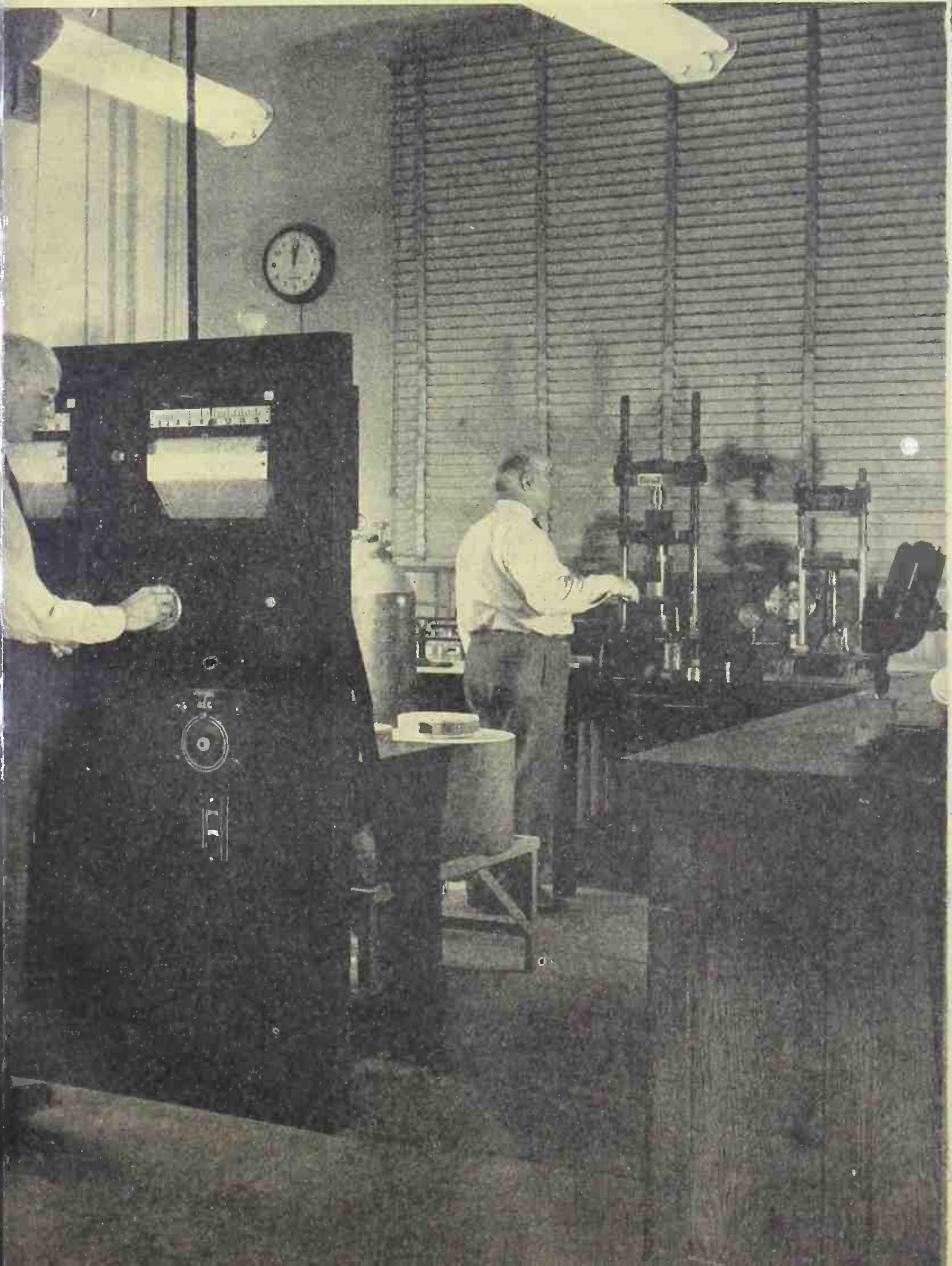
DECEMBER, 1950

RADIO & TELEVISION NEWS

# RADIO-ELECTRONIC Engineering



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# RADIO-ELECTRONIC Engineering

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## DECEMBER, 1950

### DEPARTMENTS

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COVER PHOTO—Courtesy of National Bureau of Standards

General view of the laboratory where high-dielectric ceramic capacitors under development at the National Bureau of Standards are produced. Batches of systematically varied compositions are processed into capacitor plates of 0.003 to 0.006 inch thickness, and the performance properties then studied.

# ELECTRODES for H.F. HEATING

**By R. A. WHITEMAN**
Consulting Engineer, Chicago

***A discussion of the performance of electrodes as a component part of h.f. dielectric heating equipment.***

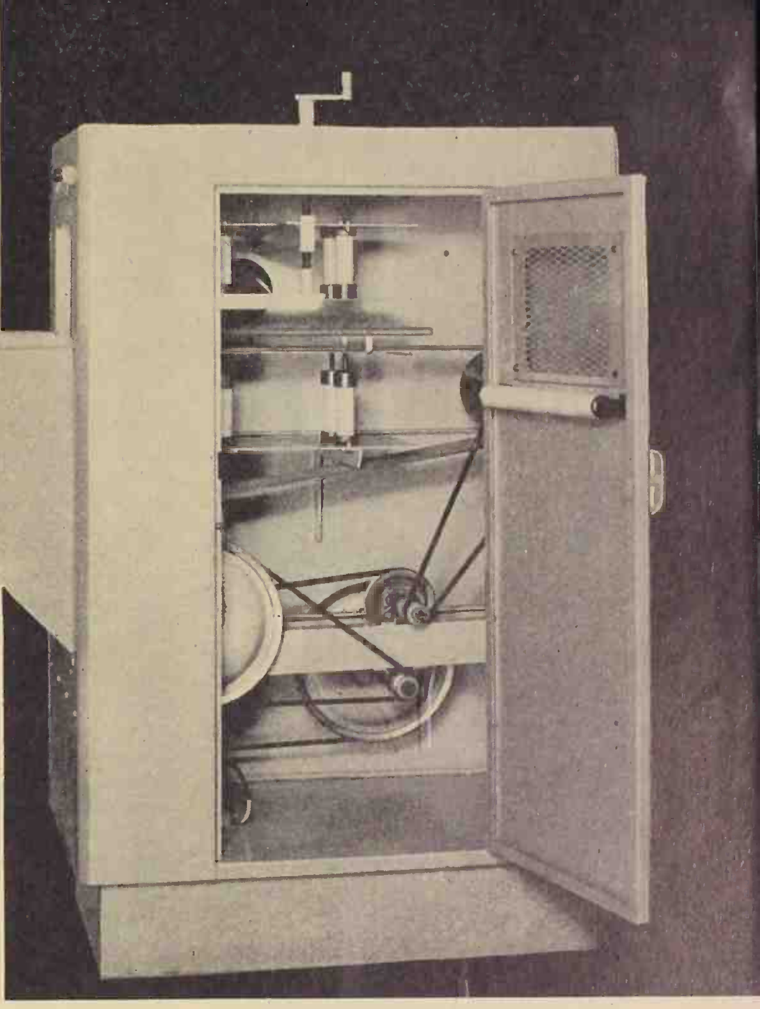


Fig. 1. Conveyor belt arrangement for dielectric heating showing electrodes and transmission line through side of cabinet.

THE h. f. heating of dielectric material has reached new production heights in recent years. With this increase in production there has been a parallel increase in the diversity of applications and also design problems. These applications include the heating of plastics, melting frozen foods, deinfestation of pharmaceutical products, and the gluing of wood. The design and application problems range from the d.c. power supply through the power-oscillator circuit to the final arrangement of the electrodes. This article will deal specifically with the performance of the electrodes as a component part of the h.f. dielectric-heating equipment. The effectiveness of the electrodes naturally depends upon their shape as well as the shape and dielectric properties of the material being heated.

A statement of certain fundamental high-frequency properties of dielectrics is of value at this time in order to emphasize the important design parameters which must be considered. These fundamental relations depend upon the theory that nonmetallic solid materials have few free electrons and the effect of the electric field on the dielectric molecules becomes very important. In accordance with this theory, the electric field causes a definite displacement of the electrons within the atoms and also a displacement of the atoms within the molecules. These displacements have translational as well as rotational components and are most important in the range of frequencies used for radio-frequency heating. As the electric field components of the molecules are rotated to line up with the electric field, a displacement of charge within the material takes place. As a result of this effect, the displacement current in the circuit is greater due to the presence of the dielectric material than that occurring due to free space. The ratio of the former to the latter displacement current for a given electric-field intensity is defined as the dielectric constant of the material.

Since the changes in the molecular configuration impart kinetic energy to the thermal motion and are due to the applied electric field, an equivalent displacement current must flow into the material in phase with the voltage impressed on the material.

If the power loss in watts for each cubic inch of the dielectric is $P$ and the electric field intensity in volts per inch is $E$, then by definition the ratio of the power $P$ to $E^2$ is the effective conductivity $\sigma$. The numerical value of $\sigma$ is generally a very involved function of the frequency and is not a constant.

For materials that have a very small conductivity, the algebraic relation for $\sigma$ is:

$$\sigma = 2\pi f \varepsilon p \quad \ldots\ldots \quad (1)$$

where $\varepsilon$ is the dielectric constant at the operating frequency $f$ and $p$ the power factor of the dielectric material. By equating the right member of Eqt. (1) to the ratio of $P$ to $E^2$ and solving for the power dissipated in a cubic inch, the result becomes:

$$P = 2\pi f \varepsilon E^2 p \quad \ldots\ldots \quad (2)$$

Although this equation is a simple relation between the power dissipated and the fundamental parameters of the dielectric circuit, it clearly indicates the dependency of the power dissipation upon the power factor, frequency, dielectric constant, and the electric field intensity. It is possible to make the frequency $f$ as well as the dielectric constant $\varepsilon$ independent of the electrode shape. It is, however, impossible to make the electric field intensity throughout the dielectric material independent of the electrode shape. In order to have a uniform distribution of power density, it is necessary to have a uniform electric field intensity, which can be achieved in practice by properly arranging the electrodes.

The objective in dielectric heating is to first develop a sufficient power density for the required heating time interval and then the problems which follow must be solved as secondary items. Thus, it may seem advisable at first to have the equipment operate on one of the lower frequency channels in order to obtain a uniform potential across

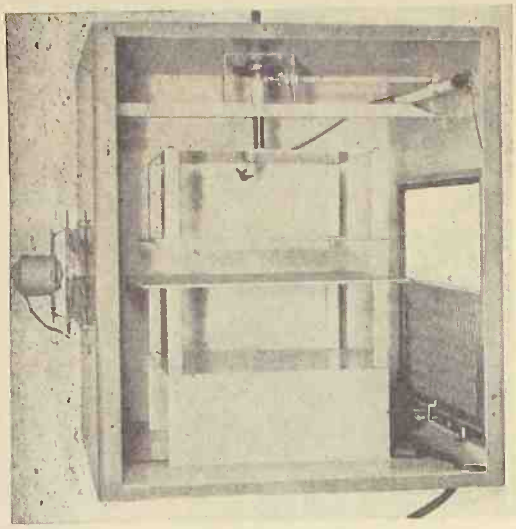


Fig. 2. Adjustable height work oven for heating dielectric materials.

the surface of the electrodes. This would be accomplished by having the wavelength large compared with the size of the electrode. However, with this condition, the voltage will be uniform on the electrodes but the electric field intensity $E$ will be high and flash-over as well as internal sparking within the dielectric material will exist. This means that a higher frequency $f$ must be used and the field intensity $E$ will then be within a numerical value which does not produce dielectric breakdown. In some applications the electrodes will then be large compared with the wavelength and means of compensating for the variation in voltage along the electrode must be included in the design of the electrodes.

This undesirable effect is very well exemplified in large electrodes used in the high-frequency wood-gluing process. It is quite common in this industry to glue sheets of wood from 10 to 20 feet and sometimes longer and use frequencies in the region of 37 megacycles. With the average dielectric constant of the wood and glue equal to $\varepsilon$, the wavelength of the voltage wave on the long electrodes is given by the formula:

$$\lambda = \frac{984}{f\sqrt{\varepsilon}} \quad . \; . \; . \; . \; . \; . \; . \; . \; . \quad (3)$$

where $\lambda$ is the wavelength measured in feet and $f$ is the frequency in megacycles. The voltage wave will have a maximum at the open end of the electrode and a minimum at a quarter wavelength from the open end. If the wavelength of the voltage wave is very much greater than the length of the electrode, it is possible that the ratio of the maximum voltage to the minimum is greater than 0.9. To accomplish this desirable voltage distribution, however, the frequency would be entirely too low and the power concentration for a permissible operating voltage would be insufficient for a reasonably short gluing cycle. This condition can be improved somewhat by feeding the voltage to the center of the electrode plates with each half of the press considered separately, thereby enabling the frequency to be twice that where the voltage is applied at one end. If it is physically permissible, it is advisable to introduce auxiliary tuning of the long electrodes by using small inductances spaced along the electrodes. These inductances should be connected in parallel across the electrodes and tune the capacitance of the press to parallel resonance. To accomplish this, each inductor should have $n$ times the total inductance required to tune the loaded electrodes. This is expressed by:

$$L = \frac{n\,10^6}{4\,\pi^2\,f^2\,C} \quad . \; . \; . \; . \; . \; . \; . \; . \quad (4)$$

where $n$ is the number of equally spaced inductors, $f$ is the frequency in megacycles, $C$ is the capacitance of the loaded electrodes in micromicrofarads and $L$ is the inductance of each coil in microhenrys. At resonance the load will present a resistive impedance of:

$$Z = \frac{2\,\pi\,f\,L}{\tan\theta} \quad . \; . \; . \; . \; . \; . \; . \; . \quad (5)$$

where $\tan\theta$ is the ratio of effective resistance to the capacitive reactance of the dielectric load.

The distance between the tuning inductors in feet is given by:

$$d = \frac{5.48}{f\sqrt{\varepsilon}}\cos^{-1}\frac{E_1}{E_2} \quad . \; . \; . \; . \; . \; . \; . \quad (6)$$

where the ratio of $E_1$ to $E_2$ is the ratio of the minimum voltage to the maximum voltage on the electrodes and is generally made greater than 0.9.

As a typical example to illustrate this method of using special electrodes for high-frequency heating, consider the following set of conditions:

$f = 37$ megacycles
$\varepsilon = 2.5$
$\tan\theta = .05$
$l = 20$ feet
$E_1/E_2 = 0.9$
$C = 1000$ micromicrofarads

$$d = \frac{5.48}{37\sqrt{2.5}}\cos^{-1} 0.9 = 2.44 \text{ feet}$$

is the spacing of the inductors. In an application very similar to this example, 7 tuning inductors were used in order to provide a fairly uniform voltage along the electrodes. The inductance of each coil was:

$$L = \frac{7 \cdot 10^6}{4\pi^3 \cdot 32^2 \cdot 1000} = .130 \text{ microhenrys}$$

and the impedance of the dielectric load:

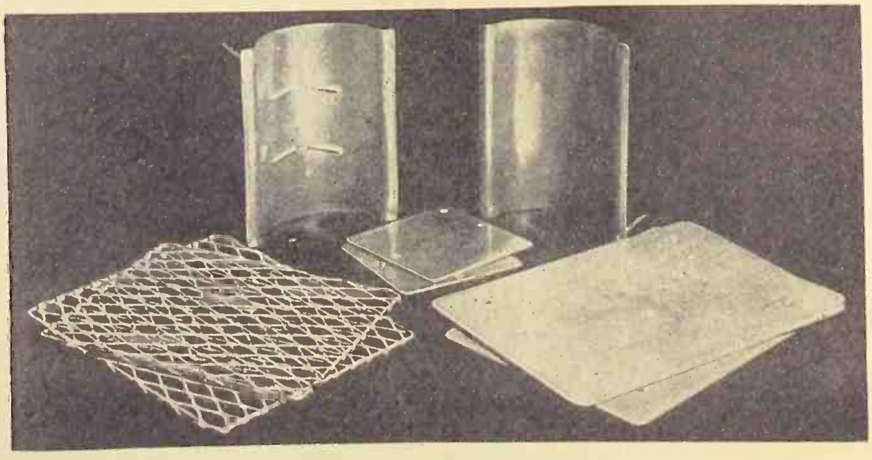
$$Z = \frac{2\pi \cdot 37 \cdot .130}{7 \cdot .05} = 86 \text{ ohms}$$

This example illustrates the application of a special electrode for high-frequency heating where the geometrical length of the electrode is very much greater than ¼ of an electrical wavelength.

In applications where the electrodes are short compared with the quarter wavelength, problems of voltage variation do not exist. Compensation by tuning is, therefore, not necessary and the only adjustment necessary is that of varying the distance between electrodes. This condition exists when heating plastic preforms and is illustrated in Fig. 2. In the case of heating plastic preforms it is usually necessary to supply an air movement parallel to the electrodes. The air movement is provided by the fan and removes water vapor and volatile chemicals, thereby preventing them from condensing on the upper electrode.

If instead of having the electrode long, it is short compared with ¼ of the electrical wavelength, then the voltage applied to the electrodes will be uniform. The electrodes under these circumstances may be considered to be equipotential surfaces and the electric field intensity will be substantially uniform within the dielectric if the dielectric has a special shape. In general, any piece of dielectric will require electrodes of a special shape in order to obtain a uniform electric field intensity within the given dielectric.

Fig. 3. Curved and flat electrodes used in dielectric heating equipment.



