AUGUST, 1950

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AUGUST, 1950

DEPARTMENTS

RADIO-ELECTRONIC ENGINEERING is published each month as a special edition in a limited number of copies of RADIO & TELEVISION NEWS, by the Ziff-Davis Publishing Company, 185 N. Wabash Avenue, Chicago 1, Illinois.

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COVER PHOTO — Courtesy of RCA

The new RCA permanent-magnet electron microscope consists of four units: the microscope, control unit, high-voltage unit, and vacuum pump. The photograph shows the microscope in use. A carefully regulated 50 kv. accelerating potential is used, and the permanent magnets eliminate the necessity of adjusting the magnetic field.

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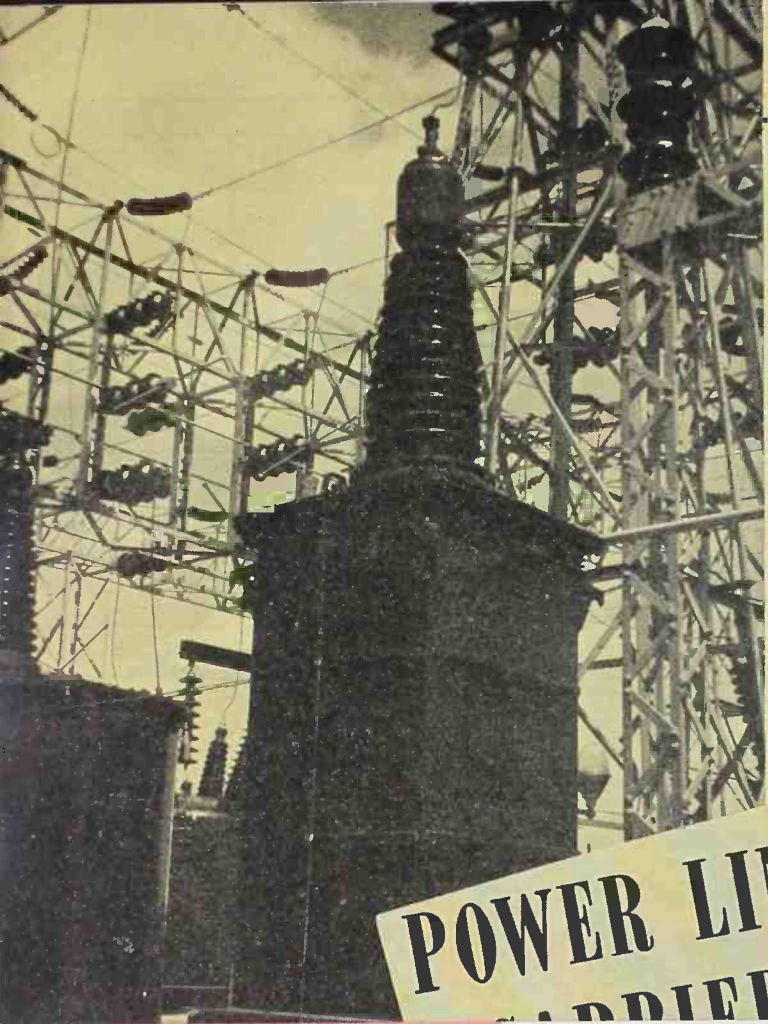

POWER LINE CARRIER

onstrates some of its applications, and Fig. 6 illustrates the basic requirements of one terminal.

The fundamental principle of power line carrier is to superimpose carrier frequencies in the radio spectrum of 50 kc. to 150 kc. on high tension lines without disrupting the transfer of energy at power line frequencies.

The carrier transmitter and receiver

The large number of generating stations on modern power systems and the increasing practice of interconnecting power systems make it necessary for the load dispatcher to know the generating conditions and load distribution at distant points of the system. Telemetering provides an ideal method of transmitting these indications to the dispatcher's office. The indications can be used by the dispatcher in manually dispatching the load or they can be used to operate automatic load control equipment.

Amplitude modulation or frequency modulation equipment is used for telemetering quantities. For control, supervisory, and remote trip indications, audio tone generators are used to modulate the carrier of a transmitter. The audio frequencies used are from 300 cycles to 3000 cycles. As many as 11 channels are possible by this method. At the receiver end, the tones are demodulated, and the rectified signal operates audio filter relay units, which in turn actuate indicating instruments. Fig. 5 shows an 11 channel telemetering receiver.

A new method of telemetering has been devised by *General Electric*, and has the trade name of frequency-shift. Frequency-shift uses a method of transmitting two different control or indicating impulses, in which the transmitted carrier frequency is shifted and the two frequencies are distinguished at the receiver by selective devices.

