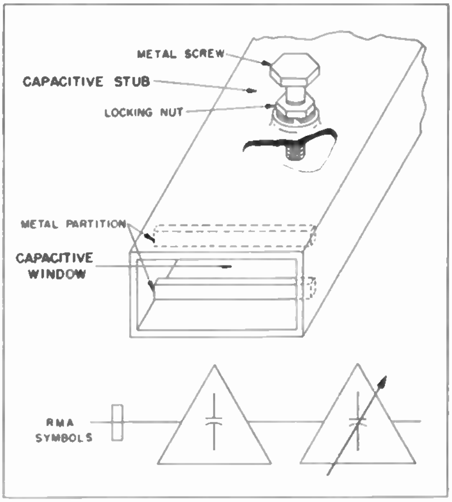

Figure 15. Capacitive Window and Capacitive Stub

is used. There are two types of T junctions—the E-type and the H-type. An E junction, shown in figure 13, is similar to two series-connected transmission lines, and the *electric field (E-lines)* is divided between the branches. An H junction, shown in figure 14, is similar to two parallel-connected transmission lines, and the *magnetic field (H-lines)* is divided between the branches. The resulting discontinuity produced by either type of T junction will cause reflection and standing waves unless compensation is provided. Compensation is usually provided by placing a reactive element at a suitable distance from the junction, so that it reflects power in the proper phase to cancel the power reflected directly from the junction.

REACTIVE ELEMENTS

Reactive elements are inserted into a wave-guide system for matching purposes. In cases where a circuit element causes power reflection, and thus sets up standing waves, it is standard practice to use a reactive element to develop an equal power reflection of the proper phase to produce cancellation. The reactive element may be fixed-tuned in cases where a limited range of frequencies is encountered. However, in equipment where a very wide range of frequencies is possible, the reactive element must be made adjustable.

In general, two types of reactive elements are used—the capacitive type and the inductive type. The capacitive type, two forms of which are shown in figure 15, causes distortion of the electrostatic field within the wave guide. The metal partition is called an iris, or window, and the screw is called a stub. The stub and window have the same effect, but while the stub is easily adjusted, the window is more suitable for fixed operation. An *inductive* iris, or window, is shown in figure 16; in this case the metal partition distorts the magnetic field. If desired, the inductive window and capacitive window may be combined together to form a *resonant* slot, or window. A resonant slot is often used as a filter element, and can be used as an antenna. Figure 17 shows an inductive stub, or, as it is commonly called, an inductive post. This element differs from the capacitive stub in that it joins together the two close-spaced wave-guide walls, and thus provides a current path. Different values of inductance are obtained by using posts of different diameter.

Where an extremely wide range of

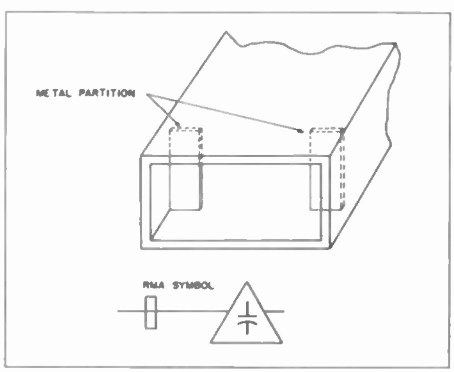

Figure 16. Inductive Window

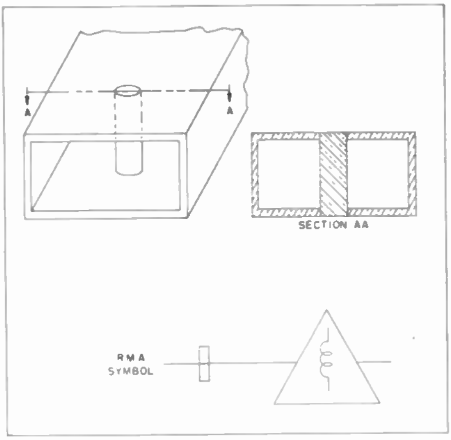

Figure 17. Inductive Post

matching is desired. reactive elements can be used in pairs, or even in groups of three. Two capacitive screws, located an odd multiple of quarter-wavelengths apart. make up a standard matching system. Assume that one of the screws is located at point A. and the other at point B. The one at B is capacitive, but its effect as seen at A (one quarter-wavelength away) appears inductive. Thus at point A, any reactance (from capacitive, through resonance. to inductive) can be obtained, making a wide range of matching available.

TRANSITION (WAVE GUIDE TO COAXIAL LINE)

Connection between a wave guide and a coaxial line is accomplished by means of a transition. The most common type of transition is shown in figure 18, where a coaxial line is terminated in a quarter-wave antenna (or probe), and the antenna is placed across the electrostatic field of the wave guide. With this type of coupling, energy from the antenna is radiated into the wave guide very efficiently. The antenna is placed a quarter-wavelength from the closed end of the wave guide, so that the short circuit produced by the closed end appears at the antenna as a high impedance; consequently, no energy is directed toward that end. This transition works equally well when used to transfer energy from the wave guide to the coaxial line, and may appear in modified form as shown in figure 19. This figure shows a broadband transition; the broadband characteristics are derived from the fact that the increased capacity of the antenna reduces the Q of the antenna circuit. Figure 20 is similar to figure 18 except that the antenna is supported by a quarter-wave stub, which not only serves as a mechanical support, but also provides a d-c return for the coaxial-line center conductor.

Another type of wave guide to coaxial line transition makes use of an inductive coupling loop (see figure 21). In this case the loop is placed into the wave guide so as to couple the magnetic field, which extends down the wave guide in the fundamental mode. The orientation of the loop affects the degree of coupling; therefore. rotation of the loop provides a convenient method of adjustment. The coupling can also be adjusted by controlling the depth of penetration of the loop into the wave guide.

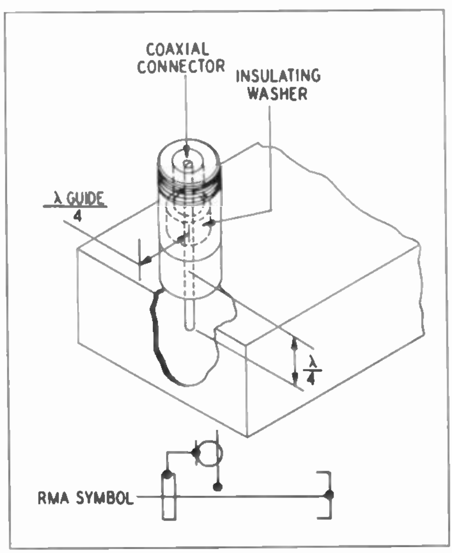

Figure 18. Wave Guide to Coaxial Line Transition

APERTURE COUPLING

There are many ways of extracting power from or inserting power into a wave guide. A simple method consists of drilling a hole into the wave-guide wall at a suitable point (see figure 22). Power transfer may be considered to occur in one of two ways — the opening either breaks a current path, or distorts the electrostatic field in an unsymmetrical manner. In figure 22, hole A will not radiate, while holes B and C will radiate. Part A of figure 23 shows how the hole affects the electrostatic field. Notice that the field intensity is the same at both sides of the hole, so that no voltage exists across the hole; therefore, the hole does not radiate energy. However, if the hole is located off-center, as shown in part B of figure 23, the electrostatic field about the hole is unsymmetrical, a voltage is developed across the hole, and, therefore, radiation results. It should be noted here that if the hole were located on the opposite side of the wave guide, a similar action would result, but the coupling would take place in the opposite phase. Part C of figure 23 shows the hole at a point of maximum current flow. The current flow

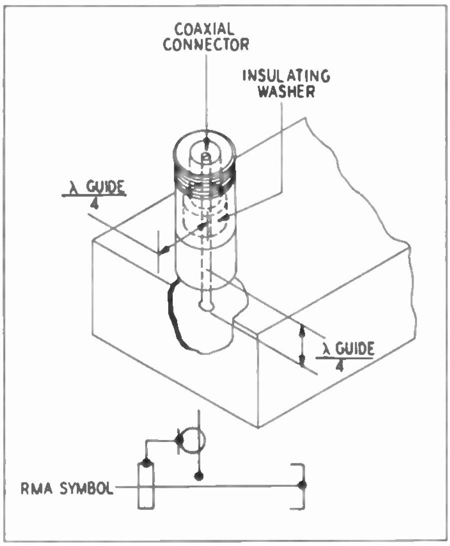

Figure 19. Wave Guide to Coaxial Line Transition (Broad-banded)

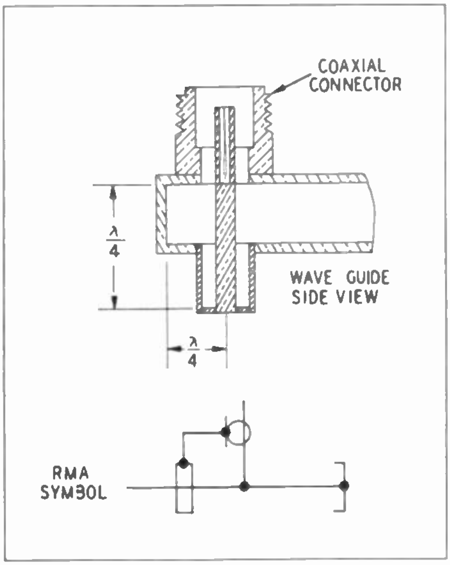

Figure 20. Wave Guide to Coaxial Line Transition (Shorted)

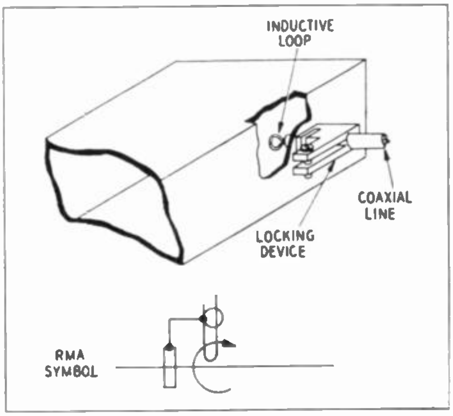

Figure 21. Wave Guide to Coaxial Line Transition (Loop)

on the inner wall surface will produce a voltage drop across the hole, and thus cause radiation. The larger the hole, the greater the amount of radiation; therefore, the degree of coupling can be changed by changing the size of the hole. Although the coupling action has been discussed only in terms of removing power from a wave guide, the action is the same for power entry.