A general solution of this type of problem may be easily expressed but not always easily evaluated. For the general solution, refer to Fig. 4 where the electrodes are above and below the $x$-axis and the electric field is parallel to the $y$-axis within the dielectric. Let the potential within the dielectric be:

$$V_i = f_i\,(\rho_i,\ \theta) \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \qquad (7)$$

and must satisfy the relation $\delta V_i / \delta y$ = a constant to insure that the field intensity within the dielectric is a constant. With a uniform field within the dielectric, the equation of the potential external to the dielectric is:

$$V_e = f_e\,(\rho_e, \theta) \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \qquad (8)$$

and the potential at an external point through which the electrode must pass:

$$V_e' = f_e\,(\rho_o, \theta) \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \qquad (9)$$

To find the equation of the equipotential surface with a potential equal to $V_e'$, set $V_e$ in Eqt. (8) equal to the constant potential of Eqt. (9) as follows:

$$V_e = V_e' = f_e\,(\rho_e, \theta) = f_e\,(\rho_o, \theta) \qquad (10)$$

and solve for $\rho_e$ in terms of the angle $\theta$ and the constant $\rho_o$. Then the equipotential surface can be expressed as:

$$\rho_e = F\,(\theta) \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \qquad (11)$$

A metallic electrode which is constructed so that it satisfies Eqt. (11) will be coincident with the equipotential surface and thereby produce the desired electric field. This mathematical approach is generally cumbersome and can be replaced by one which is not as accurate but easier to apply. By referring to Eqts. (7), (8) and (9) and rewriting them for rectangular coordinates they become:

$$V_i = g_i\,(x, y) \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \qquad (12)$$

$$V_e = g_e\,(x, y) \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \qquad (13)$$

Then since the voltage across the electrodes is a constant for any one application, this voltage is:

$$V_t = \int_0^s E_i\,dy + \int_s^a E_e\,dy \quad . \quad . \quad . \quad . \qquad (14)$$

$$V_t = E_i\,(s + a\,\varepsilon) \quad . \quad . \quad . \quad . \quad . \quad . \qquad (15)$$

where $s$ is the thickness of the dielectric and $a$ is the distance of a flow line through the air gap between the dielectric and the electrodes. Since the total applied voltage $V_t$ and the internal electric field intensity are constants, then $(s + a\,\varepsilon)$ must be a constant for small values of $a$.

It is readily observed that the position of the electrodes can be located point by point using Eqt. (15). This method not only applies to dielectrics with flat surfaces but to surfaces of any shape even though it is understood that the numerical value of $a$ must be small compared with $s$. In general, the shape of the electrode surface will not be a plane but will resemble such surfaces as shown in Fig. 3.

In many applications, it is not necessary or advisable to attempt to form the electrodes to conform with an equipotential surface providing an alternate solution to the problem is available. In many heating procedures, moving the object to be heated has long been standard practice. This same procedure may and is being used to advantage when dealing with the high-frequency heating of dielectric material. If the dielectric material is moved or passed between electrodes of the required shape the higher intensity electric fields will shift from one region of the dielectric to another. Such a change or shift will enable the heat generated within the dielectric to conduct from the hotter to the cooler portions, thereby producing a more uniform heating of the material. The procedure is very well applied in the *SIECO* dielectric belt oven shown in Fig. 1. The oven is loaded from the long end which projects to the left of the photograph. The dielectric material moves on a traveling belt of specially treated fabric. The speed of travel through the oven is adjustable over a wide range and the vertical spacing between the upper and lower electrodes is also adjustable from ½ inch to approximately 6 inches. The electrodes generally extend 18 inches along the belt and are slightly wider than the belt. These electrodes are used under an applied voltage of 8000 volts at 27 megacycles. This belt driven system is constructed to be used with a *SIECO* dielectric heating unit with the aid of a transmission line connection. This type of connection is shown in Fig. 5. The oven is equipped with a stop button both at the loading end and at the exit end so that in the event of a mechanical fouling of the dielectric material, or for any other reason, the operator can then shut off the entire machine including its radio frequency power by one simple stop button operation. The fluorescent lamp located and visible inside the door is grounded at the near end and at the far end is equipped with a 4-inch long antenna which picks up sufficient radio-frequency power from the adjacent heating electrode to operate the fluorescent tube and to give a soft uniform light of good intensity indicating normal operation of the oven. Also visible in the photograph is a safety interlock switch so that if the door is opened the entire equipment is disconnected from the power line. This equipment provides an excellent method of production-line handling as well as a method of obtaining uniform heating of dielectric materials. It has been used for protective deinfestation of pharmaceutical and plastic products.

This analysis has indicated the various conditions which must be considered when selecting electrodes for high-frequency heating of dielectric material. The primary requisite in this work is that of uniform heating of the dielectric in order to avoid "hot spots."

The author wishes to thank *Sherman Industrial Electronics Company* of Belville, N. J. for the accompanying photographs and other helpful information.

* * *

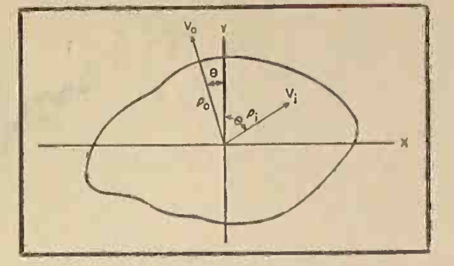
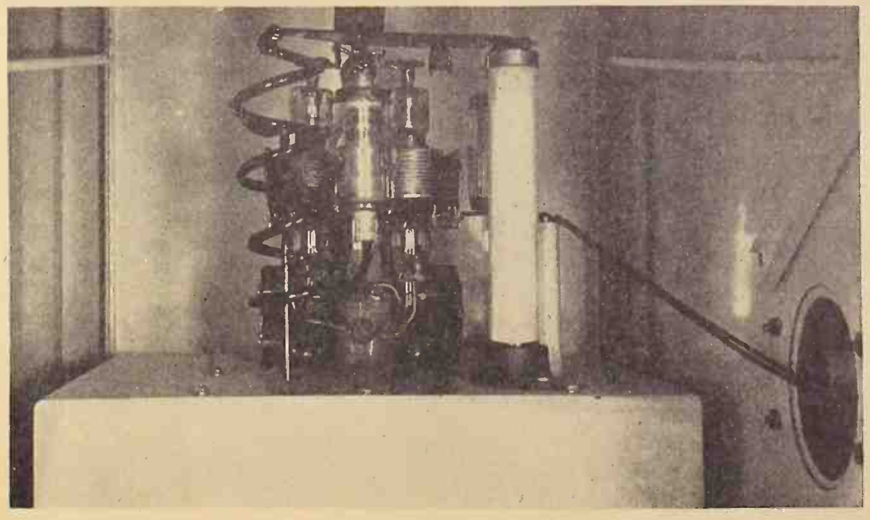
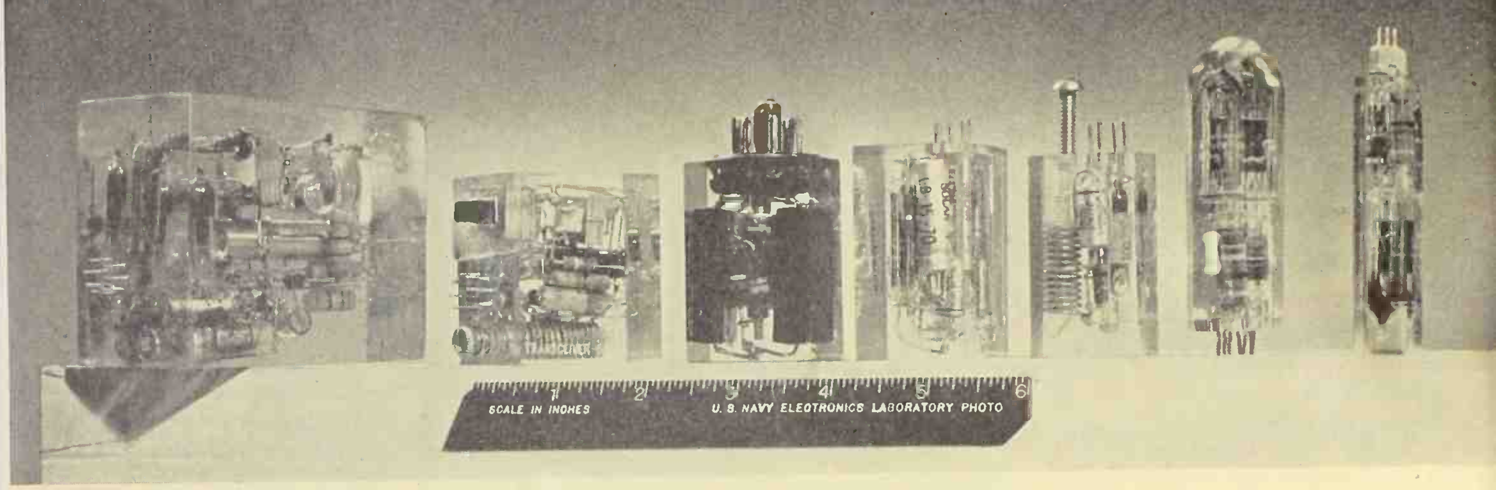


Fig. 4. A section of dielectric symmetrical about the z-axis with internal and external potentials indicated.

Fig. 5. 27 mc. oscillator showing transmission line connection to adjacent electrode enclosure.



# Potted Electronic Circuits



Miscellaneous units cast in NEL casting resin.

### By HAROLD E. BRYAN

***Casting or potting materials protect electronic circuits from effects of humidity, fungus, shock, and vibration.***

IT USED to be standard practice to construct electronic equipment for the Naval forces in a most rugged and substantial manner, in order that it might continue to operate and withstand the rigors of the service. This worked fine, and the equipment did indeed stand up in Naval shipboard operation. However, with the coming of the late war the electronics art progressed very rapidly. Ships no longer are able to carry all the electronic equipment considered desirable, or even in some cases essential, if it is built in the usual rugged and massive manner.

Attention has therefore been directed to the miniaturization and subminiaturization of the many electronic devices required by modern Naval warfare. The advent of miniature and subminiature electron tubes, as well as the production of smaller and smaller components such as resistors and capacitors, has made the reduction in size not only desirable but possible.

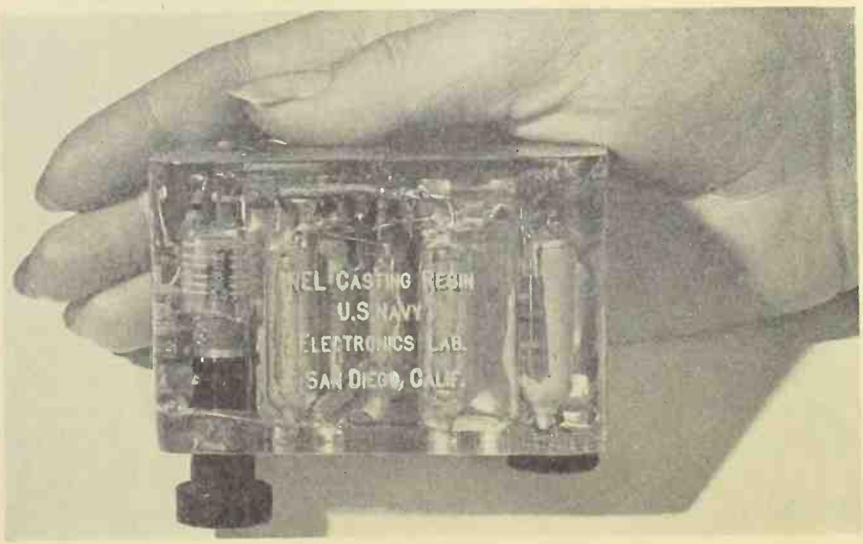
This reduction in size of components and complete units has, nevertheless, not been achieved without new complications. True miniaturization of a circuit requires the elimination of the common chassis as it is usually known, since such a base occupies considerable space by itself. Making the components more or less self-supporting, or using very small flat sub-chassis where required for support, can in itself save a very substantial amount of volume.

Other problems which must be met in the new designs are those involving damage from such things as high humidity, fungus growths, and mechanical damage from handling and shock and vibration. It has been stated that in certain areas during the last war as much as 60% of the electronic equipment was useless as received, due primarily to the effects of humidity and fungus. This is a very high casualty rate for equipment that has not even been used!

Attempts have been made to meet the attacks of humidity and fungus by means of hermetic sealing. This would seem to be ideal, since all moisture and other effects would be excluded from the start, and it would not be impossible to repair the units, given the necessary facilities. However, most such attempts have met with failure. It appears almost impossible to obtain a true seal, very small pin-holes invariably appearing in the solder joints sometimes after several cycles of expansion and contraction due to temperature changes.

The use of casting or potting materials has been put forward in an attempt to overcome these difficulties;

Top view of a complete receiver encased in NEL casting resin.

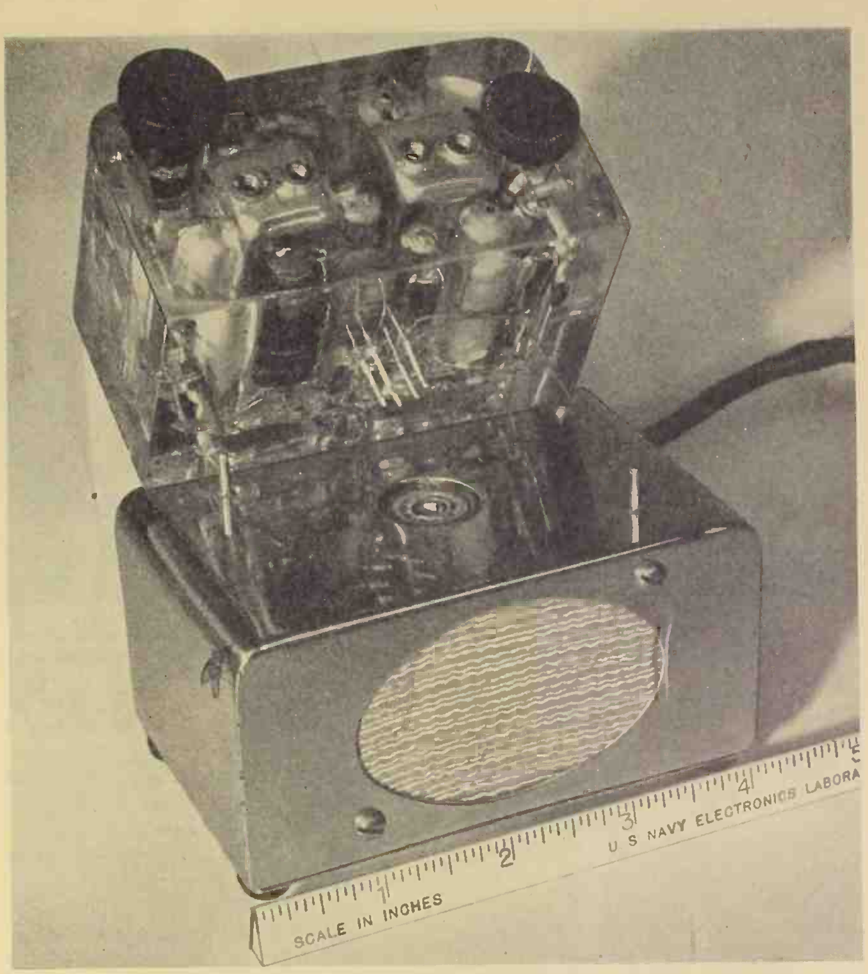


when embedded in the plastic, the electronic components are thoroughly protected from the effects of humidity and fungus. Considerable protection is, of course, also provided from mechanical damage due to handling; and the effects of vibration in particular are reduced because of the damping produced by the plastic surrounding the components.

There are other problems which appear, however, when potted circuits are used. For one thing, it is obviously impossible to repair a defective unit in most cases. This increases the maintenance problem, when it is considered in the conventional light. However, modern electronic equipment is becoming so complex that it is very difficult to obtain qualified Naval personnel for maintenance. It is desirable therefore to increase the ease of repair of the equipment itself. This may be done by designing units in small subassemblies which plug into the main assembly. In this way, when defects appear they may be remedied in service rapidly and effectively by replacing the defective subassembly. Admittedly this is expensive in that it throws away good as well as defective components; but it reduces the training necessary for the service personnel to a considerable extent and makes unnecessary the carrying of large quantities of spare parts, each individually protected from the elements. Instead, a relatively small volume of replacement subassemblies is carried.

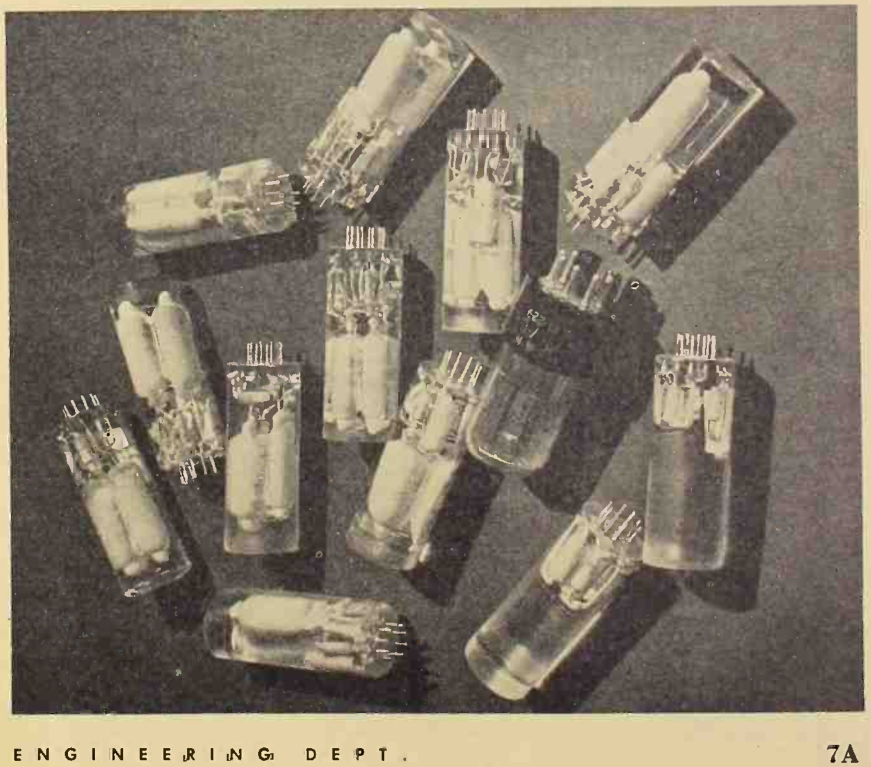
One other effect is of importance in miniaturized circuits. Usually there is just as much power dissipated in the small units as in the equivalent large ones, but there is less surface area to carry away the resultant heat. Operating temperatures are therefore in general much higher than for the larger units, making things difficult for the components involved. Fortunately new and better components are being produced all the time, and it is possible to reduce the temperatures to at least some extent by proper thermal design.

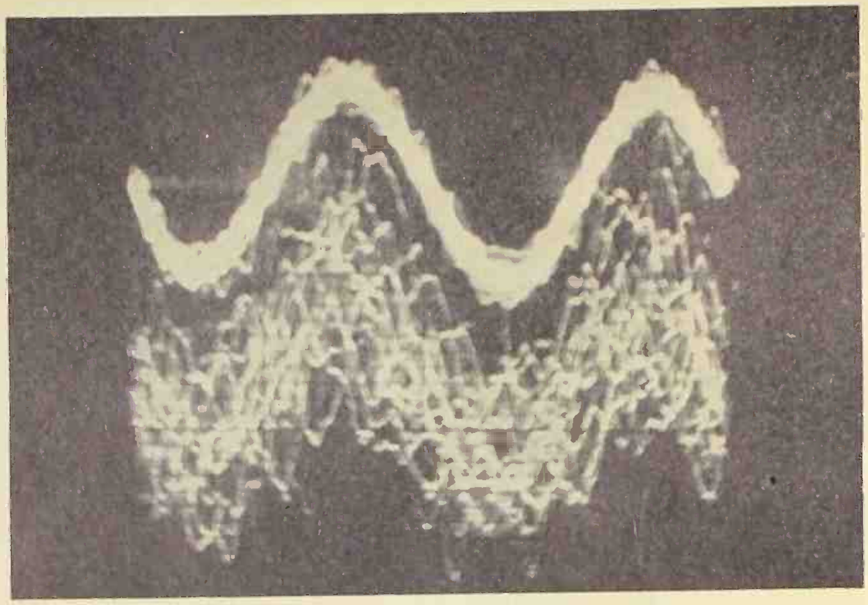
Early in the studies of potted electronic circuits at the Navy Electronics Laboratory it was found that there were no entirely suitable casting materials then available. All which could be obtained had some faults, either electrically or mechanically. The casting resin developed by the National Bureau of Standards, commonly known as NBS Resin, while excellent from the electrical standpoint was unsatisfactory mechanically. Electrically it was equivalent to polystyrene, very good indeed; but it was also otherwise similar to polystyrene, being soft, flowing at a low temperature, and fairly difficult to machine. In addition, it was expensive and the ingredients difficult to obtain.



Another view of the six tube receiver shown on P. 6A, complete with power supply and speaker. Note size of fly on speaker case.

Typical cylindrical plug-in assemblies encased in NEL casting resin. An octal tube is included to indicate size of the castings.





Comparison between two identical amplifiers mounted on vibration table. Top oscilloscope curve shows output of potted unit; bottom, conventional subminiature chassis construction.

Other materials available at that time were usually good, or at least fairly good, mechanically but poor electrically. These no doubt would be suitable for many low frequency applications encountered, but it was felt that something which could be applied to both high and low frequency units was desirable.

The material currently in use at the Navy Electronics Laboratory, designated NEL-177 Casting Resin, while not ideal, is at least usable at frequencies as high as 100 megacycles and has fair mechanical and heat-resisting properties. Dielectric constant and power factor are of the order of 3.0 and 2% respectively at 10 megacycles, and do not depart from these values appreciably at other frequencies. When properly cured, surface temperatures as high as 100 degrees Centigrade can be withstood satisfactorily, and internal temperatures of the order of 200 to 250 degrees are not excessive. Surface checking tends to take place at higher temperatures, and if the internal temperature is too high the material changes in color from almost water-white with a slight amber tint to a dark brown in a relatively short time. When operated at elevated temperatures the color normally changes slightly, gradually darkening.

While the ingredients used in NEL-177 are cheap and readily available, they must be pure and fresh if satisfactory results are to be obtained. The base stock in particular deteriorates with time even when stored under refrigeration and eventually will become unusable for electronic applications. Considerable variation in properties, particularly mechanical, will be experienced with its deterioration. In addition, the manufacture of the base ingredient is not controlled to as fine a point as might be desired. There are therefore small changes from one batch to another which in some cases result in appreciable changes in properties. Fortunately these do not appear to be reflected to any great extent in the electrical properties.

There is nothing at all difficult about the formulation of the material which makes trained personnel necessary. Anyone who can read and weigh out the ingredients accurately can do it. Also there are no critical temperatures or processes involved in the casting and curing of the resin.

Mold requirements are easily met, since the material is poured in the form of a liquid over the circuit to be cast. This makes it possible to make unusual shapes without difficulty. The main requirement of course is that the molds be made of some material that does not react with or adhere to the casting resin. This restricts the choice to a relatively small field, but no difficulty is experienced in finding suitable materials. In some cases lucite has been used. There is no reaction between the lucite and the resin, but the mold must be machined from the casting if it is desirable to remove it. Since lucite is readily formed to shape this is not a great disadvantage, especially if there is no point in removing the form. Polystyrene cannot be used in the same manner, because it is seriously affected by the casting resin. Certain acetates perform satisfactorily as forms, since they neither adhere to nor are attacked in any way by the resin; and they are very flexible in application.

The principal disadvantage of plastic molds is that they are usually either destroyed or otherwise made unsuitable for further use. A type of mold which can be used over and over, which has been found very useful at NEL, consists of chromium-plated metal. The metal is of course not destroyed in removing the casting, and the chromium plating produces a smooth mirror-like surface. Since the plastic does not adhere to the chromium surface if it is clean, the casting is easily removed after curing. This type of mold has been used extensively for manufacturing cylindrical plug-in sub-assemblies, the only machining required on the finished casting being the cutting of the unit to the required length.

Although cured in the mold, there is a certain effect on the surface of the castings, apparently caused by the oxygen of the air. This has been overcome by curing the plastic in such materials as pure linseed or cottonseed oil or glycerine. The resulting surface then has the same properties as the interior of the casting. This is accomplished by removing the casting from the mold after it is sufficiently gelled and completing the curing immersed in the oil.

The change from a liquid to a solid in these resins is due to a chemical process known as polymerization. During this process a certain amount of heat is generated, and this must be controlled if satisfactory castings are to be obtained. If too much heat is generated, the polymerization will take place too rapidly and the resulting casting will in all probability contain large cracks or fissures. If too little heat is produced curing will take too long for practical application. The control of the amount of heat generated is accomplished at NEL in part by refrigeration. After the unit has been poured, it is held at a low temperature for as long as is required for the plastic to thoroughly gel. This is usually five to six hours, but depends upon the size of the casting involved. After the casting has gelled sufficiently it is removed from the refrigerator, taken out of its mold and placed in an oven for several hours, at which time the cure is essentially complete. It is possible to allow the cure to take place at room temperature but considerably more time will be required—unless of course it is summer and room temperature is up around 100 degrees Fahrenheit. In this case no oven is needed since the room is one anyway. In any case, carrying on this latter cure at elevated temperature is best done with the casting immersed in oil as described above. No excessive temperatures are involved in such a process, so no damage is done to components.

The plastic materials generally used as casting resins all shrink from five

(*Continued on page* 27A)

# 3-PROBE METHOD of IMPEDANCE MEASUREMENT

By **EDWIN N. PHILLIPS**
Research Div., Collins Radio Co.

***Derivation of a chart for determining impedance from voltage readings at three fixed probes spaced 1/8 wavelength apart.***

WELL-KNOWN methods of impedance measurement at high frequencies include the use of a slotted coaxial line and the use of directional couplers. Another method involves the use of three voltage probes which are spaced an eighth wavelength apart. On occasion, this method can be of considerable utility: when the frequency is low (with a corresponding long wavelength), or when measurement at a fixed frequency is being performed, this latter method is more easily set up and interpreted than the first two methods.

One difficulty which, possibly, may have inhibited the wider use of this method stems from the fact that the loci of constant voltage-ratios are circular only on a rectilinear map. The chart presented herewith eliminates this difficulty, since these loci are transformed onto a closed chart.

It will be recalled that the voltage at any point $x$ units distant from the load on a lossless line is given by:

$$E_x = E_R \cos\left(\frac{x}{\lambda}360^\circ\right) + jI_RZ_0 \sin\left(\frac{x}{\lambda}360^\circ\right) = E_R\left\{\cos\left(\frac{x}{\lambda}360^\circ\right) + j\frac{Z_0}{Z_R}\sin\left(\frac{x}{\lambda}360^\circ\right)\right\} \text{ volts} \quad (1)$$

where $\lambda$ is the wavelength (the free-space wavelength for principal mode, the guide-wavelength for guided modes) and $E_R$ is the voltage across the load. If, now, a probe at the load reads $E_1$, a probe located an eighth wavelength down the line reads $E_2$, and a probe located a quarter wavelength from the load reads $E_3$, then these voltages are given by:

$$E_1 = E_R \quad (2)$$

$$E_2 = \frac{E_R}{\sqrt{2}}\left(1 + j\frac{Z_0}{Z_R}\right) \quad (3)$$

$$E_3 = j\,Z_0/Z_R \quad (4)$$

From these three quantities, two ratios can be formed:

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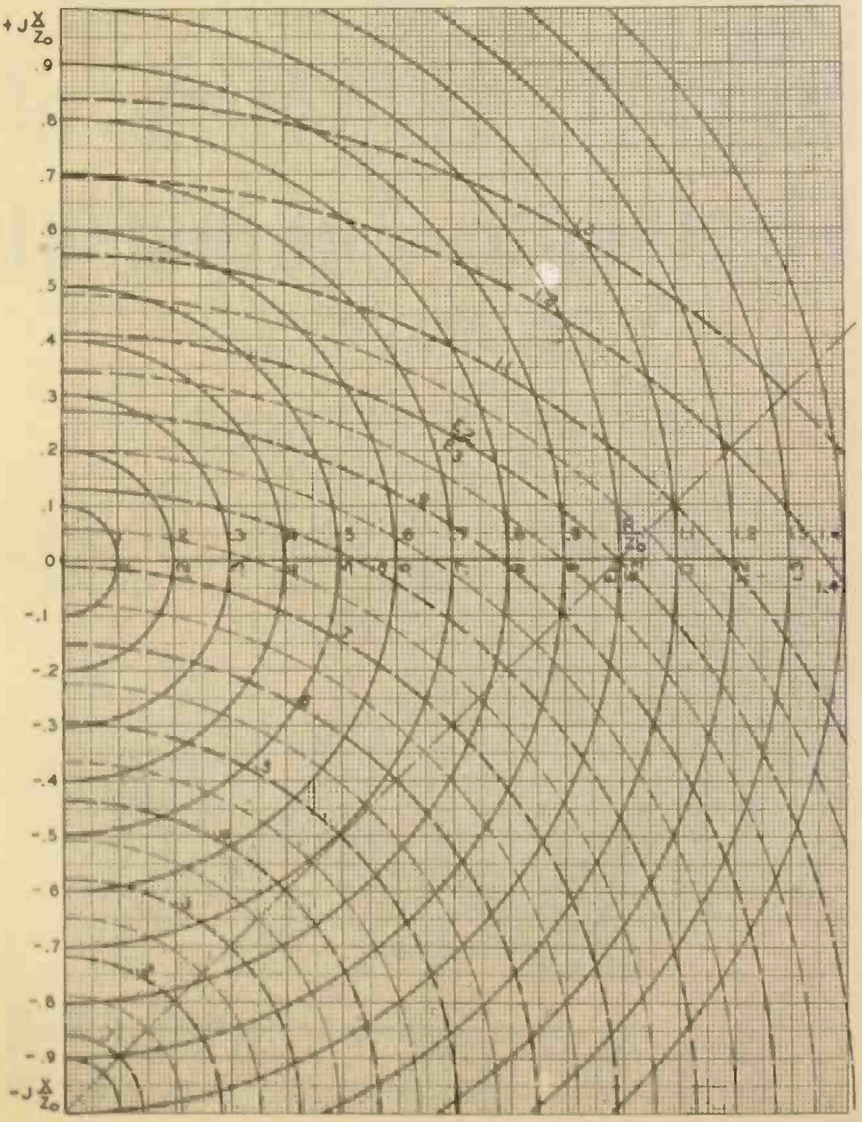


Rectilinear map for use with 3-probe method of impedance measurement.

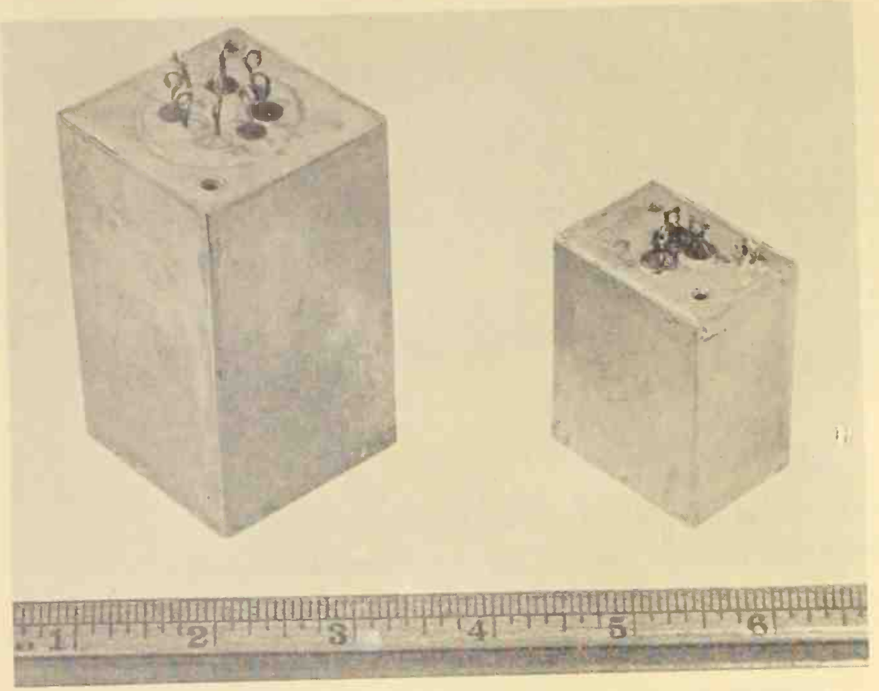
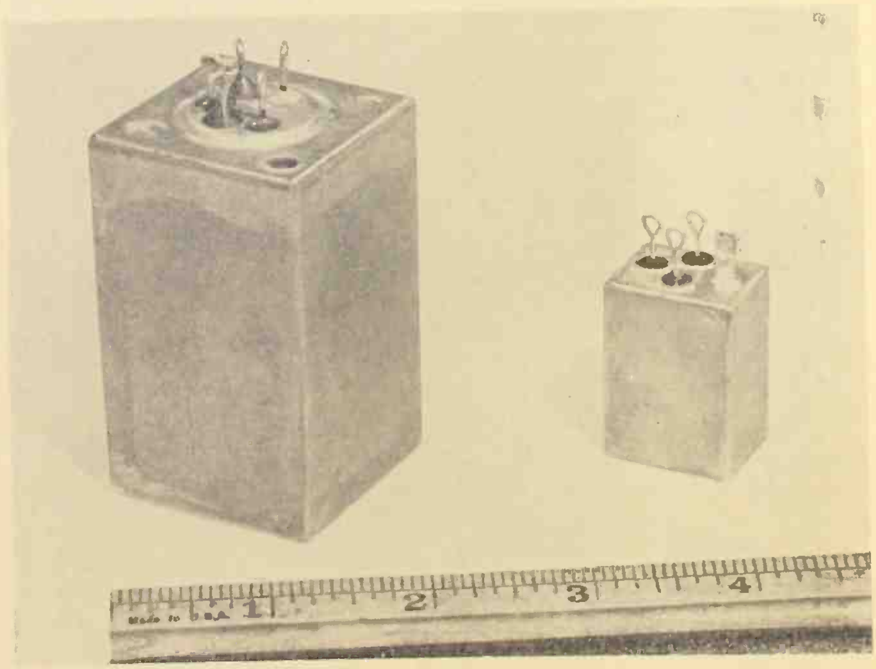


Fig. 1. Comparison between typical 400-cycle 14 VA transformer and miniaturized version, which supplies 33 VA, weighs ¼ as much, and takes up 68% less volume.

# MINIATURIZATION of Electronic Equipment

**By SAM MILBOURNE,** Eng. Dept.
Eclipse-Pioneer Div., Bendix Aviation Corp.

Fig. 2. Conventional excitation transformer (left) compared with miniaturized version, which occupies 1/9 the space and weighs 1/10 as much.



### *Smaller components and improvements in design reduce size and weight.*

A PRACTICAL, if somewhat whimsical, definition of equipment miniaturization is that it is the design process wherein the size and weight are progressively decreased while the use is progressively increased until the resulting equipment takes up no space, weighs nothing but does everything.

Miniaturization of electronic circuits and components—for both military and non-military use—is demanding the greater interest and attention of design engineers. Rapid strides have been made in lightness and compactness—particularly in modern airborne electronic equipment. Fig. 3 illustrates a recently developed plug-in assembly which is part of a complete airborne equipment. Fig. 4 shows the underside. The chassis or "card" is of plated aluminum and measures 5" x 12" x 2¾" over-all. The complete assembly weighs only one pound thirteen ounces! The assembly comprises five two-stage push-pull output amplifiers with a total of fifteen tubes. Small parts such as resistors and condensers are connected between glass-bead through-type terminals which are soldered into the card. Besides the fifteen tubes, the card carries 5 output transformers, 30 resistors, 32 condensers and two plugs—a total of 84 components!

Improvements in non-military electronic equipment are often the outgrowth of previous military equipment improvement. Present television progress can be traced in no small measure to military radar developments. Thus, it can be expected that future non-military miniaturization will be greatly aided by developments in military equipment miniaturization.

It should be remembered that the requirements for military electronic equipment are much more rigid than those for non-military use. Military airborne electronic equipment design is pointed toward a maximum of accuracy, dependability and ruggedness. Extremes of temperature, humidity and barometric pressure as well as resistance to fungus, salt-water, chemicals, vibration, etc., all add their problems. Any miniaturization must be accomplished without serious sacrifice of the above.

Specifically, the general requirements for military airborne electronic equipment can be stated in terms of "Maximum" and "Minimum" as:

*Maximum*

1. Operational life.
2. Safety in use and in the event of equipment failure.

Fig. 3. A plug-in assembly using miniature tubes and parts.

3. Ambient temperature range.
4. External pressure (altitude) range.
5. Resistance to water immersion and salt spray.
6. Ability to withstand shock and vibration.
7. Resistance to chemicals.
8. Resistance to fungus.
9. Ease of service with minimum technical knowledge.
10. Electrical and mechanical efficiency.
11. Dissipation of developed heat.
12. Use.

*Minimum*

1. Size.
2. Weight.
3. Lost space.
4. Internal heat rise compatible with minimum size.
5. Number and types of power supplies.
6. Number and types of parts.
7. Size and weight of individual parts.

Although there may be other requirements in certain applications, the above breakdown presents the designer with his major problems.

Now, how can equipment be made smaller and lighter?

To accomplish this, our first approach was to eliminate the unnecessary in circuits and parts. The design was pared to an absolute minimum of parts. Every resistor and condenser having no real purpose was eliminated. Resistance-condenser coupling between stages was used in place of transformer coupling where possible. Autotransformers and tuned chokes were utilized where practical. One master power supply was designed in place of several smaller ones.