The hole used for coupling can be in the form of a slot without causing any

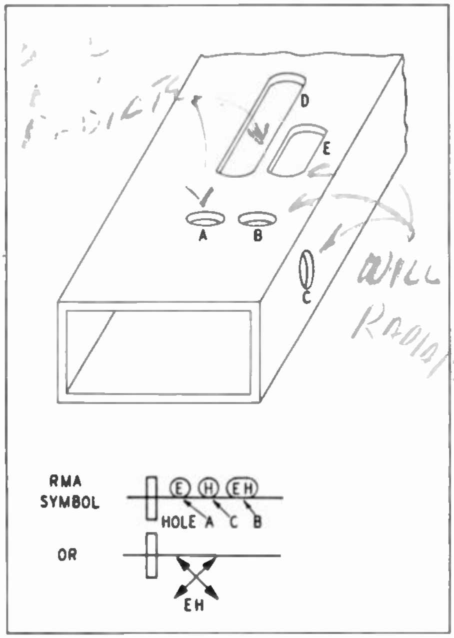

Figure 22. Wave-Guide Apertures

change in operation (see figure 22). Slot D will not radiate, because it is located in the same relative position as hole A. Slot E, however, will radiate in the same manner as hole B. A section of wave guide containing a slot is called a slotted section; such sections are used to facilitate the insertion of measurement devices into the interior of the wave guide. An r-f voltmeter probe can be inserted into the slot and moved back and forth, to determine the voltage-standing-wave ratio of the system during operation.

MICROWAVE ANTENNA ASSEMBLY

A typical microwave antenna consists of a circular parabolic reflector with the antenna feed at the focal point. For reception, the reflector serves to gather energy out of space, and to focus the energy on the feed. The over-all ability to pick up energy is called antenna gain, which is given by the formula:

Gain $= \frac{4\pi AF}{2\lambda}$ where A is the area of the reflector, λ is the wave length in the same unit used to measure the area, and F is a factor used to give the efficiency of focus (usually about .6 or 60%). It can be seen that for any given frequency, the gain is dependent upon the area of the reflector.

Figure 24 shows the construction of the antenna feed. Radiation is accomplished by means of two resonant windows formed by the capacitive-iris and inductive-iris actions previously de-

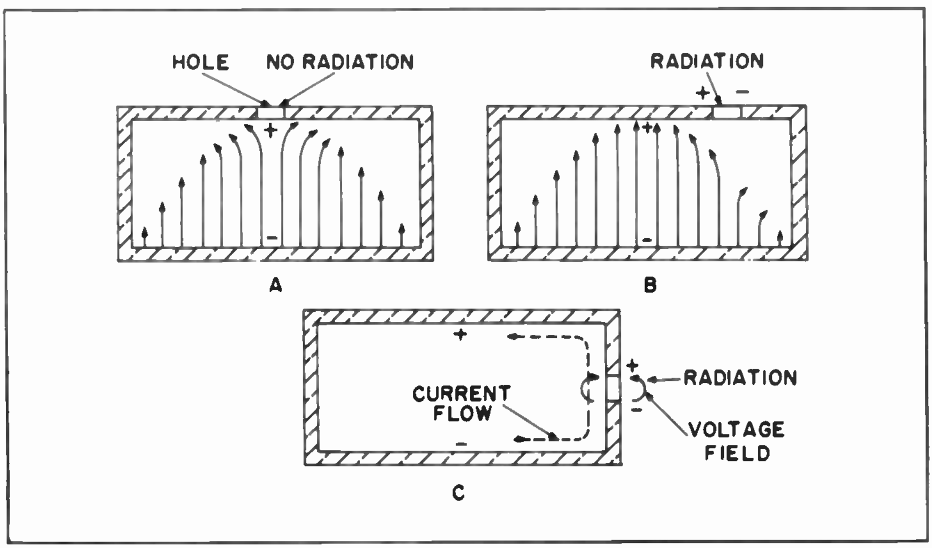

Figure 23. Action of Wave-Guide Apertures

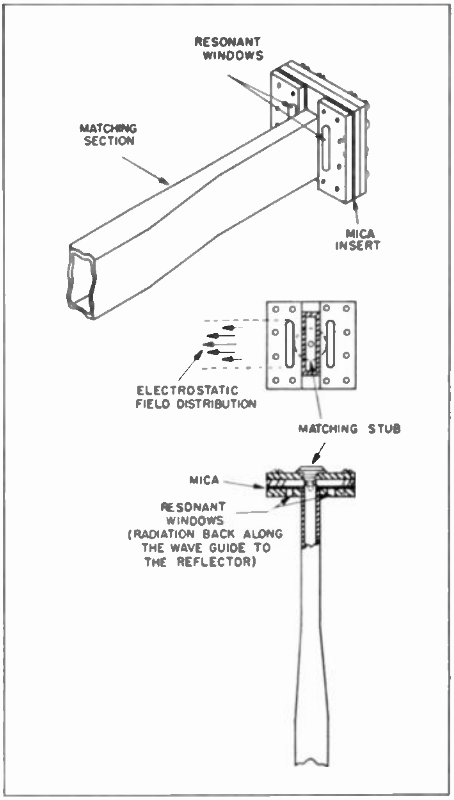

Figure 24. Construction of Antenna Feed Assembly

scribed. This type of window is broadly resonant in much the same manner as a large-diameter dipole, and the feed is matched by means of a factory-adjusted capacitive stub. This type of feed is essentially a low-impedance device, and, since the wave guide used for transmission operates at a higher impedance. a matching section must be used. The matching section, which is the tapered-wave-guide type shown in figure 9, is used because of its broadband characteristics.

MICROWAVE RECEIVER R-F SECTION

A typical wave-guide r-f section for a microwave receiver is shown in figure 25. It is the function of this circuit to pass the desired microwave signals (within a given band) to a first detector, where a local-oscillator signal is injected, and an i-f signal is produced. The i-f signal is then amplified in a conventional manner. The r-f section shown in figure 25 consists of a preselection filter, a local-oscillator injection jack, a crystal-type first detector, and two matching stubs.

R-F FILTER

Figure 25 shows the construction of the r-f filter. There are four tuned sections, each of which consists of two inductive posts and a capacitive stub. In each case, the posts are located one-half of a wave-guide wavelength apart, at the *highest* frequency of operation. The effect of this half-wave section is to form

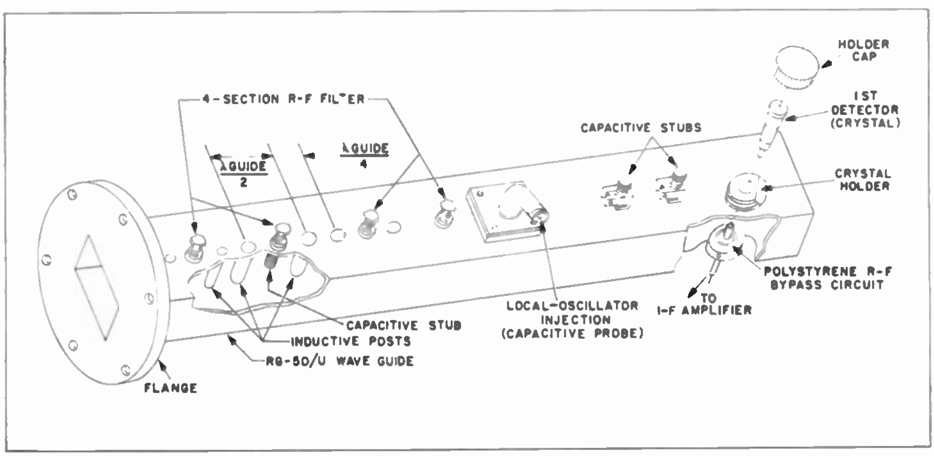

Figure 25. Microwave Receiver R-F Section

a resonant circuit as shown in part A of figure 26. The posts represent inductive elements, and the wave-guide walls form the capacitive element. When the capacitive stub is inserted into the wave guide the capacitance is effectively increased, and the resonant frequency is lowered to the desired value. If the two posts are considered to appear in parallel to each other, and in parallel with the capacitive elements, the resultant equivalent circuit will appear as in part B of figure 26. This circuit will look like a short circuit to all frequencies except at resonance, where it will have very little effect (high impedance). Since the r-f section uses four such filter sections, the entire circuit appears as shown in part C of figure 26. It will be noted that each section is located one-quarter of a wave-guide wavelength away from the next. This spacing causes the impedance to be inverted over each quarter-wave of travel, thus causing two of the sections to appear as series-rejection circuits, while the remaining two appear as shunt-acceptor circuits. (See part D of figure 26.) The desired response curve is obtained by tuning all elements to the same frequency, and is determined by the Q of the individual sections. As seen in figure 25, the two end sections have smaller posts than the center sections, a condition which results in low-Q end sections. (The smaller posts have a greater inductance value than the larger posts, and, in addition, the reduced surface area of the small posts results in increased surface resistance.) The combined bandpass of the four sections is shown in part E of figure 26. Individually, each section has a conventional-tuned-circuit response curve, with the two center sections exhibiting a bandpass of about 23 mc., and the two end sections about 50 mc. The curve in part E of figure 26 is called *maximally flat*, and is about 20 mc. wide.

LOCAL-OSCILLATOR INJECTION

The local-oscillator signal is supplied by means of a coaxial cable, which terminates in a right-angle fitting mounted upon the wave-guide section just following the r-f filter (see figure 25). The local-oscillator fitting terminates in a capacitive probe, which extends into the wave guide, thus forming a transition of the type shown in figure 18. This transition is located at the proper distance from the last filter section so that no signal is radiated toward the filter, but is directed toward the crystal detector.

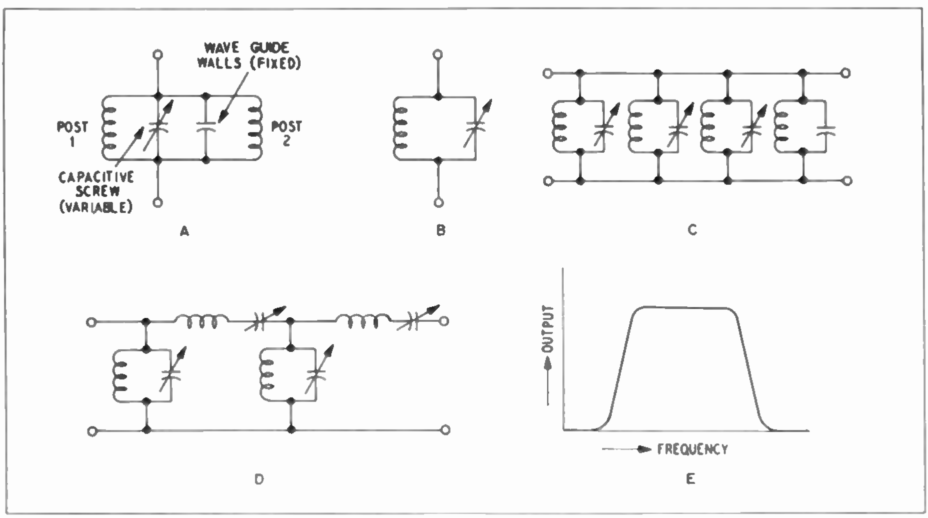

Figure 26. Development of Circuit Equivalents of R-F Filter Section

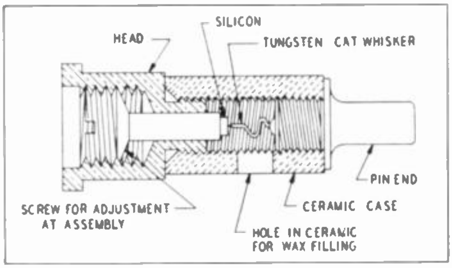

Figure 27. Microwave Crystal Unit

CRYSTAL DETECTOR

The first detector, a crystal of the 1N23 series, is placed across the voltage points of the wave guide in such a manner that it acts as a matched impedance to the r-f signals being transferred down the wave guide. The wave-guide impedance is too high at the center of the guide for an exact match, so the crystal is mounted a little off-center, and approximately one-quarter of a wave-guide wavelength from the closed end. The match exists only for the received signal; no attempt is made to match the local-oscillator signal, since it is sufficiently strong to overcome any losses due to mismatch.

The construction of the crystal is shown in figure 27; while the internal structure varies from one manufacturer to the next, the external dimensions and the electrical characteristics have been standardized. The active elements are the cat whisker and the silicon wafer, which serve as rectifying elements. Since a rectifier is electrically a non-linear impedance, heterodyning action takes place, and the difference between the r-f and local-oscillator signals appears as the i-f signal. Figure 25 shows the construction of the crystal mount, which incorporates an r-f by-pass circuit. The r-f by-pass circuit consists of two brass bushings separated by a cylinder of polystyrene; the combination forms a capacitor which is electrically one-quarter wavelength long at the r-f signal frequency. This structure causes the crystal holder to appear as a broadly tuned series-resonant filter circuit, which effectively removes the r-f signal and leaves the i-f signal. The i-f signal is then coupled into the first i-f amplifier.

The use of a crystal at microwave frequencies is desirable because of the relatively high noise level of vacuum tubes. The signal-to-noise ratio of a crystal is far superior to the best vacuum tube so far developed. TABLE I gives a comparison between the characteristics of several modern silicon crystals. The 1N21-series crystals are designed for operation below 3000 mc., but they can be used at higher frequencies if a reduction of efficiency can be tolerated for emergency operation.

MATCHING STUBS

Two capacitive stubs used for matching the crystal to the wave guide are shown in figure 25. Some form of adjustable matching is necessary because of the wide range of frequency operation. The stubs are adjusted so that minimum power is reflected from the crystal at the operating frequency.

TABLE I

TYPE	OPERATING FREQUENCY (MC.)	CONVERSION LOSS (db)	RELATIVE NOISE OUTPUT	RESISTANCE TO BURNOUT
1N21	3000	8.5	4.0	Poor
1N21A	3000	7.5	3.0	Poor
1N21B	3000	6.5	2.0	Good
1N21C	3000	5.5	1.5	Good
1N23	10000	10.0	3.0	Poor
1N23A	10000	8.0	2.7	Medium
1N23B	10000	6.5	2.7	Poor

Characteristics of GERMANIUM DIODES

By Bernard F. Osbahr
Associate Editor TELE-TECH Magazine

An analysis of the technical specifications for 36 different commercially available types of general-purpose diodes, matched duo-diodes, and varistors.

(Editor's Note: This article was originally published in the December, 1950, issue of TELE-TECH magazine, and is reprinted here through the courtesy of Caldwell-Clements, Incorporated, publishers of TELE-TECH.)

THE USE OF germanium diodes as substitutes for vacuum tubes in communications circuits has increased markedly year by year since World War II. This year, for example, nearly four million units will have been sold for initial equipment and replacement purposes, and with new circuit applications being uncovered daily, next year's production may be more than double this figure. Last year, only two TV manufacturers indicated that their receivers were using germanium diodes as video detectors; this year this number has been stepped up to twenty. In addition to this use, many units are finding their way into radios (particularly FM), test and measuring equipment, commercial and Government communications equipment, and research equipment.

Technical specifications for all currently available types of germanium crystal diodes are shown in Table I. The 36 types are listed by their RTMA type number as distinguished from a manufacturer's type number, since one manufacturer may make types whose technical specifications were registered by another. The chart also indicates, generally, the application purpose intended for each type.

General-purpose diodes are designed to operate as rectifiers, clippers, d-c restorers, and modulators, at frequencies up to several hundred megacycles. Germanium crystal diodes are also obtainable as matched units in the form of duo-diodes, varistors, or quads (1N35, 1N40, 1N41, 1N42, 1N71, 1N73, 1N74). In the case of matched duo-diodes, the technical specifications shown are for each diode. These units are matched in the forward

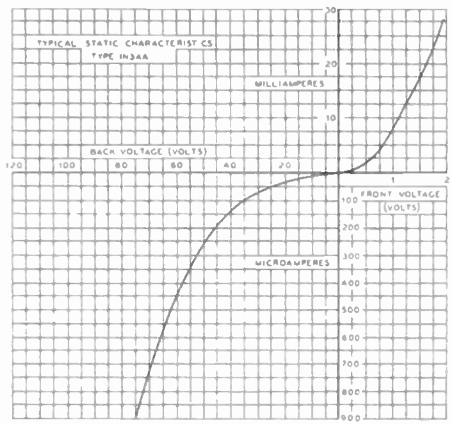

Figure 1. Static Characteristic Curve for 1N34A (Such curves are obtainable for all of the types of crystals listed in TABLE I.)

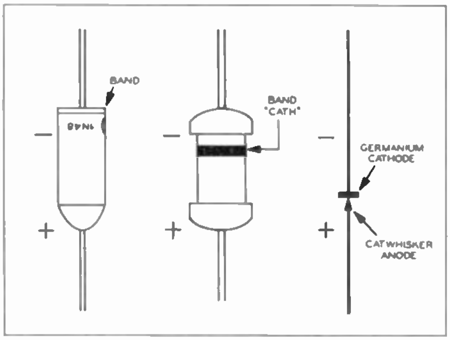

Figure 2. Cathode and Anode Identification Markings for Germanium Diodes

TABLE I. GERMANIUM DIODE SPECIFICATIONS

RTMA Type Number	Manufacturer Registered By RTMA	Description	Continuous Reverse Working Voltage (Volts. Max.)	Reverse Voltage for Zero Dynamic Resistance (Volts Min.)	Minimum Forward Current at +1 Volt (MA)	Average Anode Current (MA)	Maximum Recurrent Peak Anode Current (MA)	Maximum Forward Surge Current (MA)	Maximum Reverse Current (MA)	Remarks
1N34	Sylvania	General-Purpose Diode	60	75	5.0	50	150	500	.05@-10v., .8@-50v.	
1N34A	Sylvania	General-Purpose Diode	60	75	5.0	50	150	500	.03@-10v., .5@-50v.	Sealed in Glass
1N35	Sylvania	Matched Duo-Diode	50	75	7.5	22.5	60	100	.01@-10v.	Ratings for each Diode
1N38	Sylvania	100-Volt Diode	100	120	3.0	50	150	500	.006@-3v., .625@-100v.	
1N38A	Sylvania	100-Volt Diode	100	120	4.0	50	150	500	.005@-3v., .5@-100v.	Sealed in Glass
1N39	Sylvania	200-Volt Diode	200	225	1.5	50	150	500	.2@-100v., .8@-200v.	
1N40	Sylvania	Plug-in Varistor	25	75	12.75*	22.5	60	100	.04@-10v.	*@+1.5v., Ratings for each diode
1N41	Sylvania	Lug-type Varistor	25	75	12.75*	22.5	60	100	.04@-10v.	*@+1.5v., Ratings for each diode
1N42	Sylvania	Plug-in 100-Volt Varistor	50	120	12.75*	22.5	60	100	.006@-3v., .625@-100v.	*@+1.5v., Ratings for each diode
1N43	Western Electric	General-Purpose Diode		60	5.0	40	125	50	.9@-50v.	
1N44	Western Electric	General-Purpose Diode		115	3.0	35	100	400	1@-50v.	
1N45	Western Electric	General-Purpose Diode		75	3.0	35	100	400	.4@-50v.	
1N46	Western Electric	General-Purpose Diode		60	3.0	40	125	500	1.5@-50v.	
1N47	Western Electric	General-Purpose Diode		115	3.0	30	90	350	.004@-3v.	
1N48	General Electric	General-Purpose Diode		85	4.0	50	150	400	.83@-50v.	
1N51	General Electric	General-Purpose Diode	40	50	2.5	25	100	300	1.6@-50v.	
1N52	General Electric	General-Purpose Diode	70	85	4.0	50	150	400	.25@-50v.	
1N54	Sylvania	High-Back-Resistance Diode	35	75	5.0	50	150	500	.01@-10v.	
1N54A	Sylvania	High-Back-Resistance Diode	50	75	5.0	50	150	500	.007@-10v., .1@-50v.	Sealed in Glass
1N55	Sylvania	150-Volt Diode	150	170	3.0	50	150	500	.3@-100v., .8@-150v.	
1N55A	Sylvania	150-Volt Diode	150	170	4.0	50	150	500	.5@-150v.	Sealed in Glass
1N56	Sylvania	High-Conduction Diode	40	50	15	60	200	1000	.3@-30v.	
1N56A	Sylvania	High-Conduction Diode	40	50	15	60	200	1000	.3@-30v.	Sealed in Glass
1N58	Sylvania	100-Volt Diode	100	120	4.0	50	150	500	.8@-100v.	
1N58A	Sylvania	100-Volt Diode	100	120	5.0	50	150	500	.6@-100v.	Sealed in Glass
1N60	Sylvania	Video-Detector Diode	25	30	*	50	150	500	.03@-1.5v.	*Special test circuit in figure 4
1N63	General Electric	General-Purpose Diode	100	125	4.0	50	150	400	.05@-50v.	
1N64	General Electric	Video-Detector Diode		25	0.1*					*Special test circuit in figure 5
1N65	General Electric	High-Back-Resistance Diode		85	2.5	50	150	400	.2@-50v.	
1N69	General Electric	General-Purpose Diode	60	75	5.0	40	125	400	.05@-10v., .85@-50v.	
1N70	General Electric	General-Purpose Diode	100	125	3.0	30	90	350	.01@-10v., 41@-50v.	
1N71	Sylvania	Low-Impedance Varistor	40	50	15	60	200	1000	.3@-30v.	Ratings for each diode
1N72	General Electric	UHF Diode		5		25	75			14 19d6 Noise Figure @ 500 MC
1N73	General Electric	Germanium Quad		75	12.75*	22.5	60	100	.05@-10v.	*@+1.5v., Ratings for each diode
1N74	General Electric	Germanium Quad		75	12.75*	22.5	60	100	.05@-10v.	*@+1.5v., Ratings for each diode
1N75	General Electric	Germanium Diode	100	125	2.5	50	150	400	.05@-50v.	