Next, we attacked the problem of reducing the size and weight of component parts, starting with the largest and heaviest parts and working down to the last nut and screw. The success of such a program was positive. A twenty per-cent reduction in the final weight of the complete system under the weight requirements of the development specification was accomplished. It can be stated that in terms of previously designed comparable equipment this represents more than a fifty per-cent reduction in weight.

Transformers and chokes (power and audio) exhibit the greatest weight density per cubic inch of all electronic parts. Previous design experience indicated an average weight of one ounce per VA (approximate) for hermetically-sealed transformers and chokes. This type of airborne transformer is designed with ordinary silicone steel "E-I" laminations, potted in compound and housed in copper or brass cans.

To illustrate what can be done to reduce size and weight in this direction, Fig. 1 shows a typical 400-cycle excitation (power) transformer operating from 115 volts and supplying 14 VA. This transformer measures 3" x $1\frac{3}{4}$" x $1\frac{3}{4}$" over-all and weighs 12 ounces. Also shown is the resulting comparable miniaturized version. Supplying over twice the VA (33 VA), it measures $1\frac{7}{8}$" x $1\frac{9}{16}$" x $1\frac{1}{8}$" over-all, takes up 68% less volume and weighs but 3 oz.!

Acknowledgment by the electronic engineer must be made, not only to the mechanical engineer, but to the chemical engineer. It is through the use of new products and new processes that space and weight reductions can be accomplished. The best results can be achieved by a welding of the knowledge and efforts of these men.

Specifically, how was this size and weight reduction accomplished in the above mentioned transformer?

It was felt that sometimes unreasonable safety factors crept into the design of airborne transformers, particularly when the design passed through several hands—each adding a bit to the over-all safety factor. It was also felt that too little attention was sometimes paid to absolute limits of safe operation. Too many "rules of thumb" seemed to be used. Too little attention seemed to be paid to the requirements of each specific design.

In the design of this transformer, we first studied the core. A toroidally oriented-silicone prefabricated core was finally chosen because a 15% to 30% weight reduction was indicated over standard "E-I" laminations—without altering transformer characteristics or increasing core losses. More rigid control of magnetic characteristics, faster assembly and no high temperature de-

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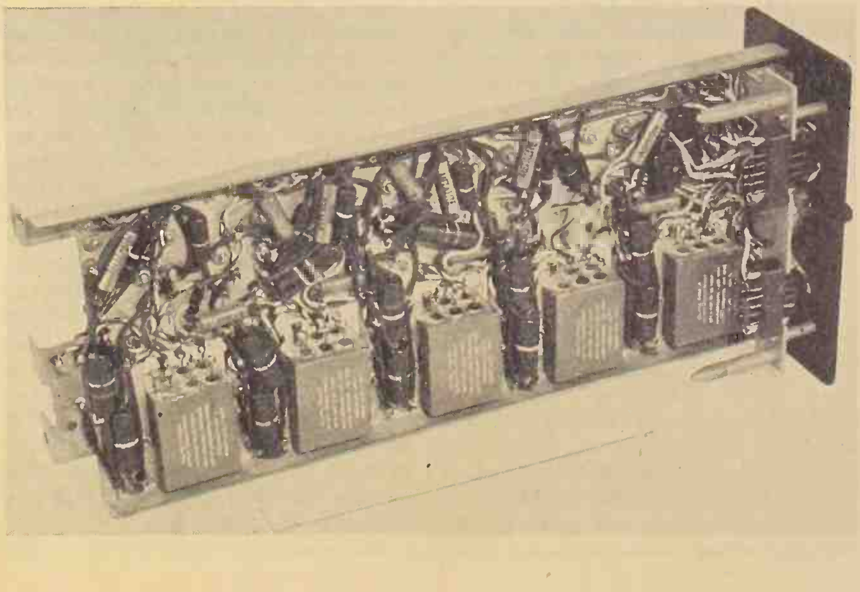
Fig. 4. Bottom view of assembly shown in Fig. 3. Five driver amplifiers are included—a total of 15 tubes, 5 output transformers, 30 resistors and 32 condensers.



By **ALVIN B. KAUFMAN**

***This useful device converts very small d.c. voltages to a.c. so they can be more readily amplified and measured.***

Experimental setup for testing the chopper with an iron-constantan thermocouple.

# Electronic D.C. Millivolt Chopper

THE amplification of small d.c. potentials is an old problem. That this problem is of much importance may be seen by examining the literature on methods of d.c. amplification.

Application of d.c. amplification systems has been made for strain gauge signals, thermoelectric potentials, electrometer output potentials and in medical, industrial, and experimental fields.

Amplification systems are generally of one of the following types:

1. Mechanical-electrical chopper (vibrator) and a.c. amplifier.
2. D.c. Amplifier.
3. D.c. Modulating R.F. Carrier System.[1]
4. Magnetic Amplifier.
5. Electronic Chopper (Bridge Type); a) Neon, b) Resistive, c) Tube, d) Varistor.

Most of these types have been discussed quite liberally in engineering literature and there is little need of recapitulation of their problems of drift, complexities and general unsatisfactoriness.

The "Electronic Chopper" discussed herein is not presented as a cure-all, but a substantial simplification of the d.c. amplification problem as applicable to many laboratory and industrial problems.

The very use of the word chopper signifies the use of an associated a.c. amplifier and detector. As the chopper may be referenced to the amplifier output voltage, a "phase sensitive system" or a conventional carrier detection system may be employed. Unlike the mechanical chopper system of a vibrator or carbon button hummer, the bridge modulator, to be described, may be operated at frequencies as high as five or six megacycles allowing full dynamic recording of d.c. potential variations up to signal intelligence of 600 kilocycles. This feature also allows simplified filtering of the rectified a.c. or carrier signal to produce smooth d.c. output, where required. Of course industrial line frequencies of 60 cycles, or aircraft frequencies of 400 cycles may be used to operate the "chopper" and regulate its frequency of operation.

The electronic chopper utilizes two basic principles. The first is that if a signal is applied to the grid of an amplifier tube through a series resistance and the grid is intermittently shorted to the cathode, the d.c. potential on the grid will vary from zero to its full value. This is roughly the action of the electronic chopper. The electronic chopper does not create a dead short from grid to ground, but does effectively create a varying resistance, which in combination with the series resistance causes voltage divider action and a variation of the d.c. potential impressed on the tube grid. The method of impressing this varying resistance between grid and cathode is the second basic innovation. It is well known that a bridge with four equal legs will have zero voltage difference between its conjungate output terminals. Therefore a bridge circuit is used. Its "output" terminals are connected between grid and ground and a.c. (the carrier frequency) applied to the source terminals. The bridge output terminals must not develop any output voltage, but must vary in resistive value at the carrier frequency rate. The bridge elements, then, are the major consideration.

The bridge modulator may consist of any of a number of entirely different elements, operating technically different, but producing the same end result to varying degrees. Four basic types with several modifications will be discussed.

The resistive bridge modulator consists of four equal resistive legs of fine wire with a high coefficient of thermal resistance change. With each half-cycle of carrier current applied to the bridge, all four legs heat up and increase in resistance. Thus the modulator output is at twice carrier frequency. This modulator is limited by the thermal characteristics of the wire, and the heating and cooling resistive changes of all four legs. The resistance change available without burning out the wire is the major limitation of output, while the dynamic resistance tracking of all four legs on their heating and cooling cycle regulates the undesirable output of bridge carrier voltage and this regulates in turn the value of the lowest possible d.c. signal that may be amplified. If all four legs do not remain equal in value, then the bridge supply will cause an output voltage to appear on the grid of the tube without a d.c. input signal and this consequently regulates the signal-to-carrier (or noise) ratio, as indicated. It is this ratio which limits the value of the d.c. potential which may be amplified. These factors limit the use of the resistance bridge modulator and it is not too satisfactory, but may be improved with research.

The neon bridge modulator has a much better resistive output variation, possessing as its does a variation between almost zero and infinity ohms. With no potential (or carrier) applied to its bridge, the neon lamps are unlit and at the low d.c. signal potential applied to them will not ionize and therefore present a high impedance to the tube grid. As each half-cycle of carrier is applied to the bridge the neon bulbs ionize (or light) and form a low impedance path from grid to cathode. Any resistive unbalance between neon lamps is balanced out by the slide wire potentiometer forming the other half of the bridge. On the non-conducting portion of the carrier cycle, the bridge is balanced capacitively by the padders. It is apparent that this modulator also puts out a frequency twice the carrier frequency.

The neon modulator works quite well and its limitations are due mainly to the dynamic resistance characteristics of the neon lamps, firing point and inverse resistance characteristics. The lowest d.c. potential which may be amplified satisfactorily because of these limitations is approximately 50 millivolts, varying with the modulator components and associated equipment. Another variation of the neon modulator is the replacement of the neon bulbs with vacuum tube diodes. The system faces similar problems of capacity balance and matching of dynamic characteristics, along with not too satisfactory a variation in impedance and Edison effects causing performance not to be any more satisfactory than the neon modulator.

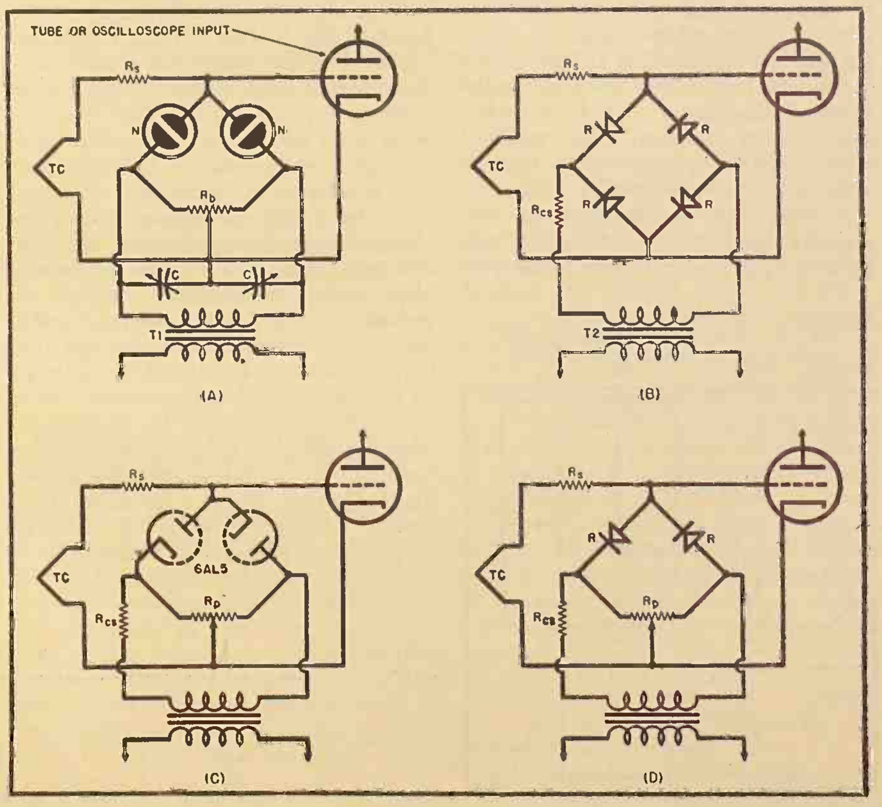
Possibly the best bridge modulator is that used in telephone and radio work as described by Terman, "non-linear rectifiers" in his Handbook of Radio Engineering. This modulator is commercially described as a "Bridge Modulator", "Copper Oxide Modulator", and under varying other synonyms. Here four non-linear rectifiers consisting either of copper oxide, selenium, or germanium are wired together identical to that circuit used in a full wave bridge instrument rectifier. However the lead away connections employed are entirely different. A.c. or the carrier is applied to what would normally be the plus and minus output connections, and what would be the input a.c. connections are tied grid to cathode.

As with the three other modulators the dynamic and static resistance characteristics of the four non-linear rectifiers which form the four legs of the bridge, limit the useful input signal level. The theory of operation is simple. The forward resistance of such rectifiers is very high at extremely low current, dropping off to a low value at full current rating. The d.c. signal potential is insufficient to cause appreciable current flow and therefore the bridge is effectively an open circuit. With application of the carrier signal, full current flows through the rectifiers on the conducting half of the cycle and the bridge output resistance (grid to cathode) falls to a low value. As the bridge only conducts every other half cycle, the amplifier will effectively produce the same output frequency as the carrier supply. Of course there is a certain shunting effect from the impedance of the carrier exciting transformer, particularly with the full bridge circuit, which may be limited however by the addition of a series resistor of high value in one of the carrier exciting leads. The carrier series resistor should be five to ten times higher than the value of the series grid limiting resistor.

For high efficiency the signal-to-carrier voltage ratio should be ten or higher. Actual values will depend on the dynamic characteristics of the rectifiers used and the d.c. signal potential. A carrier current value should be selected that gives best signal to noise (or carrier) output.

The interelectrode capacity of such rectifiers is usually insignificant at carrier frequencies up to several thou-

Fig. 1. Circuit diagrams of several versions of the d.c. millivolt chopper.



sand cycles, balancing possibly being required at five or six megacycles. Load impedances are generally of little importance except at the higher frequencies where the shunting effect of the rectifier capacities may become appreciable.

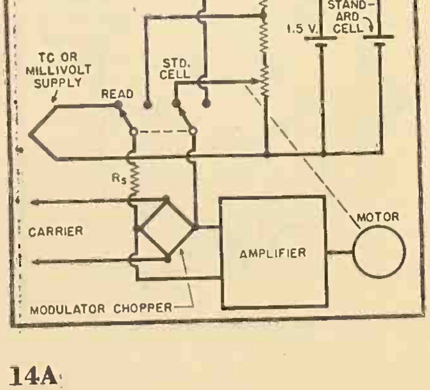
The inverse resistance characteristics of the rectifiers if dissimilar may cause an undesirable output on the non-conducting half of the carrier cycle. As indicated before, static matching is not sufficient; the dynamic, static, and inverse resistance characteristics must be the same.

Tests were conducted by the author using a *Conant* instrument rectifier Model 160 B and a *Sylvania* Varistor type 1N42. The instrument rectifier was not intended for this service. The *Sylvania Electric Company* Varistor, consisting of four matched germanium diodes, is specifically designed, among other uses, for use as a bridge modulator.

Considering the *Conant* instrument rectifier first, wired exactly as Fig. 1B, good readable signals from 10 millivolts d.c. could be observed on an oscilloscope with its gain adjusted for fairly straight line output with no d.c. input signal. Smaller signals could not be read above the noise (unbalance carrier) level. The observed signal was, of course, at carrier frequency. 60 cycle carrier was used. This rectifier is rated at 5 volts r.m.s. input, maximum current 5 milliamperes. Possibly twice this rated current could be run on the half duty cycle as used in this configuration, if necessary. The series grid limiting resistor finally selected was 300 ohms, but was not critical. The carrier supply across the bridge was 3.3 r.m.s. volts, supplied through a half-watt 10,000 ohm carrier resistor. The millivolt source was an iron-constantan thermocouple of low resistance.

A number of factors must be considered; the impedance of the d.c. signal source, the correct value of the grid or series limiting resistor, optimum impedance matching, and the optimum carrier voltage. Some of these questions may be answered quantitatively while others must be qualitative, depending largely upon components and amplification equipment.

Fig. 2. Simplified schematic diagram of an automatic potentiometer.



The impedance of the d.c. signal source should preferably be small in comparison with the input impedance of the amplifier with its modulator and series limiting resistor, or excessive shunting of the signal source impedance will occur with a corresponding reduction of the d.c. signal potential. The input impedance of the modulator and amplifier should be its highest possible value, *i.e.*, with the modulator not conducting, when used for calculations. In the conducting state we are trying to reduce the signal as much as possible and the shunting is not serious.

The series limiting resistor value may best be determined empirically, approximations being made by formula. Part of this problem hinges on the fact that in determining by test the forward or inverse impedances of these rectifiers, the analyzer current must be of the same value as the operating carrier current or erroneous data will be secured. Also, this operating carrier current in itself will depend somewhat on the value of d.c. signal and the value of the series limiting resistor. As much carrier current should be used as will allow substantial dynamic tracking of all four rectifiers, generally less than one-half rated current of the rectifier. This must be determined empirically with an oscilloscope. The carrier current should be adjusted to the value where maximum change of output signal occurs for a given d.c. input signal. This may in some cases result in a background level or signal which may be balanced out in the detector or output circuit.

If the bridge modulator effectively changed resistance between infinity and zero ohms, then no series resistance would be necessary, except to prevent loading of the signal supply source or excessive bridge current for high values of d.c. signal potentials. In practice this optimum condition does not exist. The fact that the modulator does not open to infinity ohms is the main problem in determining the series limiting value. With a very low impedance signal source, if no limiting resistor were used, the d.c. signal could not be attenuated sufficiently because the modulator resistance does not go to zero ohms. Thus to reiterate, to prevent loading or reaction on the signal source and limiting by the action of the modulator a series limiting resistor is generally necessary. Optimum series resistance value for low values of d.c. input potentials is approximately: $R$ series $= R_{nc} - R_c$ of the bridge, where $R_{nc}$ and $R_c$ are respectively the nonconducting and conducting grid-to-cathode impedance of the bridge. $R_c$ depends upon the carrier voltage and consequential rectifier currents. This series resistance may vary widely in value with variations of d.c. input signal and input impedance. High values of d.c. input potential require the use of a higher series grid resistor than normal to maintain the ratio between carrier and signal currents.

Where intermittent loading of the signal source, at carrier frequency, is permissible it is often possible to secure much higher output by eliminating the series limiting resistor. This depends, however, on three factors: the input impedance, the d.c. input level, and the characteristics of the particular modulator used. This is satisfactory only where: $R$ *source* $> R_c$ *modulator*.

A half-bridge modulator may also be used quite effectively, in which case only two rectifiers are used and these need not have identical static resistance values, as this may be balanced out by the potentiometer forming the other half of the bridge. Dynamic slope characteristics must still be close. Under these conditions, although possessing less available change in resistance than with the full bridge, it is sometimes possible to use more of the resistance change available because of decreased matching problems and thus secure higher outputs for a given d.c. input signal. The potentiometer forming the other half of the bridge should have a resistance of: $R_p \gtreqless R_c + R_c$ where $R_c$ is the conducting resistance of the rectifiers at the rated carrier current. Optimumly, the potentiometer should be of as low a value as consistent with reasonable carrier current through its half of the bridge.

The *Sylvania* Varistor bridge modulator (shown in photograph) performed excellently and is to be preferred over the other units because of its availability as matched units, making hunting for individual matched components unnecessary. This would be especially important for commercial equipment.

As with the other circuits, the millivolt d.c. supply consisted of an iron-constantan thermocouple with about five ohms lead resistance. This supply was chosen as a convenient standard signal source for comparsion of chopper efficiency. The thermocouple (about 28 gauge wire) was heated to cherry red with a match, supplying about 40-50 mv. peak. $R_s$ was empirically determined optimum at 500 to 1000 ohms when $R_c$. was 10,000 ohms. With 60 cycles as the carrier, optimum carrier voltage was 6-10 volts r.m.s. across the

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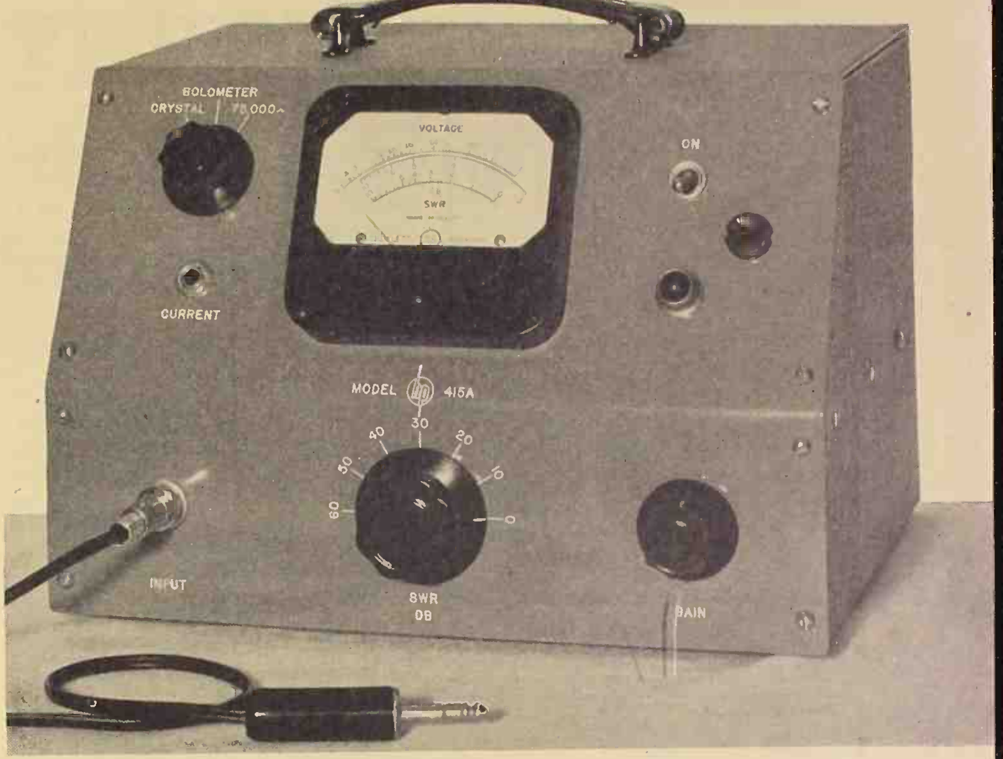


Power meter for measuring standard wave ratios on slotted line. The r.f. is modulated at 1000 cycles and is detected, amplified, and measured by this instrument.

# MICROWAVE MEASUREMENTS

## *Part I. Equipment and techniques for measuring impedance and power at microwave frequencies.*

By
**J. RACKER**
**Federal Telecommunication Labs.**

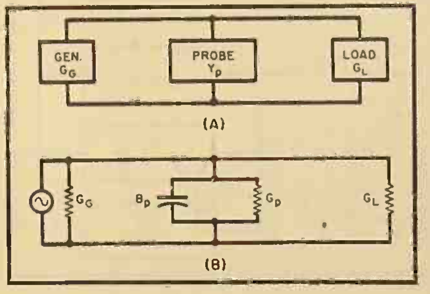
THE design principles described in the previous articles in this series provide the engineer with the basic approach to a given microwave problem. Once this approach is established the details of design, *i.e.*, configuration, size, and position of the various elements, will be determined primarily on an empirical trial and error basis. This procedure is employed in many other fields, but it is particularly necessary in microwave techniques because of the mathematical complexity of even relatively simple circuit calculations. For example, a cavity filter is desired between antenna and mixer in a receiver. The exact position of the input and output probes must be determined to effect proper matching between antenna and filter, and filter and mixer. The mathematical solution to this problem involves very complex computations and is completely impractical, while it is a relatively simple matter to adjust probes for maximum power output and minimum standing-wave ratio.

Several years ago the measuring equipment necessary to effect such a trial and error procedure was not readily available and the engineer frequently had to design and build these instruments himself. Sometimes this was the most difficult part of the project. Today, due to the expanded use of microwave systems, a wide selection of commercial equipment is on the market. In this article, the author will be primarily concerned with the theory of operation and practical application of these units.

Not only are the techniques themselves different, but the electrical characteristics of interest in the microwave region are, in several instances, not the same as those of interest at longer wavelengths. For example, we are more concerned with the electric field than in potential difference; in fact, voltages are difficult to define in most microwave elements other than coaxial lines. For this reason, the output of an oscillator or signal generator is specified in terms of power delivered to a load matched to the transmission line rather than in terms of available voltage across a given impedance. Similarly, instead of inductance or capacitance, measurements are usually made of the normalized impedance of unknown loads.

In previous articles[1,2] it was shown that when a transmission line was terminated in an impedance equal to the line characteristic impedance, no reflections from the load occurred. If the load impedance was not equal to the line characteristic impedance, part of the energy would be reflected back down the line and standing waves would be developed. The magnitude of these standing waves would be proportional to the impedance mismatch ratio. The position of the maxima and minima along the line would be a function of the phase difference between the two impedances. Thus it is seen that knowing the impedance of the line, the standing-wave ratio, and position of the maxima and minima points, it would be possible to determine the impedance of a given load.

Fig. 1. Equivalent circuit for probe in (A) block diagram form and (B) schematically with probe represented by $G_p + jB_p$.



The impedance $Z_n$ is known as the normalized impedance and is equal to the actual impedance, $Z$, divided by the characteristic impedance of the line, $Z_o$. For most transmission line matching applications (particularly in wave guides) the normalized impedance is a more useful parameter than the actual impedance. The normalized impedance of the load is related to the reflection coefficient, $K$, by the following expression:

$$Z_n = \frac{1 + K}{1 - K} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

The reflection coefficient, $K$, may be a complex number having both magnitude

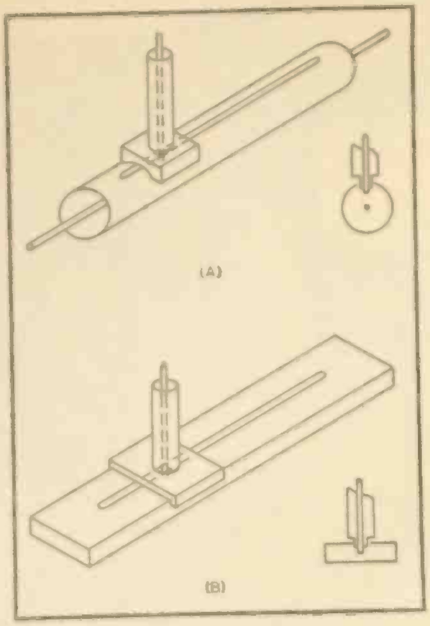


Fig. 2. Slotted line with probe for (A) coaxial and (B) wave guide line.

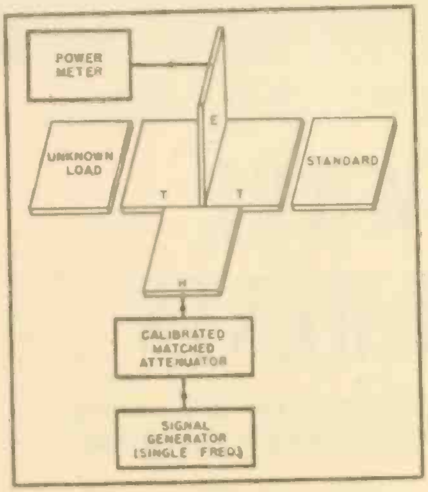


Fig. 3. Single frequency "magic T" impedance measuring bridge.

and phase components. The magnitude of $K$ can be determined empirically by measuring the standing-wave ratio (power) and using the following equation:

$$|K| = \frac{\sqrt{\rho} - 1}{\sqrt{\rho} + 1} \quad (2)$$

The phase of $K$ is determined by the minima on the line, being equal to zero at these points. (Maxima points correspond to a phase angle of 90°). Knowing the phase at any one point, $N$, it is possible to determine the phase at point, $M$, by the following expression:

$$K_N = K_M e^{j2\beta x} \quad (3)$$

where $x$, the distance between the two points, will be positive when $M$ is nearer to the generator, and negative when $N$ is nearer to the generator. $\beta$ is equal to $2\pi/\lambda$, where $\lambda$ is the signal wavelength in the transmission line.

Standing waves are measured through the use of slotted lines which consist of a section of transmission line into which a small probe is introduced through the slot. Fig. 2 shows two types of slotted lines, *i.e.*, coaxial and wave guide. The probe extracts a small fraction of the power flowing through the line and this power is measured in an external circuit. (The external circuit will be described later in this article when power measurements will be discussed).

By moving the probe along the slot, the standing waves can be determined. In order to minimize distortion of the field configuration, the slot should run parallel to the lines of surface current. In coaxial lines, due to symmetry the slot can be placed anywhere parallel to the axis, but in the $TE_{1,0}$ wave guide the slot should be placed at the center of each of the broad sides. It is very important to maintain the width and depth of the slot constant and its direction parallel to the axis. If the probe insertion varies by as much as 0.001 inch as the carriage is moved along the slot an error of several per-cent in the measured value of standing-wave ratio may be introduced. This imposes very close mechanical tolerances in the construction of the line.

Assuming proper construction and reflection-free matching between line and load, there are two factors that may require consideration in the evaluation of the results. One is the standing wave introduced by the slot and the other is the standing waves caused by the presence of the probe. It can be shown that the characteristic impedance of a coaxial line varies due to the insertion of the slot by approximately:[5]

$$\frac{\Delta Z_0}{Z_0} = \frac{1}{4\pi^2} \frac{w^2}{D^2 d^2} \quad (4)$$

where $\Delta Z_0$ is the change in original characteristic impedance, $w$ is the slot width, and $D$ and $d$ are the radii of the outer and inner conductors. The reflection coefficient $K$ will therefore be equal to:

$$K = \frac{\Delta Z_0}{2 Z_0} \quad (5)$$

For very precise work this effect may be troublesome. It may be avoided by compensating for the change in impedance caused by the slot by increasing the diameter of the inner conductor.

In a wave guide the impedance due to the presence of a slot is approximately equal to:[6]

$$Z_s = Z_0 \left(1 + \frac{w^2 \lambda_g^2}{8 \pi b a^3}\right) \quad (6)$$

where $Z_0$ is original guide impedance.

The presence of the slot also affects the propagation constant of the wave guide, and increases the guide wavelength by the same factor as the impedance. This factor is not too important, however, since (as will be indicated later) it is recommended that the wavelength of the guide be determined by actual measurement.

### Probe Effects

An ideal probe would be one whose presence in no way altered the field within the transmission line and which, nevertheless, provided an indication of the intensity of the electric field within the line. This ideal is not attainable, of course, and usually some compromise between probe sensitivity and mismatch effects must be made.

For most practical applications where the probe dimension parallel to the slot is small compared to a wavelength, the probe can be represented as a shunt admittance across the line as shown in Fig. 1A and B. The impedances shown in this diagram are normalised so that $Y = 1$ represents an admittance equal to $1/Z_0$.

Assuming that the generator is matched to the line and that $B_s$ can be made to be equal to zero by tuning it out (both assumptions are valid for most applications) then $G_L = 1$, the ratio between standing-wave ratios with and without probe can be given by:

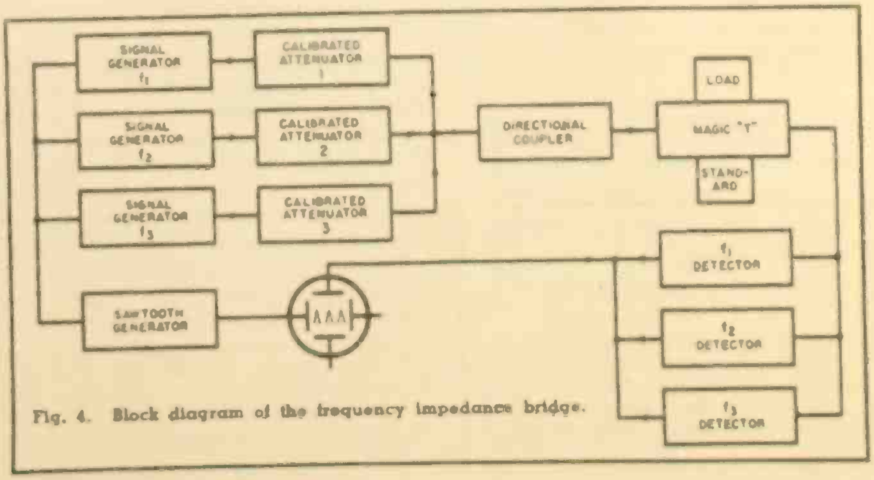


Fig. 4. Block diagram of the frequency impedance bridge.

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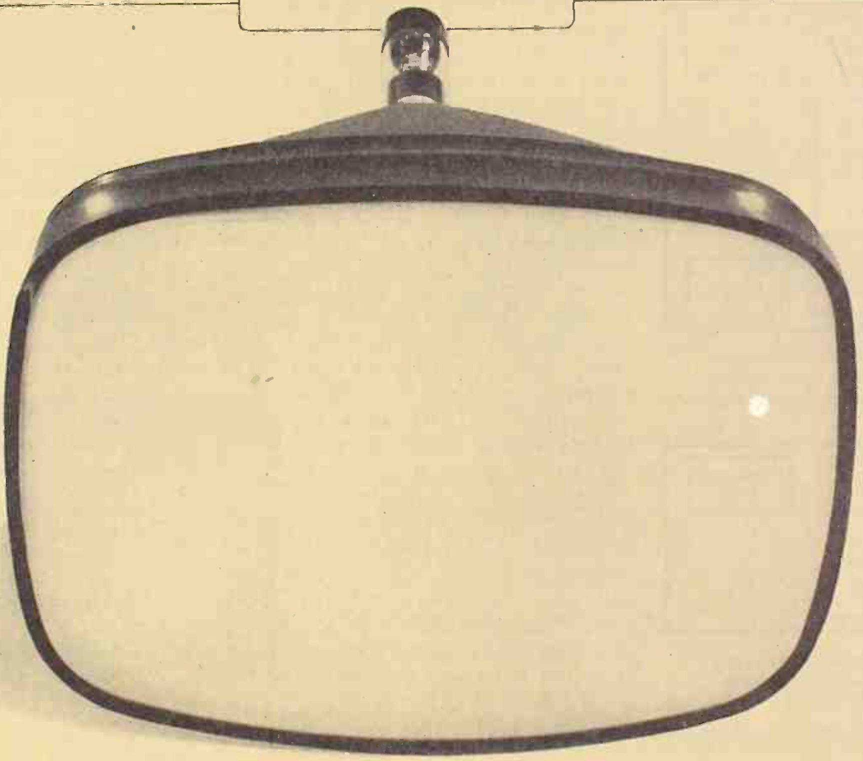


This unique combination of experience naturally fits Sylvania for top position in the TV Picture Tube field.

Maintaining this leadership is a continuing program of research and engineering. A Sylvania engineer, for example, invented the famous "Ion Trap," now licensed to numerous other picture tube makers.

Sylvania achievements in fluorescent powders, tungsten wire, and precision parts are some of the other reasons which lie behind the consistent color, greater clarity, and longer life of all Sylvania TV Picture Tubes.

Backing up each Sylvania advance is a rigid system of quality control . . . of checking and rechecking every step of every process . . . so that TV set owners everywhere will continue to look to Sylvania for the finest performance possible. New booklet gives information concerning the complete line of Sylvania Picture Tubes. Write for your copy today. Address Sylvania Electric Products Inc., Dept. R-2312, Emporium, Pa.

# SYLVANIA ELECTRIC

TELEVISION PICTURE TUBES; RADIO TUBES; ELECTRONIC PRODUCTS; ELECTRONIC TEST EQUIPMENT; FLUORESCENT TUBES, FIXTURES, SIGN TUBING, WIRING DEVICES; LIGHT BULBS, PHOTOLAMPS; TELEVISION SETS

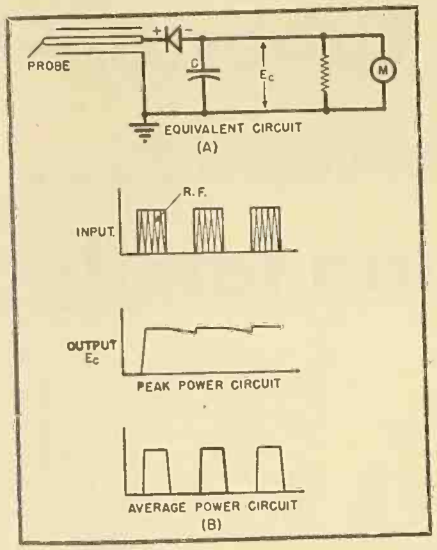


**Fig. 5.** Crystal detector power meter.

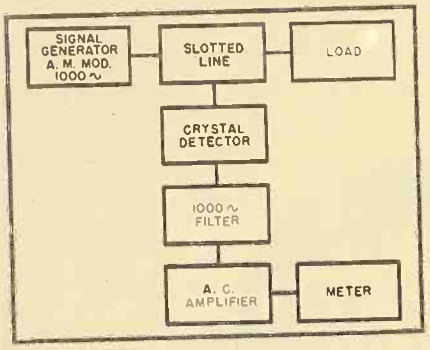


**Fig. 6.** Power meter (with slotted line) employing 1000 cycle modulation of r.f. to effect sensitivity.

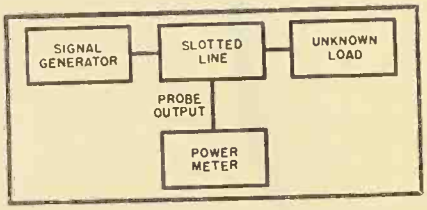


**Fig. 7.** Block diagram of typical impedance measuring setup.

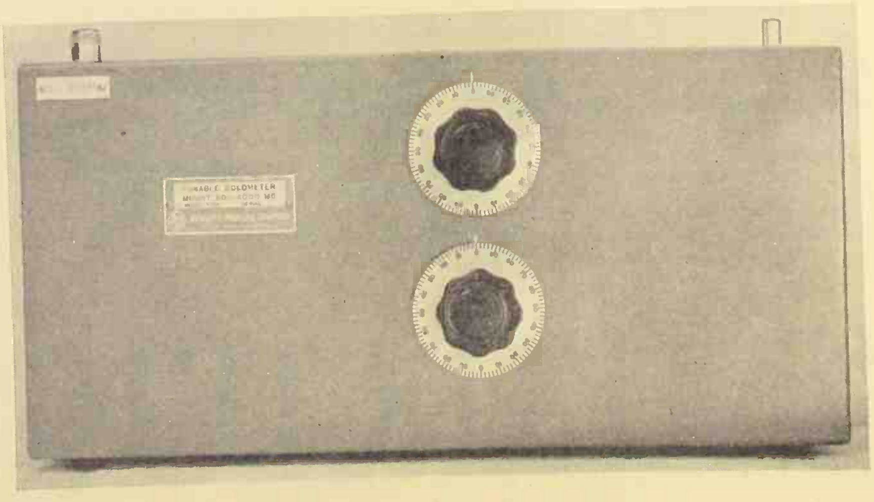
$$\frac{P_s}{P} = \frac{1+P+G_p}{1+P+PG_p} \qquad \ldots\ldots \quad (7)$$

The quantity $G_p$ determines the amount of power extracted by the probe, and is called the coupling coefficient of the probe. As previously indicated, every standing-wave measurement poses the problem of compromise between a very small coupling coefficient necessitating relatively elaborate external equipment to obtain sensitivity, or extracting more power by increasing $G_p$ and thereby encountering a larger discrepancy between measured and true readings.