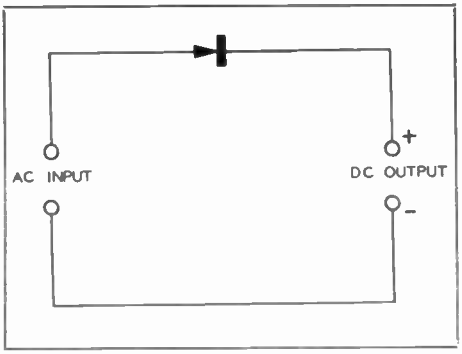

Figure 3. Polarity of Rectified D-C Output Obtained from A-C Input

direction at 1 volt, so that the current flowing through the lower-resistance unit is within 10% of that flowing through the higher-resistance unit. For varistors or quads, four diodes are specially selected so that their resistances are balanced within ±2.5% in the forward direction, at 1.5 volts. For additional balance, the diodes of each pair of varistor crystals are matched within 2 to 3 ohms. As before, the rating shown on the chart is for each diode.

The first column of the technical specifications shows the maximum continuous reverse working voltage for each diode type. This value is most important, because it indicates the amount of d.c. which can be applied across the diode for an indefinite period of time without damage to the crystal. Usually, the maximum continuous reverse working voltage is about 80% of the minimum reverse voltage for zero dynamic resistance. This latter value is more academic than practical, and is defined as the maximum reverse voltage that may be applied without danger of voltage breakdown of the unit. Voltage breakdown is the point at which the dynamic back resistance of the unit becomes zero, at which time the reverse current rapidly increases to a high value.

The average anode (rectified) current is the maximum average current which can be carried by the diode without appreciable heating of the unit, or appreciable change in its characteristics. These measurements are made at a temperature of +25° C., although the normal operating range is considered as being from —50° C. to +75° C. Greater currents than those shown may be obtained on an intermittent basis.

The maximum recurrent peak anode current indicates the peaks to which the rectified current may be allowed to rise, provided the duty cycle and wave shape are such that the RMS current does not exceed the average rating, and provided the frequency of application is at least 25 c.p.s. The maximum forward surge current is the maximum current which can flow for one second without damage to the unit.

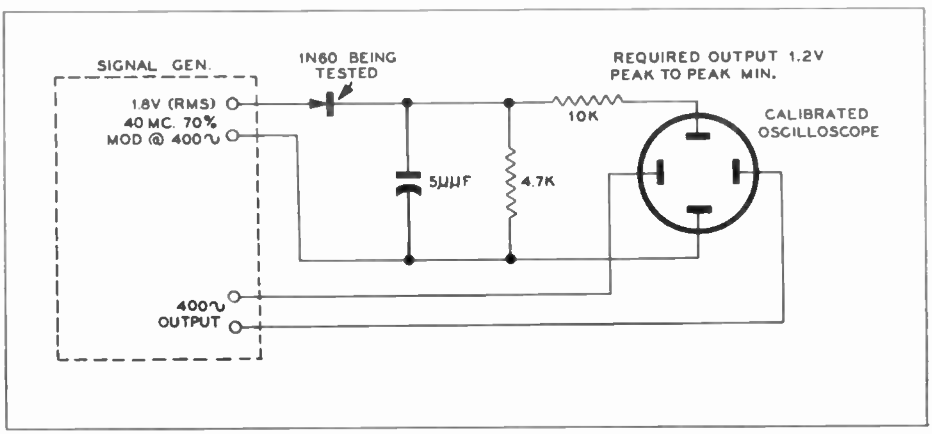

Figure 4. Special Test Circuit for 1N60

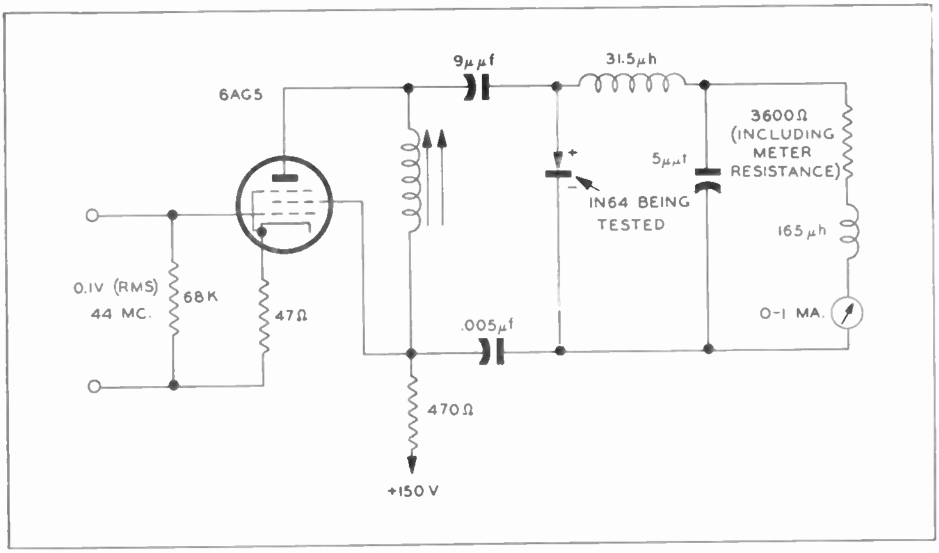

Figure 5. Special Test Circuit for 1N64 (See TABLE I for required indication.)

Minimum forward current is the smallest amount of current which can be expected to flow through the diode when a specified d-c potential is applied across it, assuming that the positive lead is connected to the anode. The corresponding forward resistance may be calculated by dividing the applied voltage by the resulting current flow. The forward-resistance characteristic is non-linear with voltage, and measurements of this characteristic are normally made at one volt. The maximum reverse current is the greatest amount of current that will flow through the diode when a specified d-c potential is applied, with the anode connected to the negative lead. By dividing the applied voltage by this current flow, the value of back resistance can be calculated.

In present-day commercial germanium diodes, the catwhisker is the anode, and the germanium pellet is the cathode. Electrically, the cathode provides positive d-c output. Figure 2 shows how the anode and cathode can be identified by the outer markings on a diode unit, and Figure 3 presents a simplified sketch showing the proper circuit connections. When inserted into a circuit, these diodes offer approximately 0.8 to 1 μμf shunt capacitance.

With deep regret we announce the passing, on August 27, of J. Carl Drumm, member of the Headquarters Technical Staff, and frequent contributor to the BULLETIN. Carl, who was in charge of the instruction in the Advanced Electronics phase of the Philco—Air Force School, will be greatly missed by his many close friends and associates throughout the TechRep Division.

Although only 42 years old, Carl was extremely capable in at least three fields: he was an accomplished author, teacher, and electronics engineer. He was an ardent "ham," a graduate of two radio-engineering schools, and at various times prior to joining Philco, had been an airline radio operator, a newspaper columnist, and the senior laboratory instructor for Capitol Radio Engineering Institute, in Washington, D. C. He joined the Philco TechRep Division in 1948, and was officially commended by the Air Force for his field-engineering work in Germany, in 1949.

Carl's very popular column, "What's Your Answer?," will continue to be a regular BULLETIN feature, because, in spite of his many other activities, he found time to prepare a number of problems for publication in future issues.

We join with the entire TechRep Division in extending our deepest sympathy to Carl's family.

PATCH FIELDS

By M. M. Elson

Philco Field Engineer

Theory and construction details of multiple-jack panels for the monitoring, switching, and testing of wired-communications circuits.

FIELD engineers are frequently required to provide for the monitoring, switching and testing of remote-control cables within the installation on which they work. While several methods are available, one of the easiest and most logical is the use of one or more jacks for each circuit. These multiple jacks, when assembled into one panel, are commonly called a *patch field*.

OPEN PATCH FIELD

Several types of patch fields can be constructed, the most common being the *open* field. This type of field (illustrated in figure 1) has the advantage of simplicity, but the disadvantage that a patch cord must be used to complete each circuit. Also, this type of field does not provide the flexibility in switching that other types provide. About all that this

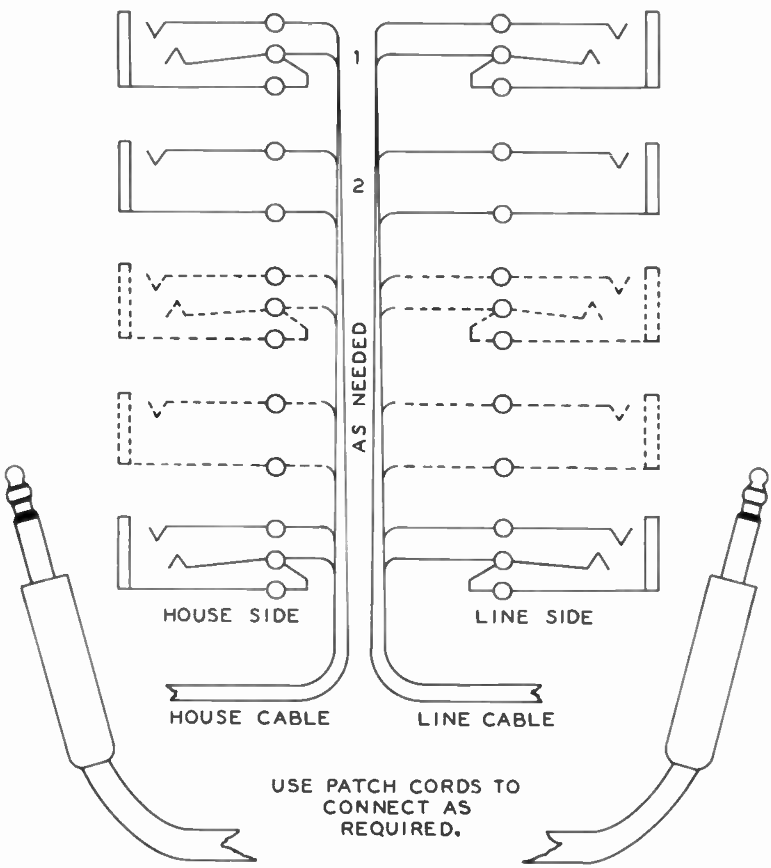

Figure 1. Open Patch Field

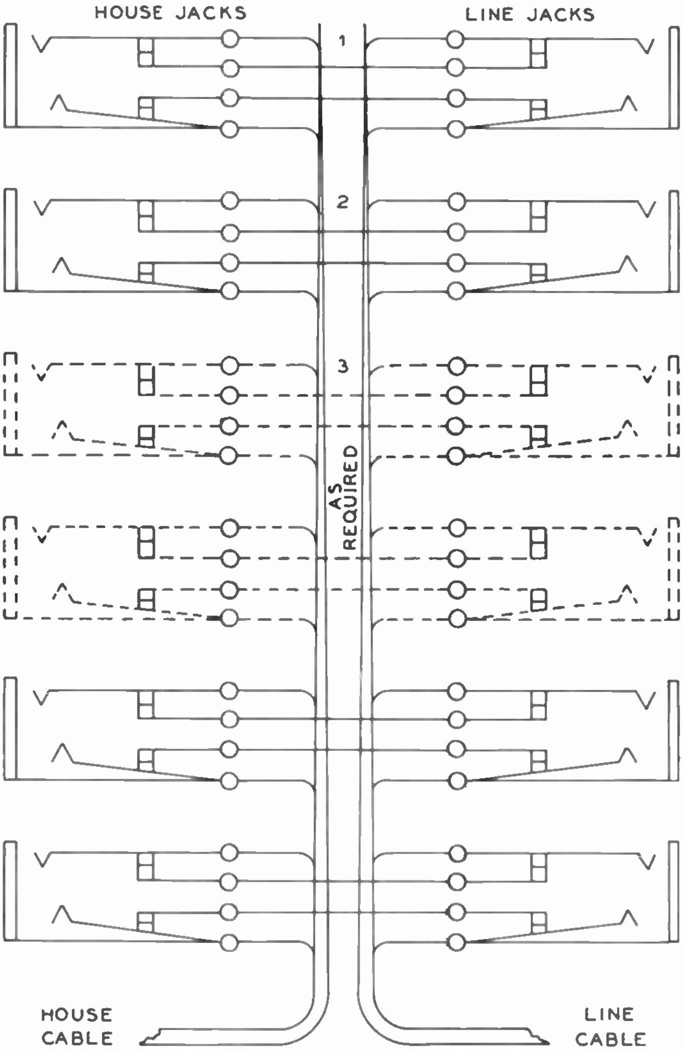

Figure 2. Parallel Patch Field

TABLE I. USE OF PARALLEL PATCH FIELD (FIGURE 2)

FUNCTION	PATCHING REQUIRED
Normal use	No patching required.
To switch equipment or lines	Patch from house (or line) jack to desired jack.
To test equipment on house side	Insert dummy plug in line jack; patch test equipment into house jack.
To test line	Insert dummy plug in house jack; patch test equipment into line jack.
To monitor house or line	Patch from house to spare jacks, and from spare jacks to line; patch monitoring equipment to spare jack.

type of field can be used for is to complete a circuit through a patch cord to any other jack within reach of the cord.

PARALLEL PATCH FIELD

An improvement on the open field is the *parallel* field, illustrated in figure 2. This field has the advantage that no patch cords are required to complete the circuits between the remote-control "line" cable and the equipments assigned to the "house" cable. Further switching between the pairs in the line cable and the pairs in the house cable is readily accomplished by the use of patch cords. If "spare-jack" sequences of jacks are provided, monitoring of either the line circuits or the house circuits is performed by use of the appropriate jack. The principal disadvantages of this type of patch field are:

a. It is impossible to measure current in a circuit without additional provisions.

b. It is impossible to series lines without additional provisions.

c. It is impossible to monitor the circuits without interference with signal channels.

NORMALLED PATCH FIELD

A further improvement on the parallel field is the use of multiple jacks for each line. A patch field of this type, commonly called a *normalled* patch field, is illustrated in figure 3. A close examination of this type of field will reveal that anything which can be accomplished by the switching of two wires (a pair) can also be accomplished with this circuit. The principal disadvantage of this type of field (and it is no small one) is that untrained personnel may make mistakes while attempting to switch within the field. Operation with the normalled patch field can often be restricted to trained personnel, however, so that this disadvantage may not exist.

It is customary to prefabricate the jack sequences before installation. After installation, all line pairs should be soldered permanently into place, and an excess of sequences provided (beyond the demand of the equipments in use by the house cable) for future expansion.

It is suggested that a voltmeter, an ammeter, and an output meter be mounted convenient to the patch field; the voltmeter and ammeter should be of the zero-centered type, or incorporate a polarity-reversing switch. One normalled jack sequence should be provided for each individual piece of equipment within the house cable. This will probably provide more jack sequences than there are lines, but the sequences will then be available to take care of possible expansion of the equipment.

All jacks should be of the "long-frame" type, commonly used in telephone service, and should be mounted on an insulating panel. Cords used should be of the shielded type, and of a length which depends on the size of the patch-field panel. Jacks designated as "J-13" by the Air Force may be used throughout, if properly wired to produce the circuits of figure 3, but a saving may be effected by the use of jacks with Air Force designations as follows:

a. For jack "A" J-13

b. For jack "B" J-3

c. For jack "C" J-7

d. For jack "D" J-13

In the wiring of patch fields, good construction practices should be observed, to reduce troubles and to simplify maintenance procedures. One of the saddest of all sights is a field engineer inquiring around for a wire stretcher. Figure 4 illustrates a desirable method of terminating a cable. Looping the cable in the manner illustrated practically eliminates wire breakage; however, if break-

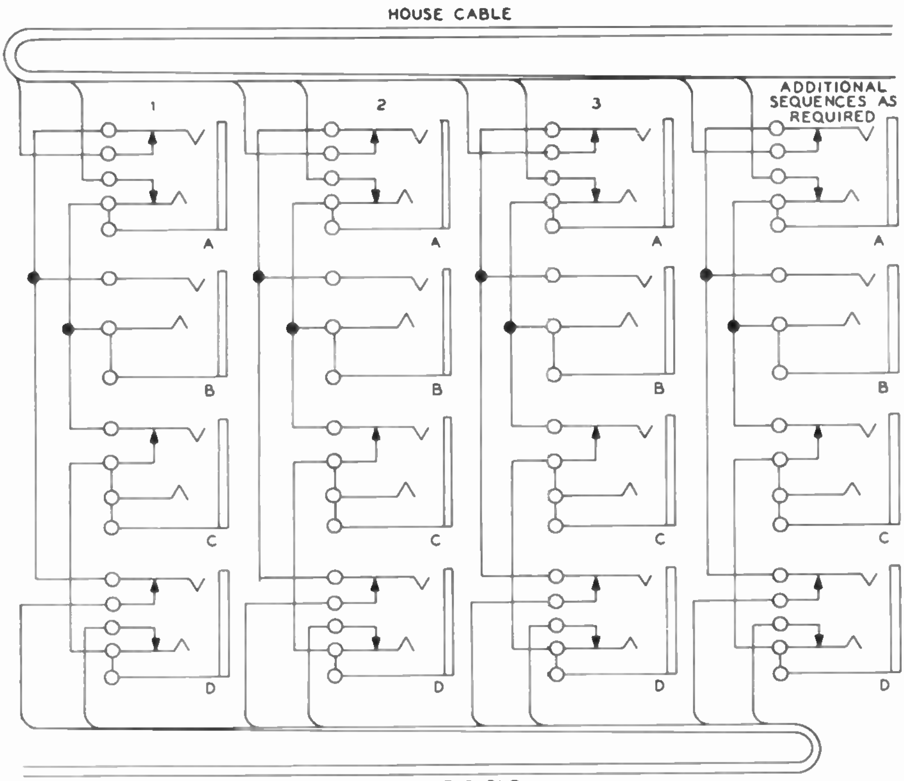

Figure 3. Normalled Patch Field

TABLE II. USE OF NORMALLED PATCH FIELD (FIGURE 3)

FUNCTION	PATCHING REQUIRED
Normal use	No patching required.
To patch from house to line	Patch from "D" to "A".
To patch from line to house	Patch from "A" to "D".
To parallel more than one equipment in house to a single line	Patch from "B" to "D", starting at jack sequence which contains line to be used.
To series more than one equipment in house to a single line.	Patch from "C" to "D", starting at jack sequence which contains line to be used.
To test line alone	Patch test equipment into "A".
To test house alone	Patch test equipment into "D".
To monitor without interruption of service	Patch monitoring equipment into "B".
To measure voltage in circuit	Patch voltmeter into "B".
To measure current in circuit	Patch ammeter into "C".
To measure db level of circuit	Patch output meter into "B".
To open house (or line) circuit	Insert dummy plug into "D" or "A", depending on whether "B" or "C" jacks are required.