Fig. 7 shows a typical impedance measuring test setup using a slotted line. To determine the impedance of the load it is necessary to measure the standing-wave ratio and distance from the first minimum to the load impedance. Equations (1) and (2) are then used to calculate the load impedance. It should be noted that Eqt. (2) is expressed in terms of power standing-wave ratio. In some instruments voltage standing-wave ratios, $\rho_v$, are measured, in which case the substitution $\rho = \rho_v^2$ should be made in Eqt. (2).

The determination of the distance from load to first minimum may be complicated by the fact that for some loads, such as antennas, it is not apparent exactly where the load starts. In these cases it is necessary to calibrate the slotted line in terms of distance from the desired load. The following procedure is used to effect this calibration: the signal generator is first set to a frequency at which the load acts as an open or short circuit. Antennas, for example, have a finite bandwidth beyond which they act as short circuits. With the load presenting a short, the first minimum will occur exactly one-half wavelength away from the load, with the second minimum the same distance ($\lambda/2$) from the first one. By measuring the distance between second and first minima, the load starting point becomes known and the line can be calibrated in terms of distance from this point.

Tunable bolometer mount used for measuring power at microwave frequencies.



## "Magic T" Impedance Measuring Bridge

Slotted lines are used extensively for impedance measurements; however, they do have the disadvantage of being time consuming. In some applications it is desirable to determine the reflection coefficient instantaneously. For these applications a device, somewhat equivalent to the Wheatstone bridge, using a "magic T" wave guide, shown in Fig. 3, is employed. The magic T, as previously described,[1] has the property of dividing the power fed into the H-arm equally between the two test arms if these arms are properly terminated in reflectionless loads. If these arms are not properly terminated, some of the energy reflected from the load will go into the E-plane arm.

For impedance measurements, one test arm is terminated in a standard load which represents a reflectionless match over the desired band. The load of unknown impedance is placed across the other test arm. A signal at the desired frequency is applied to the H-plane arm. If this load does not match the test arm at this frequency, some power will be reflected into the E-plane arm. The magnitude of this power is proportional to the square of the reflection coefficient, $K$. Hence, a detector placed in the E-plane arm can be calibrated in terms of voltage or power standing-wave ratios. It should be noted that the phase of the unknown impedance cannot be measured by this method. However, in many applications, only the value of the reflection coefficient is of interest since the power lost due to mismatch is a function of this parameter only.[2]

The simplest type of magic T device is the single frequency bridge shown in Fig. 3. In this equipment impedance measurements can be made at one frequency only for a given setting of the instrument. To determine the impedance characteristics over the desired band, the frequency must be varied and individual readings noted at each point.

It is possible to measure the reflection coefficient over a band of frequencies instantaneously by feeding several radio frequency signals simultaneously, as shown in Fig. 4, through proper attenuators into the H-plane arm of the magic T. The output of the E-plane arm is fed to an oscilloscope through appropriate frequency selection, amplifying, and detecting apparatus. The radio frequency signals are so spaced in time that they will appear in proper time sequence on the scope. Hence, the scope

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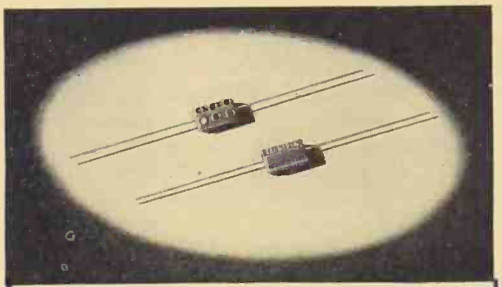

**CM-15 MINIATURE CAPACITOR**

Actual Size 9/32" x 1/2" x 3/16"
For Television, Radio and other Electronic Applications.
2 to 420 mmf. cap. at 500v DCw.
2 to 525 mmf. cap. at 300v DCw.
Temp. Co-efficient ± 50 parts per million per degree C for most capacity values.
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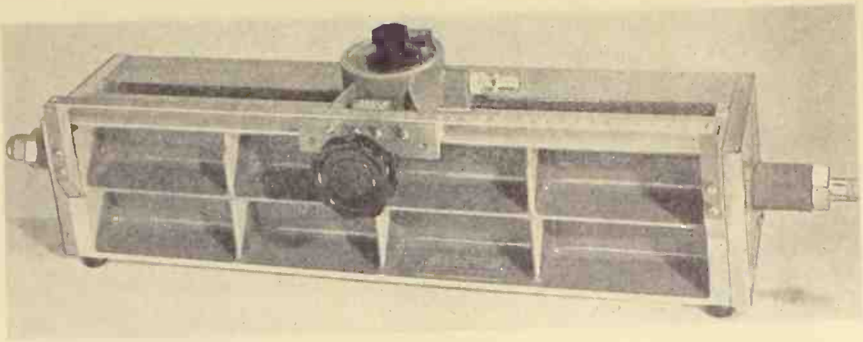


Typical slotted line used for microwave impedance measurements.

will display these signals as a series of pips, the magnitude of each pip being proportional to the reflection coefficient at the frequency involved. Thus the bandwidth response will be noted instantaneously.

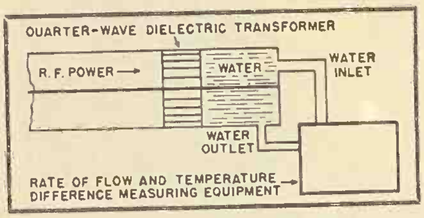


Fig. 8. Coaxial line calorimeter.

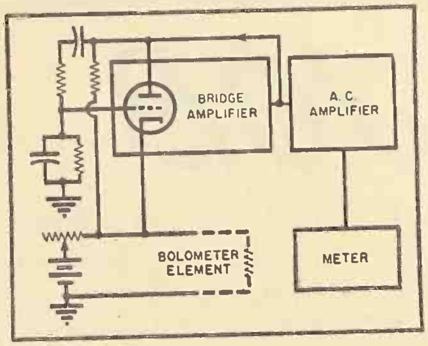
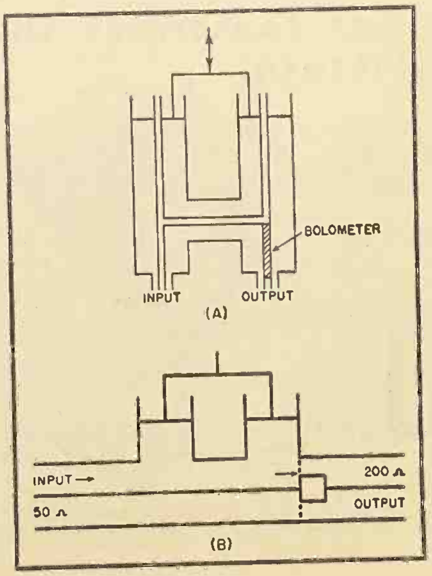


Fig. 9. Basic circuit of a self-balancing bridge power meter.

Fig. 10. Cross-section (A) and equivalent schematic (B) of a tunable bolometer mount.



### Power Measurement

Power measurement, rather than voltage, is of much greater interest at microwave frequencies. There are two basic approaches to the measurement of power. In one case the power is measured directly at microwaves, while in the other the microwave energy is modulated by a low frequency signal and measurement is actually made of the modulation, after the microwave energy is detected and bypassed. This latter method is particularly useful in standing-wave measurements where it is possible to apply any convenient signal to the slotted line. Of course, in checking the output of microwave equipment where convenient modulation cannot be applied, microwave energy must be measured directly.

Microwave power is measured by one of three equipment methods, namely, crystal detectors in which power is measured directly; calorimeters in which power is converted to thermal energy and this energy measured; and bolometers in which power variation is converted, via thermal energy, into a varying resistance, and this variation is measured.

The crystal detector, shown in Fig. 5, is the simplest and least accurate method of measurement. The r.f. power is rectified by the crystal and used to charge up condenser *C* which also acts as an r.f. bypass. Two designs are possible, one for peak power measurement and the other for average power measurement. For peak power measurement the *RC* constant of the circuit is such that the condenser will charge to the peak of the applied voltage and will discharge only slightly during the remainder of the cycle as shown in Fig. 5B. For average power measurements the *RC* constant is made such that the video or modulation voltage will appear across *C*. This output is then fed to a meter which measures average voltage. For example, if the r.f. is pulsed at 1000 times per second, the peak power design will provide an essentially constant output equal to the peak of the pulse, while an average current design would have a series of pulses at the condenser output. This type of power indicator is calibrated with a signal generator whose power output can be varied and is matched into the crystal detector.

Sufficient power must be extracted from the device under measurement to actuate the meter over an appreciable portion of its range. This is a serious disadvantage of this method since, in the slotted line for example, the more power absorbed by the probe, the greater the error in standing-wave ratio measured. As a consequence, the circuit shown in Fig. 5 is used only to provide a relative indication of power output in a microwave transmitter (thereby indicating proper operation) or in some standing-wave equipment which is not required to be very accurate.

The sensitivity of this method can be increased and the SWR error decreased by modulating the microwave signal and employing an a.c. amplifier following the crystal detector, as shown in Fig. 6, tuned to the modulating frequency. The amplitude of the modulating signal obtained at the crystal detector output will be proportional to the amount of r.f. power pickup. By providing sufficient amplification it is possible to effect a highly sensitive and accurate measurement of power with a very small probe $G_p$.

### Calorimeters

Another method of measuring power involves conversion of microwave energy into thermal (caloric) energy and measuring the latter. This can be done by feeding the microwave power into a water load matched to the output of the generator. The temperature rise of the water will be a function of the power absorbed and thereby provides an indication of output power.

A typical coaxial line calorimeter that employs this principle is shown in Fig. 8. This instrument consists of a coaxial line into which r.f. power is fed at its input and is terminated into a water load. The water is contained within a section of the coaxial line with inlet and outlet connections permitting a continuous flow of water throughout this section of line. The temperature of the inlet and outlet streams is measured as well as the rate of flow, and from this data it is possible to calculate the absorbed power.

Due to the high dielectric constant of water, a mismatch is produced at the junction of the air- and water-filled sections. This mismatch can be eliminated through the use of a dielectric transformer placed between the two sections. This transformer, which also

(*Continued on page 26A*)

## *Time and Cost Saving Features*

TYPE "A"
ASSEMBLY BIT

COMMON
SCREWDRIVER

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**1. What Other Screw** equals the high visibility of the CLUTCH HEAD recess to check out the slow-down of hesitation . . . even with "green" operators?

**2. What Other Screw** frees the line from burred and chewed-up heads with automatic straight driving . . . with Center Pivot entry that prevents driver canting?

**3. What Other Screw** has a non-tapered driving engagement (without dangerous "ride-out" as set up by tapered driving) to eliminate the hazard of skid damage? . . . and the need for fatiguing end pressure?

**4. What Other Screw** provides a simple Lock-On which unites screw and bit as a unit to hurdle "fumble spots" by permitting one-handed reaching and driving from any angle?

**5. What Other Driver** can begin to approach the durability record of the rugged Type "A" Bit . . . 214,000 screws driven non-stop?

**6. What Other Assembly Bit** can be *repeatedly* reconditioned on the spot by a 60-second application of the end surface to a grinding wheel . . . for unmatched tool economy.

**7. What Other Modern Screw** has a recess basically designed for operation with a common screwdriver . . . for the simplification of field service adjustments?

*According to users of CLUTCH HEAD Screws, you may confidently expect these time- and cost-saving features to deliver assembly production increases ranging from 15% to 50%.*

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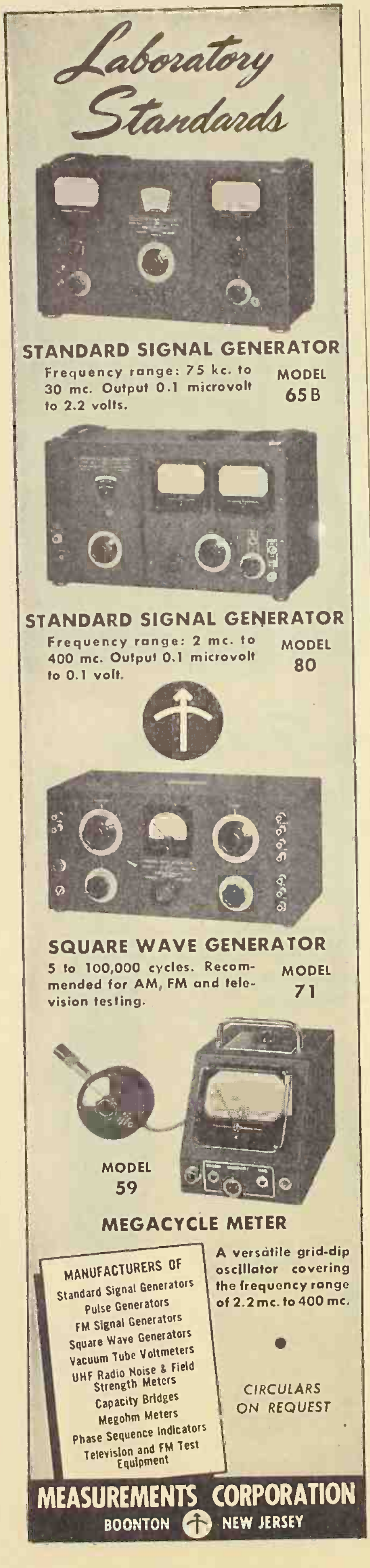
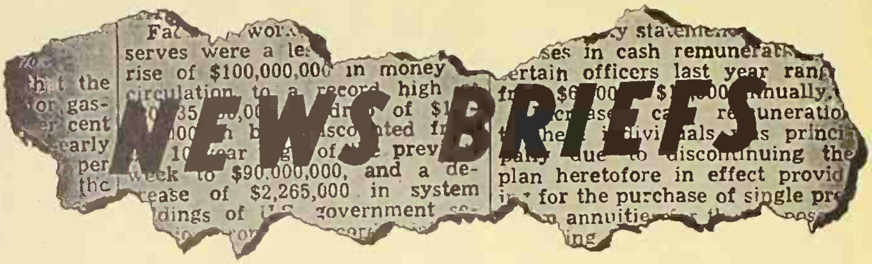




# NEWS BRIEFS

## AUDIO SYMPOSIUM

One of the main features of the IRE West Coast Convention held recently in Long Beach, California, was an Audio Symposium led by John Hilliard, Chief Engineer, *Altec Lansing Corp.*, which featured a discussion of contemporary problems in television audio.

Bryan Cole (seated at right) of KFI-TV illustrated the principal problems in television audio such as: microphone placement, acoustics of sets and studio, reduction of noise in studio, etc. Dr. J. G. Frayne (left), Chief Engineer of *Westrix Corp.*, led the discussion on magnetic recording and stated that magnetic recording in the movie industry is here to stay.

Fred Albin, Supervisor of Video Recording of *ABC*, headed the discussion on sound on film problems, and E. B. Harrison of *Altec Lansing Corp.*, discussed the engineering specifications for high fidelity audio transformers.

## TO BUILD ELECTROSTATIC ACCELERATOR

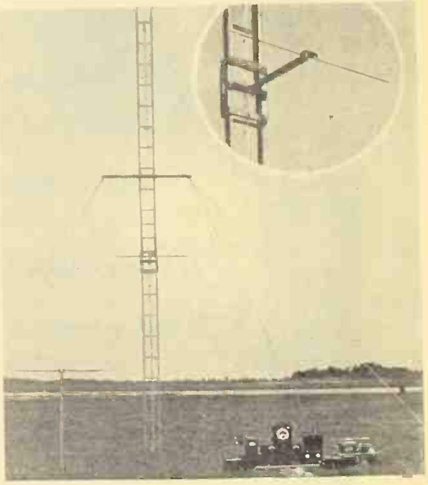
Northwestern University recently announced that it will build a 4½ million volt electrostatic accelerator for nuclear research. Dr. Russell A. Fisher, acting chairman of the Physics Department, said construction of the 28-ton atom smashing equipment will begin shortly, but it will be at least two years before the work is completed.

The instrument, a Van de Graaf type ion accelerator, will be installed in a structure to be built just south of the Northwestern Technological Institute. The lower part of the building probably will be underground so that the earth will serve as a shield for protection against the powerful rays produced by the generator. Additional protection will be provided by concrete walls and ceiling at least two feet thick.

The apparatus will be designed and constructed under supervision of Dr. Edward N. Strait, Jr., assistant professor in the Physics Department, who formerly worked with Professor R. J. Van de Graaf at MIT, and Dr. James H. Roberts, another nuclear physicist and assistant professor of physics. Consulting with them will be Dr. Paul E. Klopsteg, Director of Research in the N.U. Tech. Institute.

## CALIBRATION SERVICE

A calibration service for field-intensity meters at all radio frequencies of broadcast and commercial importance up to 300 megacycles is now offered by the National Bureau of Standards.