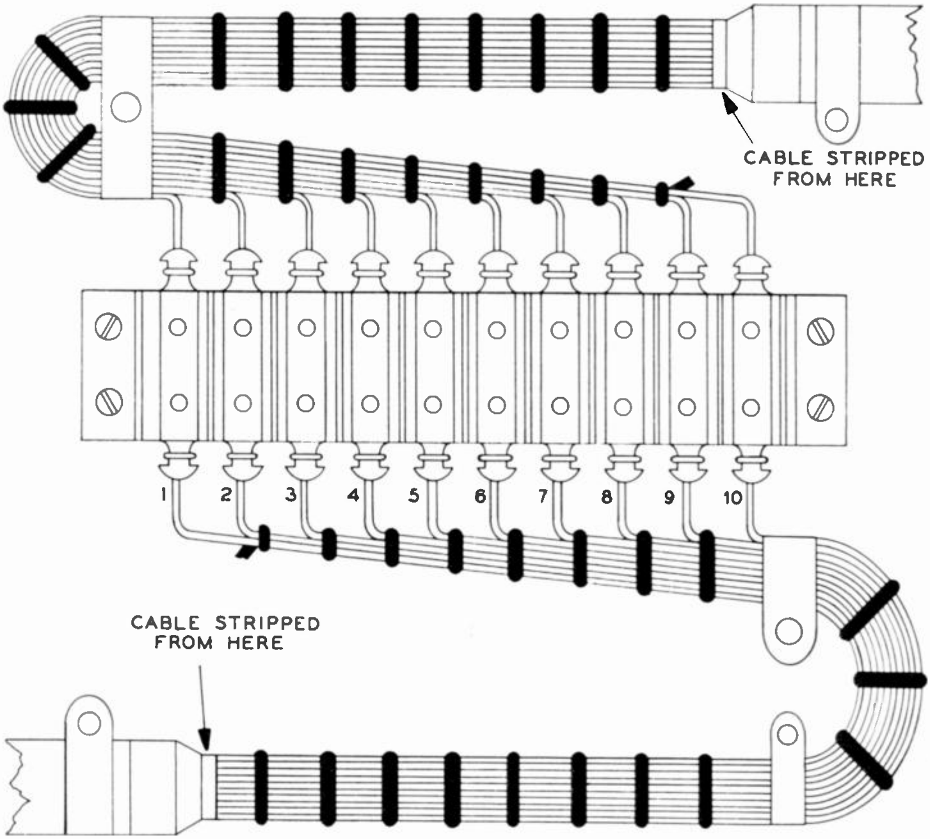

Figure 4. Proper Method of Connecting Five-Pair Cables to Terminal Strip

age should occur, there is sufficient slack available for a simple repair.

It may be that an individual field engineer will not be able to obtain the necessary jacks and panels for this type of installation. For the man who is working with the Air Force, authority for a requisition may be found in the AACS *Manual of Standard Installations* (section on "Station Wiring").

Field experience has shown that serious consideration to the methods outlined here will pay off in many hours of trouble-free circuit operation, and in considerably greater operating efficiency and flexibility of communications equipment.

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MICROWAVE PROPAGATION

By William H. Forster

Executive Engineer
Government and Industrial Engineering

A discussion of some of the factors involved in microwave propagation, and a comparison of two methods used to increase the reliability of microwave communications systems.

MICROWAVE systems are designed for a specific degree of reliability, which must be a compromise between technical and economic considerations. Equipment reliability, primary power reliability, station-house and tower stability—these factors can all be made to approach 100% performance. Propagation reliability, however, is subject to less control. Although the probability of deep fades can be minimized by locating stations close together, and by providing adequate terrain clearance, yet both of these factors increase the cost of the system. Therefore, it is desirable to study microwave propagation thoroughly, so that systems can be designed at a minimum cost consistent with the required degree of reliability.

During the years from 1945 to 1950, a great deal of effort has been expended in the study of microwave-signal propagation. This work has been summarized in several military reports; in the Proceedings of the IRE, February, 1948, by Durkee; and in the Proceedings of the IRE, June, 1950, by Millar. The emphasis in the studies was on the design characteristics for point-to-point microwave systems approaching 100% reliability as closely as economics factors permit.

The first post-War microwave system installed by Western Union was designed with a propagation fading margin of approximately 15 db, and the system used complementary diversity reception. The signal-strength records made on this system indicated that the use of diversity reception very materially improved the propagation reliability of the system. Diversity reception has also been used by the Bonneville Power Administration. All other commercial microwave systems in this country have been designed without diversity reception, but the general trend in the highest-quality systems is toward appreciably higher fading margins.

The propagation-reliability curves published by both Durkee and Millar indicate that on an overland 40-mile path, a 30-db fading margin is sufficient to provide a reliability of approximately 99.99%. It is their general conclusion that fading is caused by several factors:

Multipath fading, due both to the atmosphere and to specular reflection for which diversity reception is a cure, causes fades of approximately 15-20 db.

Tropospheric ducting causes fades of the same order of magnitude.

The most serious fades, however, are caused by a so-called sub-standard M-curve, which results from stratification of the atmosphere. Under this condition, the beam from a microwave antenna is refracted in a concave-upward fashion. In general, this means that if the beam is to reach the receiving antenna, it must start out at an angle depressed from the true line-of-sight direction.

The percentage of time that a sub-standard M-curve will cause severe fading is related to the percentage of time that the concave-upward radius of curvature is smaller than the curvature of an arc connecting the transmitting antenna site and the receiving antenna site, and passing through the minimum-clearance point in the path. The latter arc essentially defines the minimum permissible radius of curvature. For a beam refracted upward more sharply, the receiving antenna will be in the shadow of the minimum-clearance point in the path. There are obviously no cures for such a condition, except greater clearance, to permit a sharper radius of curvature, or greater signal strength, to produce a greater fading margin.

Any conservatively designed microwave system should have sufficient fading margin to insure that no tropospheric disturbances except the substandard M-curve will cause outages in the system, and the system should be designed to reduce difficulties from substandard M-curve conditions, as far as possible.

Experience indicates that a 30-db fading margin is generally adequate for paths on the order of 30 miles. Most microwave experience has been obtained in the Eastern part of the United States; therefore, a great deal of data is still to be obtained for other parts of the country. As a general guide for laying out microwave systems, it is recomended that clearances for hops greater than 25 miles in length be laid out, using grazing paths with a 2/3 earth radius. This is equivalent to a clearance of 75 feet for a 30-mile path. The required clearance then increases as the square of the path length, and is 300 feet at 60 miles. A system which is designed according to this 2/3 true-earth radius will be able to stand signal bends having a minimum radius of curvature that is independent of the length of the path. The probability of fading will then be proportional to the probability that the upward radius of curvature will be less than the permissible minimum. The substandard M-curve type of fading is not helped appreciably by the use of complementary diversity reception; therefore, for a system designed for high reliability, the propagation irregularities caused by multipath fading will be compensated for, by the fading margin of the system.

For long hops that do exhibit propagation fading greater than 30 db, it is recommended that larger antennas be installed, in order to increase the fading margin to 35 or 40 db. Diversity reception can be tried, and simultaneous recordings of field strength made, to determine whether the diversity receiver ever shows adequate signal strength during a fade-out on a primary circuit. Dollar for dollar, it is anticipated that a better circuit will always be obtained by the use of larger transmitting and receiving antennas.

To summarize, it is the consensus of the industry at the present time, as demonstrated by the kinds of microwave systems currently being installed or planned, that the best economic and technical compromise in the solution of the reliability problem is the use of greater path clearance and greater fading margins, rather than the use of diversity reception.

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Standard Frequencies and Time Signals WWV and WWVH

A summary of the information transmitted by WWV and WWVH, the two U. S. Government radio stations operated by the National Bureau of Standards

(Editor's Note: The information contained in this article was extracted from Letter Circular LC 974, January 12, 1950, published by the Central Radio Propagation Laboratory, National Bureau of Standards, U. S. Department of Commerce. The material appears in the BULLETIN through the courtesy of the National Bureau of Standards.)

The following six types of information are transmitted by WWV (N 38° 59′ 33″, W 76° 50′ 52″; near Washington, D. C.) and WWVH (N 20° 46′ 02″, W 156° 27′ 42″; near Puunene, T. H.):

(1) Standard radio frequencies
(2) Standard time intervals
(3) Time announcements
(4) Standard musical pitch
(5) Standard audio frequencies
(6) Radio propagation notices

The carrier frequencies transmitted, power levels, and modulation frequencies are as follows:

AUDIO FREQUENCIES AND MUSICAL PITCH

Two standard audio frequencies, 440 c.p.s. and 600 c.p.s., are broadcast on all radio carrier frequencies except 30 mc. and 35 mc. The audio frequencies are given alternately, starting on the hour with 600 c.p.s. for four minutes, interrupted one minute, followed by 440 c.p.s. for four minutes, and interrupted one minute. Each ten-minute period is the same. The 440-cycle signal is the standard musical pitch, A above middle C.

TIME SIGNALS AND STANDARD TIME INTERVALS

The audio frequencies are interrupted for intervals of precisely one minute. They are resumed precisely on the hour, and each five minutes thereafter. They are in agreement with the basic time serv-

TABLE I. WWV CONTINUOUS BROADCASTS

FREQ. (mc.)	POWER (kw.)	MOD. FREQ. (c.p.s.)
2.5	0.7	1, 440, and 600
5.0	8.0	1, 440, and 600
10.0	9.0	1, 440, and 600
15.0	9.0	1, 440, and 600
20.0	8.5*	1, 440, and 600
25	0.1	1, 440, and 600
30	0.1	1
35	0.1	1

* Reduced to 0.1 kw. for first 4 work days after first Sunday of even months.

TABLE II. WWVH CONTINUOUS BROADCASTS**

FREQ. (mc.)	POWER (kw.)	MOD. FREQ. (c.p.s.)
5	0.4	1, 440, and 600
10	0.4	1, 440, and 600
15	0.4	1, 440, and 600

** Entire broadcast is interrupted for 4 minutes following each hour and half-hour, and for periods of 40 minutes beginning at 0700 and 1700 UT.

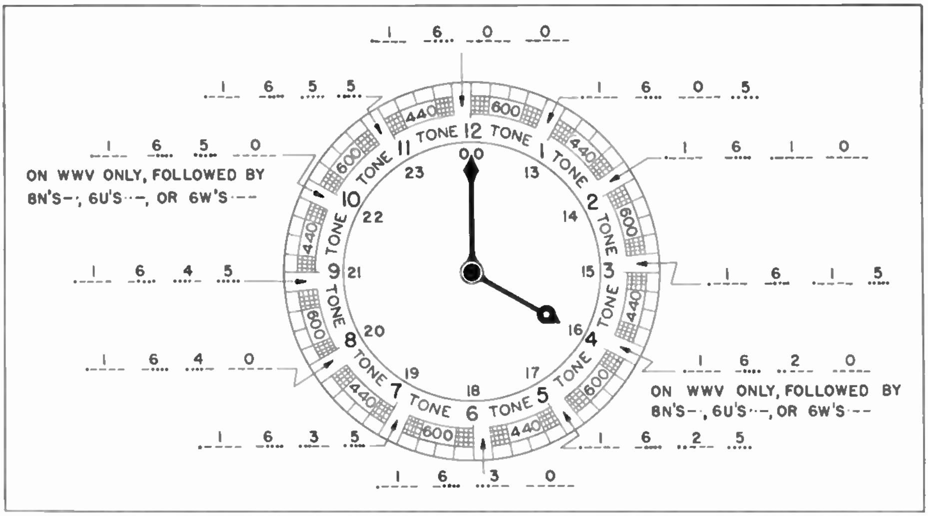


Figure 1. Universal Time Transmissions (Coded) and Standard-Tone Transmissions. The hour illustrated is 1600 to 1700, in 24-hour time. (See figure 3 for a complete analysis of the information transmitted during the one-minute intervals between the tone transmissions.)

ice of the U. S. Naval Observatory, so that they accurately mark the hour and the successive 5-minute periods.

Universal Time (Greenwich Civil Time, or Greenwich Mean Time) is announced in International Morse code each five minutes, starting with 0000 at midnight (see figure 1). Time announcements are with reference to resumption of the audio-frequency transmissions.

A voice announcement of *Eastern* Standard Time follows each telegraphic code announcement from station WWV.

At intervals of precisely one second, a pulse of 0.005-second duration is transmitted on each carrier frequency. The pulse consists of five cycles, each of 0.001-second duration (see figure 2), and is heard as a faint tick when listening to the broadcast. The seconds pulse is not transmitted at the beginning of the last second of each minute.

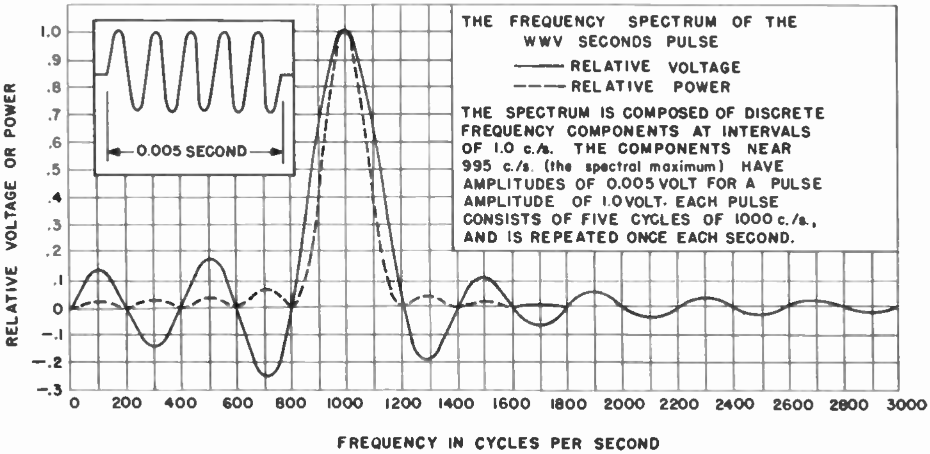


Figure 2. Frequency Spectrum of the WWV Seconds Pulse

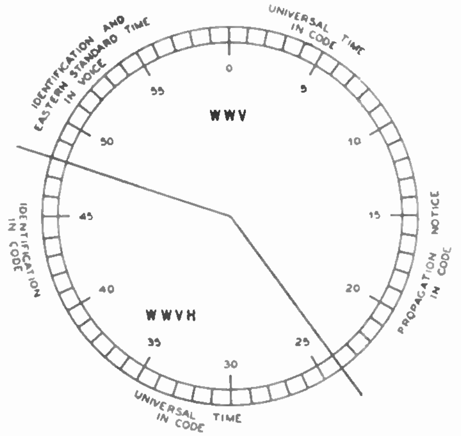


Figure 3. Information Transmitted During One-Minute Intervals Between Tone Transmissions (Note that the transmissions by each station are spaced in time so that a receiver within the range of both stations can hear each transmission separately.)

ACCURACY OF TRANSMISSIONS

The frequencies transmitted from WWV and WWVH are accurate to within 2 parts in 10^8; this is with reference to the mean solar second, 100-day interval; as determined by the U. S. Naval Observatory with a precision of better than 3 parts in 10^9. The time intervals transmitted are accurate to within $\pm$ (2 parts in 10^8 + 1 microsecond).

The frequencies received at locations in the service range are as accurate as those transmitted, for several hours per day during total light or total darkness over the transmission path. During the course of the day, errors in the received frequencies vary approximately $\pm$ 3 parts in 10^7. During ionospheric storms, transient conditions in the propagating medium may cause momentary changes as large as 1 part in 10^6.

The time intervals received are normally accurate to within $\pm$ (2 parts in 10^8 + 1 millisecond). At times, transient conditions in the ionosphere cause received pulses to scatter by several milliseconds.

RADIO PROPAGATION NOTICES FROM WWV

An announcement of radio propagation conditions is broadcast in code on each of the standard radio frequencies at 19 and 49 minutes past the hour, by station WWV. If a warning is in effect, the letter "W" (in International Morse code) is repeated 6 times following the time announcement; if unstable conditions are expected, the letter "U" is repeated 6 times; if there is no warning, the letter "N" is repeated 8 times.

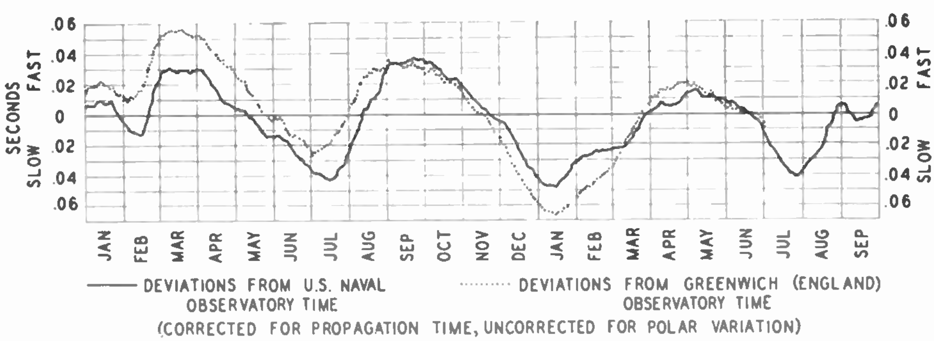


Figure 4. WWV Time Signals, January, 1948, Through September, 1949

Improved Tuning Procedure for the 5-Mc. Stages of Receiver R-19(*)/TRC-1

By Jack L. Porep
Philco Field Engineer

A simplified method of obtaining 5-mc. signal for alignment of Receiver R-19(*)/TRC-1.

(Editor's Note: This article was received after the July issue went to press. Its similarity to Ken Seaton's article in that issue is interesting: two Philco men, working independently, arrived at basically similar solutions to the problem of improving and simplifying maintenance techniques.)

In many cases in the field, R-19 receivers have been badly misaligned, making it very difficult to retune them by the procedures outlined in TM 11-2601. This condition has been the result of attempts by inexperienced personnel to align the 5-mc. i-f amplifier, limiter, and discriminator stages by the use of the wrong harmonics. Once these stages are badly detuned, it is extremely difficult, if not impossible, to tune up the receiver in the normal manner.

The procedure outlined below allows the i-f amplifier, limiter, and discriminator stages to be easily retuned to 5 mc. in a short time, without the use of extra test equipment or the modification of any component. Test Oscillator TS-32(*)/TRC-1 is the only device needed.

DETAILED TUNING PROCEDURE

First, remove any crystal that may be installed in the receiver. Remove the bottom cover plate and the top lid, and stand the receiver on the left end to allow easy access to the necessary parts. Plug the test oscillator into the power socket at the center of the set. The short length of coaxial cable should be connected to the test oscillator, in the usual manner, but the other end of this cable should not be connected to the antenna socket on the receiver panel. Instead, it should remain free, with the center conductor serving as a probe. (The collar on the coaxial fitting should be taped back temporarily with rubber tape.)