Of special interest are the new standards and methods which have been developed at the Bureau for calibrating field-intensity meters in the v.h.f. region from 30 to 300 megacycles. Two distinct experimental methods are used in the Bureau's field-intensity standardization work: the standard-antenna method for frequencies greater than 30 megacycles, and the standard-field method for lower frequencies.

In calibrating a commercial v.h.f. field-intensity meter by the standard-antenna method, the field strength at some arbitrary distance from a special v.h.f. transmitter is determined by a standard receiving antenna employing a crystal voltmeter. The antenna of the commercial set is substituted at the same position. The field strength, height of the antenna above ground, and the meter readings obtained with the two antennas enable one to compute the antenna coefficient that must be applied to the commercial instrument to relate field intensity to its meter readings.

### H.F. MEASUREMENTS CONFERENCE

The second High Frequency Measurements Conference sponsored jointly by the American Institute of Electrical Engineers, the Institute of Radio Engineers, and the National Bureau of Standards will be held in Washington, D. C. on January 10 to 12 in celebration of the semicentennial of NBS.

The Conference will be a forum at which leading engineers will exchange information on progress made since the previous Conference held in 1949. The Conference program will include about 25 technical papers, an evening demonstration, a luncheon, and conducted inspection tours of selected institutions.

The technical sessions will be held in the auditorium of the Department of the Interior. Conference Headquarters will be at the Hotel Statler. The Conference is under the general direction of Prof. Ernst Weber of the Microwave Research Institute of the Polytechnic Institute of Brooklyn. Dr. Harold Lyons of NBS is Chairman of the Local Arrangements Committee, and Dr. Frank Gaffney of the *Polytechnic Research and Development Company* is Chairman of the Technical Program Committee.

### METEOROLOGICAL BATTERY

A battery capable of powering meteorological equipment such as radiosonde to heights of 30 miles into the stratosphere has been developed by the Signal Corps Engineering Laboratories at Fort Monmouth, N. J. According to Signal Corps engineers, this battery is the result of years of work to find a compact, lightweight battery that could meet the desired standards of easy activation, long storage life, good low temperature operation, low weight, low cost, and high service.

### GERMANIUM PHOTOCELLS

George D. O'Neill, head of the Solid State Section of *Sylvania Electric Products Inc.*, presented a paper at the recent National Electronics Conference in Chicago describing a device consisting of a tiny piece of germanium placed in contact with a pointed wire for use as a light-actuated valve to control the flow of an electric current.

Such a device, called a germanium photocell, is smaller in diameter than a match stick and less than ½″ long, requiring only a single pair of connecting wires.

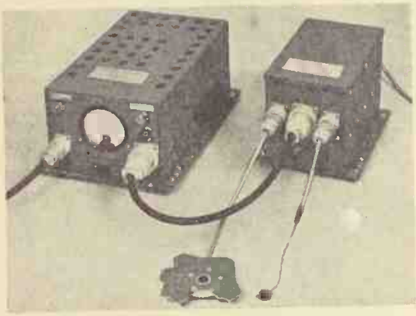
According to Mr. O'Neill, germanium

(*Continued on page* 31A)

# NEW PRODUCTS

### PRESSURE MEASURING SYSTEM

Elimination of the tubing connections in the new flush-mounted pressure cells announced by *Sierra Electronic Corp.*, 1110 County Rd., San Carlos, California, eliminates phase and amplitude errors arising from flexible-tube connection of pressure cells used in dynamic measurement. Cells mount directly in the surfaces upon which the measured pressures impinge.



Several styles are offered to fit various applications, including wedge-shaped types small enough to be used within ½″ of an airfoil trailing edge. Units are listed with these sensitivities: 0 to ± 2.5 psi, 0 to ± 5 psi, and 0 to ± 10 psi.

### SCALER

The *Atomic Instrument Co.*, 84 Massachusetts Ave., Boston 39, Mass., is now in production on its new 1010 Scaler, described as the first of a series of "building block" units.

While the standard Model 1010 is furnished optionally with a scale-of-100 or scale-of-256, added scaling assemblies to make either a scale-of-1000 or a scale-of-4096 may be specified. Other custom modifications include: scaling factor selector switch (10-100 or 4-16-64-256); precision calibrated 50 to 100 volt discriminator on front panel; electrical reset register; omission of manual reset register; omission of regulated high voltage power supply; etc.

A complete description of the 1010 Scaler may be obtained by writing the company.

### OSCILLOGRAPH

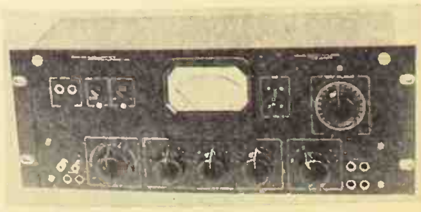
*Consolidated Engineering Corp.*, 620 No. Lake Ave., Pasadena 4, California, is now offering a small, low-cost recording oscillograph. Designated Type 5-116, the instrument is similar to the larger oscillographs manufactured by *Consolidated* but over-all dimensions have been appreciably reduced, and its weight cut by a factor of almost 50 per-cent.

A new record transport system, recently adopted by *Consolidated* for their standard recording oscillographs, is also included in the 5-116. In this system neither sprocket teeth nor a pressure roller is required to provide positive record engagement, thus removing the major sources of record-drive malfunction, and at the same time reducing the power required to drive the recording medium. Source of power for the record transport system is an exceptionally powerful governor-controlled motor directly connected through gearing to the record drive roll in the magazine.

The 5-116 is available in 9- or 14-trace block capacity for either 24-28 volt d. c. or 115-volt, 60-cycle a. c. drive.

### TRANSMISSION MEASURING SET

A transmission measuring set which eliminates lengthy calculations and intricate setups for checking audio gain or loss, measurement of matching and bridging devices, complex circuit readings, and mismatch loss and frequency response has been announced by the *RCA* Engineering Products Department.



The new equipment, *RCA* Type BI-11A, consists of a volume indicator meter, input and output attenuators, impedance matching system, and jacks for convenient connection. A meter multiplier, geared to the load-impedance shaft, provides an automatic correction for changes in load impedance. Convenient switches facilitate connection of the volume indicator to the input of the attenuator system, or to jacks for external connection. An output impedance switch allows matching to 600, 250, 150, 16, 8, or 4-ohm circuits.

Illustrated literature and specifications on the equipment are available on written request to the Broadcast Equipment Section of *RCA* in Camden, New Jersey.

### SURVEY METER

*Nuclear Instrument and Chemical Corp.*, 229 W. Erie St., Chicago, Illinois, now has available a survey meter for checking all types of radiation found in clinics, hospitals, and laboratories.

Model 2581 is a battery powered, non-discriminating rate meter, and is an improved version of the wartime "Zeuto." The detecting ionization chamber is covered with a rubber hydrochloride film on the underside of the instrument and is capable of detecting 25 k.e.v. beta particles and 2 m.e.v. alpha particles as well as gamma and x-radiation. This film is easily replaced and is protected by a removable wire grille.

The operating life of Model 2581 is at least 200 hours, and the batteries may be easily replaced by a minor disassembling of the instrument.

### DISTORTION AND NOISE METER

*The Daven Company* of Newark, New Jersey, announces the availability of its Type 35-A Distortion and Noise Meter with two amplifier gain controls provided; one an accurate step type covering the range +40 to —60 in steps of 10 db., the other a continuously variable control covering the range ±10 db.

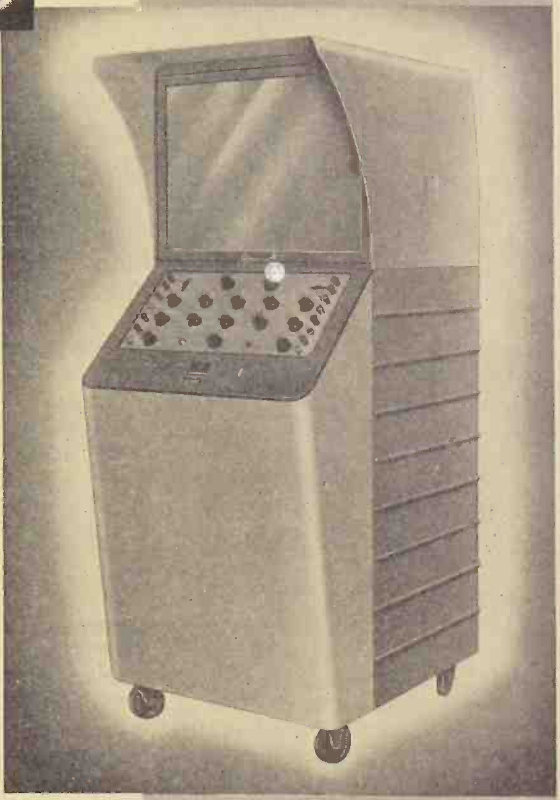
The indicating meter covers the range 0 to —15 db., the range of 35-A is +40 to —60 db. full scale meter reading or +40 to —75 db. utilizing the meter scale. The range of noise and distortion that can be measured depends upon the level of the source being investigated. At a level of —15 dbm. the limit is 60 db. below or 0.1% distortion. A pair of output jacks is provided for connecting an external scope in place of the output meter.

### MASS SPECTROMETER TUBE

Available from the *General Electric Company*, Schenectady, N. Y., is an ion resonance mass spectrometer tube to aid scientists in the analysis of chemical compounds, especially gases. High-precision measurements can be made by the new tube in a number of special analysis problems.

In the ion resonance tube, when in operation, electrons from the filament

(*Continued on page* 29A)



# Personals

**C. A. HAINES,** formerly general manager of the Photoflash Division of *Sylvania Electric Products Inc.*, is now general manager of operations for the Radio Tube Div. and the Television Picture Tube Div. Mr. Haines joined the factory engineering staff of *Sylvania* in 1929 and served as superintendent of their Salem radio tube plant and as general manufacturing manager of proximity fuse tube operations during World War II.

**JOHN H. HOWARD,** a consultant in the field of electronic control systems, has joined the staff of the Research Division of the *Burroughs Adding Machine Company* in Philadelphia. Following service with the U.S. Navy, Mr. Howard was appointed Director of Development at *Engineering Research Associates* and was later associated with the *Sperry Gyroscope Company.* He is a member of the AIEE and the Institute of Radio Engineers.

**DR. R. G. E. HUTTER,** head of the electronics research section of the Physics Laboratory, *Sylvania Electric Products Inc.*, has been appointed adjunct professor at the Brooklyn Polytechnic Institute where he will conduct classes in electron tube theory and electron optics. Dr. Hutter, a native of Berlin, Germany, served several years as a research physicist in the *Telefunken* transmitter laboratories and has been associated with *Sylvania* since 1944.

**PHILIPS B. PATTON** has been named Manager of the Sales Engineering Department of *Lenkurt Electric Co., Inc.*, San Carlos, California. Before joining the company as field engineer, Mr. Patton was chief of FCC's Radio Telephone-Telegraph Section, Common Carrier branch; flight radio officer with *Pan American World Airways;* a field engineer and technical coordinator with *Farnsworth Mobile Radio;* and was associated with *Western Union Telegraph Co.*

**BENJAMIN SAMPSON** has been appointed General Sales Manager of the *K. H. Huppert Company,* Chicago. Mr. Sampson was formerly District Sales Manager of the Stewart Div. of the *Sunbeam Corp.*, and recently Manager of the Industrial Furnace and Oven Div. of the *Claud S. Gordon Company.* Plans are under way to expand the Industrial Furnace Div. to meet the demand of their special applications on industrial furnaces, ovens, and ceramic kilns.

**DR. LAURISTON S. TAYLOR,** Chief of the Radiation Physics Laboratory of the National Bureau of Standards, delivered the Sylvanus Thompson Memorial Lecture before the meeting of the British Institute of Radiology recently. The first American scientist to receive this honor since the lectures were begun in 1916, Dr. Taylor is an internationally known authority on x-rays and has contributed extensively to scientific journals in the field of radiology.

## Microwave Measurements

(*Continued from page* 20A)

acts to keep water within its section, has an effective length of a quarter of a wavelength and a dielectric constant such that it acts as a quarter-wave transformer. It should be noted that in this setup matching is effected at one frequency only, with no provision for tuning to other frequencies.

Water calorimeters are useful only for measuring fairly large powers. Their operation is sluggish and heat losses are such as to prohibit their use for power smaller than a few watts. For larger powers, however, water loads may serve as very reliable power standards.

For the measurement of radio-frequency power in the range from 1 microwatt to several milliwatts, bolometer type instruments are usually used. The bolometer employs the characteristics of some conductors whose resistance varies as a function of power absorbed. This variation in resistance can be measured by bridge circuits and indicates the magnitude of power flow.

Bolometer elements, such as the thermistor, are manufactured by a number of companies and are usually contained within glass enclosures. As in other instruments described in this article an important problem in this equipment is effecting a match between generator and bolometer element. Matching at one frequency is relatively simple, but over a range of frequencies becomes more complex.

Fig. 10 shows a tunable mount which matches a 50-ohm line to a 200-ohm bolometer from 1000 to 4000 megacycles. This circuit uses a double stub tuner and employs the bolometer as the center conductor of the output connector. The r.f. power is fed across the input and the output operates into a bridge circuit.

Many types of bridge circuits can be used to determine bolometer resistance. It should be noted that if the bolometer element is allowed to vary in resistance appreciably, due to the absorption of r.f. power, the match between generator and this element may be affected. A circuit which overcomes this problem is shown in simplified block diagram form in Fig. 9. This circuit consists of a self-balancing bridge, with a bolometer as one of its arms, and a audio voltmeter. A high-gain amplifier is connected across the bridge as a detector and the output of this amplifier is fed back as a driving source for the bridge. This circuit will oscillate to maintain the bridge balanced and the design is such that bridge balance occurs when bolometer resistance is 200 ohms. The amplitude of oscillation is, therefore, such as to obtain a 200-ohm bolometer resistance. The frequency of the oscillator is determined by the bridge circuit and is a convenient audio frequency.

When r.f. power is applied through the bolometer element its resistance tends to increase and, to maintain bridge balance, the amplitude of the oscillator decreases so that the power flowing through this element remains the same. The reduction of audio power is equal to r.f. power, and hence an audio voltmeter (zero set for r.f. power) can be calibrated in terms of r.f. power. It should be noted that in this system the bolometer element has a 200 ohm impedance at all times and hence the proper matching is assured. An accuracy within 5 per-cent can be effected with this circuit.

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*(To be continued)*

## Potted Circuits

(*Continued from page* 8A)

to eight per-cent in the setting process. However, the great bulk of this shrinkage takes place at an early gel stage, when the material is still quite soft, and no undue pressures are therefore exerted on the components embedded.

On the other hand, these materials have a greater coefficient of thermal expansion than most of the metals and glass usually enclosed in them. No difficulty is experienced at higher than normal temperature operation from this effect. When low temperature operation is to be experienced—temperatures below about minus 30 degrees Centigrade —special treatment of the circuit to be embedded is required. Equipment has been successfully operated at temperatures as low as minus 85 degrees C.

Silicone plastics, such as *Dow-Corning* Silastic 181, remain flexible at extreme temperatures and make suitable cushions when placed between the circuit and the plastic which surrounds it. The Silastic is dispersed in a solvent so that the circuit to be protected can be dipped in it. Several dips are required, with treatment in an oven after each dip to drive off the volatile material and vulcanize the Silastic.

Unless the Silastic is vulcanized, trouble may be experienced. Under the heat of operation of the tubes embedded in the plastic, it will undergo the vulcanization process, liberating certain materials. This results in the development of a pressure within the casting which may result in fissures.

The photographs accompanying this article show some of the applications to which the casting resin has been put at the Navy Electronics Laboratory. A large number of cylindrical subassemblies one inch in diameter and two to two and one-half inches long have been made. These contain single stages or at most two stages, such as multivibrators, oscillators, amplifiers, modulators, etc., as complete units.

It will be noted that no chassis as such has been used in any of these units. Leads are necessarily short and as a result the circuits are normally sufficiently self-supporting to get them embedded in the plastic before something happens. Most of those illustrated use a seven or a nine pin miniature base in order to plug into the corresponding socket. However, any type of connection means may be used which is desirable for the application in mind. Circuits are wired in a jig which holds the plug pins in position and the components built up by point-to-point wiring.

Also illustrated is a superheterodyne receiver containing six tubes and cast as a unit. This particular receiver covers the standard broadcast band and was built as a "propaganda" unit to prove to some skeptics that the things would continue to operate after being potted. It is not intended as a representative design, either electrically or mechanically, for military applications. It has, nevertheless, created considerable interest. The speaker and power supply are enclosed in the base into which the receiver plugs. The power supply uses a selenium rectifier which is also potted for protection.

The formulation and processing of NEL-177 Casting Resin are fully covered by a patent disclosure in the name of Mr. Edward Rolle, NEL chemist, who is chiefly responsible for the development of the materials and processes involved. The details of these may be obtained for use by firms legitimately involved in production for Government use which requires the use of such compounds.

Others who contributed to the program are: Messrs. J. C. McAdam, A. T. Steinkamp and J. R. Potthoff, who did the bulk of the circuit work; Mr. R. J. Violette, who developed the molds and did a large amount of the actual casting; Mr. A. H. Attebery, whose assistance in selection and procurement of suitable components was extremely valuable; and last but not least, Mr. E. B. Robinson, whose foresight started the whole program. -⊛-

# TECHNICAL BOOKS

**"SURVEY OF MODERN ELECTRONICS"** by Paul G. Andres, Associate Prof. Electrical Engineering, Illinois Institute of Technology. Published by *John Wiley & Sons, Inc.*, 440 Fourth Ave., New York 16, N. Y. 522 pages. $5.75.

This is a textbook for a short course in electronics based on lectures and classroom notes from a course given to mechanical, chemical, and industrial engineering students to familiarize them with the basic principles of construction, operation, and application of electron tubes.

It explains the fundamentals of electronics and provides a summary of applications. The material is treated in a descriptive manner rather than mathematical, and the description of nearly every tube is followed by one or more practical applications. Final chapters review applications of electronics in instrumentation, communication, control, and heating, and offer practice in reading circuit diagrams.

General references in the appendix and specific references and problems at the end of each chapter are given for additional study. Data, circuits, and illustrations are also included.

**"ANTENNAS"** by John D. Kraus, Prof. Electrical Engineering, The Ohio State University. Published by *The McGraw-Hill Book Co.*, 330 West 42nd St., New York 18, N. Y. 553 pages. $8.00.

Here is a clear, systematic treatment of basic antenna theory and its applications to a great number of antenna systems, compiled from lectures given in recent years by the author in a course on antennas at Ohio State.