The 80.0-mc. transmitter crystal should then be placed in the test oscillator. This is the key to the whole method. The crystal marked "80.0 MC." has a fundamental frequency of 833.33 kc., the sixth harmonic of which is 5000 kc., or 5 mc. Place the test-oscillator switch to the CARRIER ON position. The MODULATION ON switch has no effect on the alignment process, so it is left in the "off" position. (This is explained by the fact that at the sixth harmonic of the test-oscillator crystal there is insufficient deviation produced by the phase modulator to make the receiver discriminator stage respond to the 1000-cycle tone; therefore, the tone is not heard from the speaker when the 5-mc. signal is passed to the discriminator circuit.)

With the receiver meter switch on position 1, the coaxial probe is placed in contact with the grid (pin 4) of the second mixer tube, V105. Transformers T106, T107, T108, and T109 are adjusted (the meter is used in the same way as described in TM 11-2601). The settings of all the other tuning controls have a negligible effect on this portion of the alignment procedure. When the receiver is properly aligned, a reading of 40 or higher is indicated by the meter (position 2). On every receiver tested, this method was easy to use, and the lack of a 1000-cycle modulating tone was no handicap, since there were no spurious responses involved.

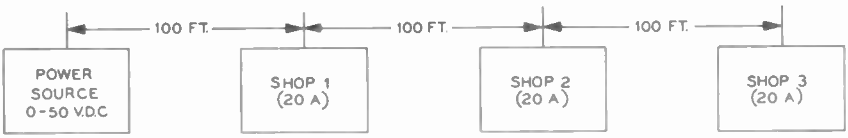
After the 5-mc. i-f amplifier, limiter, and discriminator stages are properly aligned, the remainder of the receiver circuits are tuned in the normal manner, using the test oscillator and the proper crystal for setting up any desired operating frequency.

As a point of interest, the 5-mc. i-f amplifier, limiter, and discriminator stages of a good receiver were completely detuned purposely. Then the receiver was completely realigned in less than ten minutes by the new method of tuning. The author has used this technique with much success in Korea and Japan, where a great deal of AN/TRC-1 equipment is used.

WHAT'S YOUR ANSWER?

By J. Carl Drumm, Hq. Technical Staff

This month's problem was submitted by Al Calahan, an instructor at Headquarters. The problem is to connect a d-c power source (rated at 0—50 volts, 100 amperes) to three shops that are spaced at intervals of 100 feet (see the figure).



In each shop, the power line must feed a piece of electronic equipment which draws 20 amperes with an input voltage of 28 volts, d.c. The only wire available for this power line is a 900-foot roll of single-conductor cable, the conductivity of which is equivalent to No. 6 B. & S. gage copper wire (resistance, 0.4 ohm/1000 feet). Calculation shows that if the system were wired by the usual parallel-connection method, and the power-supply voltage were adjusted so as to deliver 28 volts at Shop No. 1, then, if each shop drew 20 amperes, the voltage at Shop No. 3 would be only 23.2 volts. Or, if the power-supply voltage were adjusted to deliver 28 volts to Shop No. 3, then the voltage at Shop No. 1 would be 32.8 volts. Of course, the situation would not be quite as bad as indicated by these figures, because the shops with reduced voltage would draw less current than 20 amperes, and the shops with excessive voltage would draw more than 20 amperes. But in either case, part of the equipment would be operating improperly.

A simple method can be devised, however, for wiring this system so that with the shops operating singly, jointly, or collectively in any combination, the power-supply voltage can be so adjusted that no shop will have an input that differs by more than 1 volt from the prescribed 28-volt figure.

What wiring diagram would *you* use, remembering that only 900 feet of single-conductor cable is available?

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

The black box contains a vibrator equipped with a pair of contacts, at B, operating on a 50% duty cycle. (Note: the ammeter was misplaced in the original problem; it should have been shown as illustrated below.)

When the contacts at B are open, the voltmeter reads practically the full battery voltage (6 volts). The only current flowing through the current coil of the wattmeter is about 0.6 ma. drawn through the 10K resistance of the voltmeter. Under this condition, the three meters read as follows:

Voltmeter —6 volts
Ammeter —0 amperes
Wattmeter—3.6 milliwatts

When the contacts at B are closed, the voltmeter is short-circuited, and reads zero voltage. Current through the ammeter is then limited only by the internal resistance of the meters and the battery. and amounts to 10 amperes. The wattmeter. however, reads zero because it has no voltage impressed across it. The meter readings are then:

Voltmeter — 0 volts
Ammeter —10 amperes
Wattmeter— 0 watts

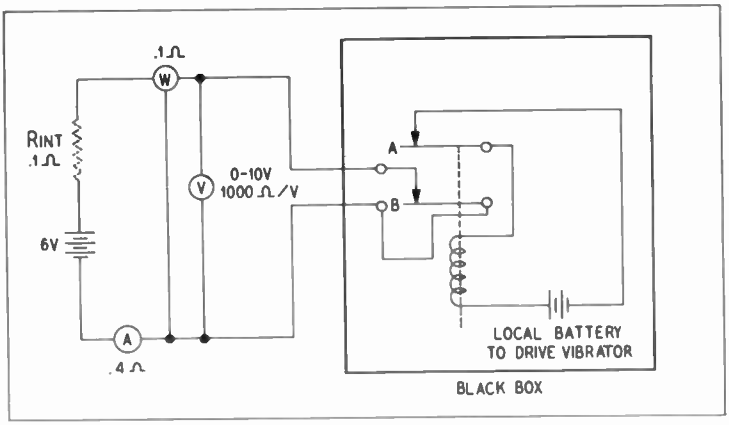
Assuming that the vibrator can turn out perfect square waves; and that there is no sparking at the contacts; and that the inductance of the meter windings does not cause current to rise and decay exponentially when the contacts close and open; and that the wiring of the circuit has zero resistance; and that the battery does not heat under intermittent short circuits, thus raising its internal resistance . . .

Then, the three meters will maintain a steady reading equal to the average of the two sets of readings given above:

Voltmeter —$(6 + 0)/2 = 3$ volts
Ammeter —$(10 + 0)/2 = 5$ amperes
Wattmeter—$(3.6 + 0)/2 = 1.8$ milliwatts

(If you think this one is pretty old, try assigning inductances to each of the three meters, and calculating the frequency of the vibrator necessary to produce a given set of readings, taking into account the exponential rise and decay of current mentioned previously.)

J. C. D.



ELECTRICAL DEFLECTION SYSTEM FOR A-1C GUNSIGHT

Submitted by Bud M. Compton, Philco Team Leader

Cabling data and circuit details of an electrical system for deflecting the A-1C gunsight during "bore-sighting."

Here is a neat trick for anyone working with the A-1C gunsight (often used in conjunction with gun-laying radar). The idea and the prototype model were developed by Mr. Robert Fouitz, field engineer for the AC Spark Plug Division, General Motors Corporation. The information is passed along for the benefit of anyone else who may be working with this gunsight (which is quite standard for fighter aircraft). With this system, the sight can be deflected during bore-sighting by simply operating the switches. This eliminates the necessity for removing the sight cover and mechanically moving the gyros to deflect the sight.

The required parts are as follows:

QUANTITY	DESCRIPTION
1 ea.	Switch, wafer, double-pole six-position
1 ea.	Switch, toggle, single-pole, single-throw
1 ea.	Plug, Amphenol size 28, AN-3106B-28-16P
1 ea.	Plug, Cannon #2083-50, AN-3106B-28-16S
As req.	Cable, interconnecting

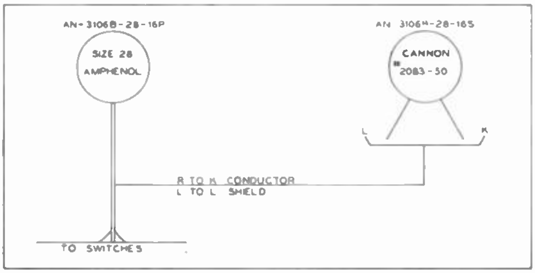


Figure 1. Interconnection Diagram of Deflection System (The Cannon plug goes to the gunsight field-testing unit used for checking the frequency of the radar inverter. The Amphenol plug goes to the gun-sight deflection computer.)

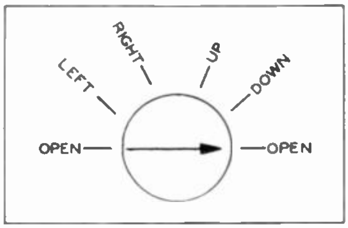


Figure 3. Front View of Deflection-Control Switch, Showing Function of Switch in Each Position

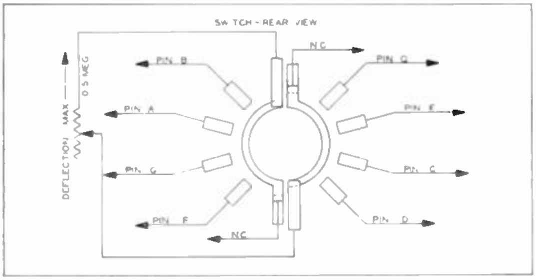


Figure 2. Schematic Diagram of Deflection Control Switch (The pin designations refer to the pins in the Amphenol plug connected to the computer.)

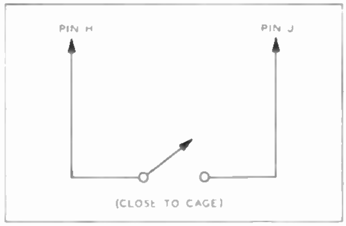


Figure 4. Gyro-Caging-Switch Connections (The pin designations refer to the pins in the Amphenol plug connected to the computer.)

PHILCO TECHREP DIVISION PUBLICATIONS

Reprints of the following Philco technical publications are currently available at Headquarters, and can be obtained by any Philco Field Engineer whose duties require them. Requests should be addressed to Potter Hallinger, Correspondence Department. There is, of course, no charge involved in such requests.

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Orders should be addressed to Publications Department, Philco TechRep Division, 22nd St. and Lehigh Ave., Philadelphia 32, Pa. Remittances should be in cash, check, or postal money order, made payable to Philco Corporation.

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Vacation Issue

This issue of the BULLETIN bears a double dateline, August *and* September, in consideration of the vacation season.

You will note that the issue is larger than usual, and that there is a larger percentage of technical information than in previous issues. The "Contributors" column has been dropped, and the number of technical articles has been increased. We believe that our readers will endorse this change in policy.

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