The text presents a unified treatment of antennas from the electromagnetic theory point of view with stress upon those aspects which are of engineering importance. The principles given are basic and are applied to antennas for all frequency ranges. Some of the material is published here for the first time, particularly portions of the treatment on point sources and on helical antennas. Recent advances in the field are described, and problem sets are included at the end of each chapter. The rationalized mks system is employed and a complete table of units in this system is included in the Appendix.

Although primarily written to serve as a textbook, it is hoped that practicing engineers and scientists will find it a valuable reference book.

## Miniaturization

(*Continued from page* 11A)

terioration were also promised.

Next, the general problem of insulation was considered. The hotter a transformer is run, the smaller can be its size. However, the usual organic material used to insulate wire, and the insulating tapes used for start, interlayer and finish wraps, will carbonize if they are pushed beyond their safe high temperature limits.

Thus, an inorganic insulated wire and an inorganic paper for the insulating tape were chosen. The former is a wire coated with a ceramic material. The latter is composed of asbestos with a binder, and is available in sheet or tape form, in a variety of widths, and in thicknesses down to 0.003". Its ability to withstand heat without disintegration is unusual. A few thousandths of asbestos paper is sufficient to stop the passage of the direct flame of a blow torch. Finally, as an impregnant, a silicone varnish was picked for its inorganic composition and relative ease of production use.

It was found that interlayer insulation in many instances was unnecessary. Where possible, random winding was used—that is—machine-winding without interlayer insulation. This meant, of course, that the potential between the two adjacent layers was the difference between the potential of the first turn of the first layer and the last turn of the next layer. This potential is usually a matter of only a few volts and, if the coil is properly impregnated, sufficient insulation appears to exist. Inter-winding insulation of asbestos paper was always provided.

When impregnation was first considered, a material was sought which would act as impregnant, seal and outer shell. However, no such material was found which could qualify, as available impregnants were not true hermetic seals. Furthermore, it was calculated that if a dipped seal was used, the weight of the seal would be greater than the weight of a conventional brass can.

The use of a silicone varnish as an impregnant consists of first driving off the moisture in the coil and then subjecting it to a series of impregnation and bake cycles. The impregnant fills in the spaces between the wires, seals the coil and allows maximum heat transfer.

Military requirements usually call for hermetically sealed transformers and chokes. As previously stated, this is normally done by using a solder-sealed brass or copper can in which the transformer is mounted and held in place by an impregnant fill or compound commonly known as "gunk".

Our survey started with the knowledge that the specific gravity of copper is 8.91 as compared with 2.7 for aluminum. Heretofore, aluminum could not be used for cans because of the difficulty of solder-sealing. Iron or zinc with 7.85 and 7.1 specific gravities respectively showed little weight advantage. Magnesium, with a 1.8 specific gravity, could not be obtained in comparable wall thicknesses. It appeared that aluminum was the desired material if some process could be evolved for solder-sealing it.

Aluminum-plating techniques were developed whereby the fabricated can parts could be plated with copper or nickel. This resulted in a can which could be soldered readily, yet gave a 70% reduction in weight.

Early in the development of the plated aluminum, some discussion arose over the possibility of electrolysis developing between the base aluminum and the copper plating (due to their removed position in the electromotive series). However, extended tests indicate that, when the developed plating process is followed, this condition is not existent in practice. Plated strips of aluminum have been subjected to 500 hours of salt-spray without deleterious effects.

The use of a compound or "gunk," which anchors the transformer in the can and conducts heat from the transformer to the can, has been the standard practice for some years. However, the weight of the compound often approached the weight of the transformer and can. By using an alternative method, we accomplished the same result at a saving in weight and cost.

This we did by mounting the transformer with special aluminum braces, evacuating the can, re-filling the can with a non-explosive, non-inflammable gas and sealing off the can against leakage.

For transformer terminals, the small glass bead type was used. They withstand applied voltages and exhibit no leakage between can and terminal stem so long as the glass bead does not become cracked in assembly.

As a final illustration of what has been done in transformer miniaturization with newer materials and techniques, Fig. 5 illustrates a 400-cycle 115-volt excitation transformer which delivers 4.2 watts at 24.25 volts into a resistive load with an efficiency of 73.9%. It can deliver 7.06 watts into a resistive load with an efficiency of 62%. Yet, this hermetically-sealed transformer measures 1¼" x ¾" x ⅞" and weighs only *one ounce*! Compare this with a previous comparable design which occupies approximately 900% more space and weighs approximately 10 times as much.

The use of subminiature tubes (as shown in the assembly illustrated in Fig. 1) allows the designer to utilize better the space available. Newer and more compact condensers save much additional weight and space. The failure of electrolytic condensers to hold capacity at low ambient temperatures (such as -55 degrees C.) results in the forced use of paper and mica types. Thus, a 30% to 50% saving in weight and size of paper condensers used in a power supply filter circuit can mean an appreciable saving in total size and weight of the complete electronic system.

The Scotch saying, "Many a mickle makes a muckle" aptly describes the process of miniaturization. A fraction of an ounce saved and multiplied many times means pounds saved in the final electronic equipment. However, this process can not be accomplished over night. It is a slow, painstaking inching forward requiring the best of laboratory equipment and personnel working as a team to advance the cause of science and to provide our military with the very best in operational equipment —modern as tomorrow.

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## Millivolt Chopper

(*Continued from page* 14A)

modulator, while the supply was approximately 12-25 volts. A *Sylvania* 1N42 Varistor was employed as the modulator. The carrier voltage was supplied through an isolation transformer so as to not introduce capacitively unbalanced line frequency signals into the oscilloscope.

The output of the electronic chopper may be connected into an oscilloscope or amplifier. In the typical circuits illustrated, input into a schematic tube is shown. Tube biasing was not shown, as this may depend on a number of variables. No biasing at all may be required with a high mu tube with low plate potential; while fixed bias, where bias is required, will give higher gain than cathode bias due to elimination of negative regeneration from the cathode circuit. The total amplifier gain required will depend on the operating conditions of input signal available and output signal required.

The amplified d.c. signal may be read or indicated four ways. Visual observation of an oscilloscope, a.c. output of the amplifier measured, or the a.c. output rectified into d.c., any residual (or hash) signal balanced out and the differential read on a d.c. meter. The fourth method and possibly one of the best is the use of an automatic potentiometer circuit. Here the amplified signal is used to turn a potentiometer arm and balance out the d.c. input signal. The potentiometer arm position is then read in terms of d.c. input (the voltage across the potentiometer is standardized) and any amplifier or chopper gain variation affects the accuracy of reading only to a small percentage of the total accuracy.

The *Dumont* 274 oscilloscope used by the author in these experiments was not a standard instrument. Its gain had been increased by the addition of a cascade 6J7 and it compared in gain with the *Dumont* 208B and other higher priced instruments. With full gain and some unbalance carrier appearing on the screen, the output signal level was sufficient to cause about a one- to two-inch variation in trace. The chopper efficiency was not established, but was 50% or lower; therefore the input signal did not exceed 25 millivolts.

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## New Products

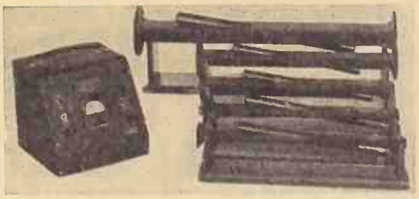
(*Continued from page* 25A)

are accelerated and form ions which are in turn accelerated in a space where crossed magnetic and r. f. electric fields are maintained. In these crossed fields, ions describe spiral paths. Those ions with the proper mass to resonate with the crossed fields will gain energy and describe larger and larger orbits until they reach the collector and are measured.

Approximately 150 volts d.c. and r.f. voltages of about 1 volt at frequencies up to 5 megacycles are employed. The filament utilizes 5 amperes at 15 watts.

### NOISE GENERATORS

A series of random noise generators designed to produce a known output noise in the frequency range of 2600 to 12,400 mc. has been announced by *The Kay Electric Co.*, Maple Ave., Pine Brook, N. J.



Designated Microwave Mega-Nodes, they are a group of five separate noise sources, each utilizing a standard commercial fluorescent lamp in a JAN size wave guide, each mounted on a hardwood stand. A separate cabinet houses the power supply for the noise sources.

The following wave guide sizes are available: RG48/U, RG49/U, RG50/U, RG51/U, and RG52/U. Noise output is 15.84 db. above thermal noise at a wave guide temperature of 32° C.

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## News Briefs

*(Continued from page* 23A)

photocells are of three types: the photoresistance cell, able to produce electric signals at least as great as 25 volts when a light shining on the cell is interrupted or modulated; the photoconductance cell, having a large current response for operating a relay directly in applications such as alarms and door openers; and the photovoltaic cell, similar to a tiny battery but having the peculiar property of supplying a voltage in proportion to the amount of light shining on the germanium.

* * *

## 3-Probe Method

*(Continued from page* 9A)

$$\frac{E_1}{E_3} = -j\frac{Z_R}{Z_0};\ \frac{E_2}{E_3} = \frac{1}{\sqrt{2}}\left(1 - j\frac{Z_R}{Z_0}\right) \qquad (5)$$

Substituting for $Z_R/Z_0$ its value $(R/Z_0) + j(X/Z_0)$, these are:

$$\frac{R}{Z_0} + j\frac{X}{Z_0} = j\frac{E_1}{E_3} \quad . \ . \ . \ . \ . \ . \ . \quad (6)$$

$$\frac{R}{Z_0} + j\left(\frac{X}{Z_0} + 1\right) = j\sqrt{2}\frac{E_2}{E_3} \quad . \ . \ . \quad (7)$$

It is seen that these two equations, when the moduli are squared,

$$\left(\frac{R}{Z_0}\right)^2 + \left(\frac{X}{Z_0}\right)^2 = \left(\frac{E_1}{E_3}\right)^2 \quad . \ . \ . \ . \quad (8)$$

$$\left(\frac{R}{Z_0}\right)^2 + \left(\frac{X}{Z_0} - [-1]\right)^2 = \left(\sqrt{2}\frac{E_2}{E_3}\right)^2 \quad (9)$$

define two circle families; a family of circles centered at (0, 0) with radii of $E_1/E_3$, and a family of circles centered at (0,-1) with radii of $\sqrt{2}(E_2/E_3)$. These are plotted in Fig. 1. The intersection of any two of these circles defines normalized impedance uniquely, and additionally, if so desired, the voltage standing-wave ratio.

For practical use, however, a chart in a form other than this is to be preferred. The present map extends to infinity in three directions. These infinite boundaries can be collapsed into a single point if the coordinates are distorted according to the bilinear transformation:

$$\frac{W}{Z_0} = \frac{1 + Z/Z_0}{1 - Z/Z_0} \quad . \ . \ . \ . \ . \ . \ . \ . \quad (10)$$

This particular complex-plane transformation can be employed to produce either the Smith Chart ("Transmission Line Calculator," *Electronics*, January 1939; "An Improved Transmission Line Calculator," *Electronics*, January 1944, both written by Phillip H. Smith) or the Carter Chart ("Charts For Transmission-Line Measurements and Computation" by P. S. Carter, *RCA Review*, January 1939).

At the outset, the circle family defined by:

$$\left(\frac{R}{Z_0}\right)^2 + \left(\frac{X}{Z_0}\right)^2 = \left(\frac{E_1}{E_3}\right)^2,$$

$$\text{or } \left|\frac{Z_R}{Z_0}\right| = \left|\frac{E_1}{E_3}\right| \quad . \ . \ . \ . \ . \quad (11)$$

is seen to be identical with the modulus of the Carter Chart, $Z_R/Z_0$, and so it can be transformed into the new $W/Z_0$ by:

$$\left(\frac{R}{Z_0} - \frac{(E_1/E_3)^2 + 1}{(E_1/E_3)^2 - 1}\right)^2 + \left(\frac{X}{Z_0}\right)^2 =$$

$$\left(\frac{2\,E_1/E_3}{(E_1/E_3)^2 - 1}\right)^2 \quad . \ . \ . \ . \ . \quad (12)$$

It is difficult to so treat the other circle family. However, since these circles can easily be drawn graphically in the original $Z/Z_0$, and since the coordinates are transformed conformally in the Smith Chart, a point-for-point plot enables this latter family to be carried over to the new $W/Z_0$ plane. These circle families are shown on P. 32A.

Aside from the drafting error in the construction of this map, P. 32A, an additional inaccuracy lies in the fact that, in some regions of this map, the intersecting circles are not orthogonal. Thus, in these regions, it is difficult to obtain a clear-cut "fix." As examples of this, it will be noted that the voltage ratios of .8 and .5 for $E_1/E_3$ and $E_2/E_3$, respectively, intersect nearly at right angles, and, so, a definite value of impedance can be obtained if this chart is used as an overlay for either the Smith- or the Carter-Chart, with which charts this one has a one-to-one interior correspondence. On the other hand, ratios of 1.3 and 1.6 for $E_1/E_3$ and $E_2/E_3$, respectively, intersect at such an acute angle that any chosen impedance value would be rather doubtful. This fault of interpretation is not caused by the transformation, since the transformation is conformal, but lies, rather, in the setup of the original chart, Fig. 1.

However, this new map has several advantages. In common with both the Smith- and Carter-Charts, the entire right-hand half of the infinite plane is contained within the bounding circle, which contains all impedance values within the range $(0 < R/Z_0 < \infty)$, $(-\infty < X/Z_0 < \infty)$. Again, in common with these charts, the circles of constant VSWR are concentric with the center of the chart. In this connection, the arm shown in the figure may be pivoted at the center, or a circle family in contrasting color may be drawn using the chart center and a radius equal to $E_1/E_3$ equal to 1.1, 1.2, 1.3, 1.4, . . . . Thus, ratios of 1.4 and .7 for $E_1/E_3$ and $E_2/E_3$, respectively, indicate a VSWR of 2.6.

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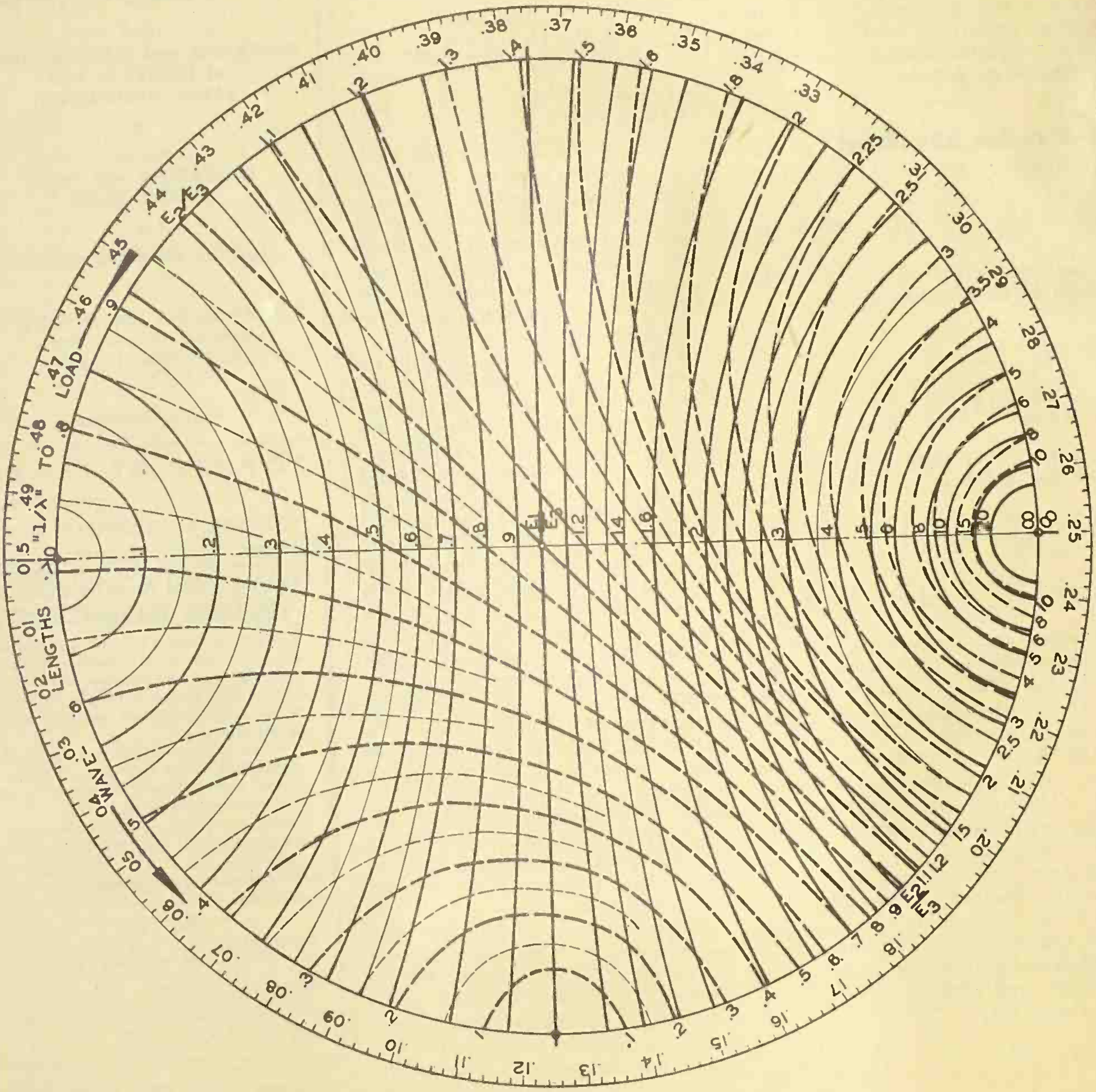
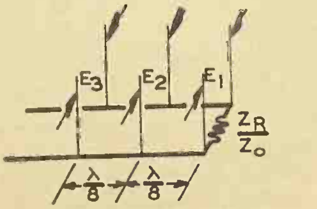
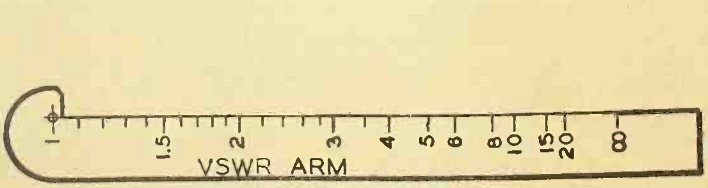
# 3-PROBE IMPEDANCE MEASUREMENT CHART

***A chart for determining load impedance from voltage readings taken from three fixed probes spaced 1/8 wavelength apart.***

*(See page* 9A)





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