Capacitance coupling has proved to be the most economical method of obtaining small amounts of low voltage energy from high potential sources. Its application in combination with potential devices makes possible a wider field of application for carrier-current communication, relaying, telemetering, and supervisory control than was previously
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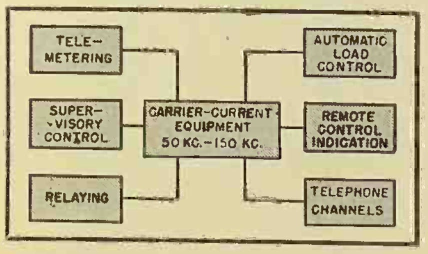

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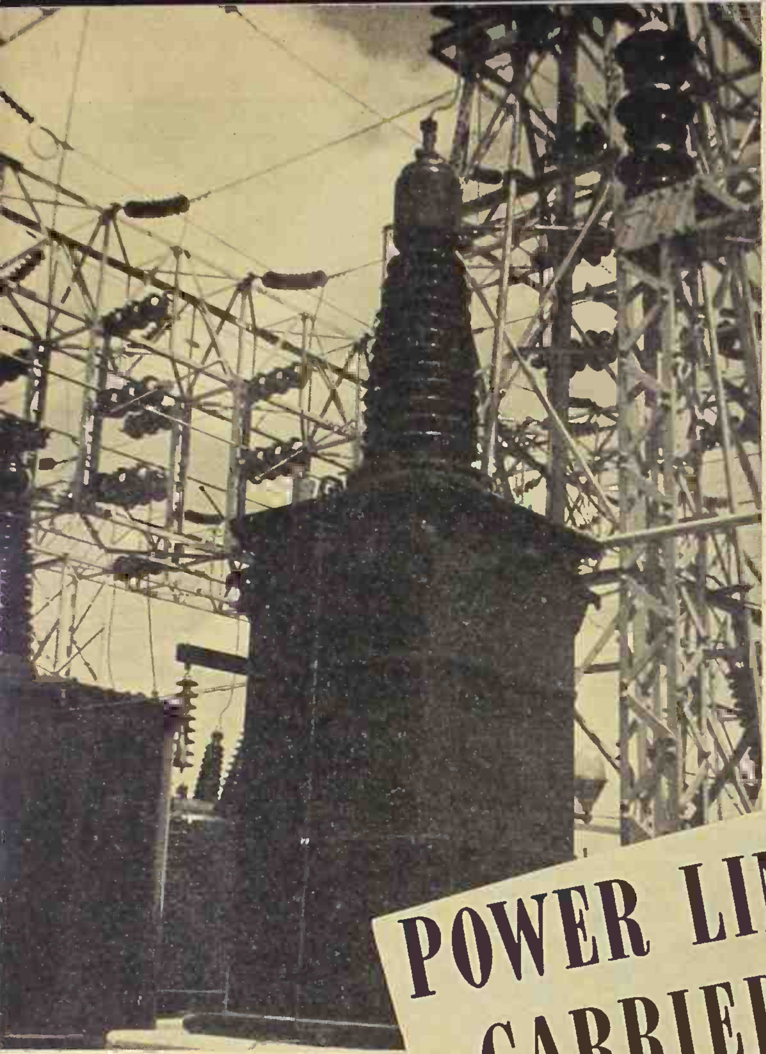


Fig. 1. Typical potential transformer.

POWER LINE CARRIER

By NICHOLAS ALCHUK

Equipment and techniques for carrier-current power line communications and telemetering.

THE utilization of power line carrier-current equipment has advanced rapidly within the past fifteen years. It has demonstrated that many services are better and more economically performed than by means of its predecessor, the pilot or telephone wire.

Today many variations of power line carrier apparatus are being used in the United States and Canada. Fig. 2 demonstrates some of its applications, and Fig. 6 illustrates the basic requirements of one terminal.

The fundamental principle of power line carrier is to superimpose carrier frequencies in the radio spectrum of 50 kc. to 150 kc. on high tension lines without disrupting the transfer of energy at power line frequencies.

The carrier transmitter and receiver bear a certain resemblance to some radio equipment, but differ in that carrier frequencies are transmitted over power lines instead of being radiated into space. Telemetering, relaying, and supervisory control are but a few features performed by power line carrier apparatus.

Telemetering

The large number of generating stations on modern power systems and the increasing practice of interconnecting power systems make it necessary for the load dispatcher to know the generating conditions and load distribution at distant points of the system. Telemetering provides an ideal method of transmitting these indications to the dispatcher's office. The indications can be used by the dispatcher in manually dispatching the load or they can be used to operate automatic load control equipment.

Amplitude modulation or frequency modulation equipment is used for telemetering quantities. For control, supervisory, and remote trip indications, audio tone generators are used to modulate the carrier of a transmitter. The audio frequencies used are from 300 cycles to 3000 cycles. As many as 11 channels are possible by this method. At the receiver end, the tones are demodulated, and the rectified signal operates audio filter relay units, which in turn actuate indicating instruments. Fig. 5 shows an 11 channel telemetering receiver.

A new method of telemetering has been devised by *General Electric*, and has the trade name of frequency-shift. Frequency-shift uses a method of transmitting two different control or indicating impulses, in which the transmitted carrier frequency is shifted and the two frequencies are distinguished at the receiver by selective devices.

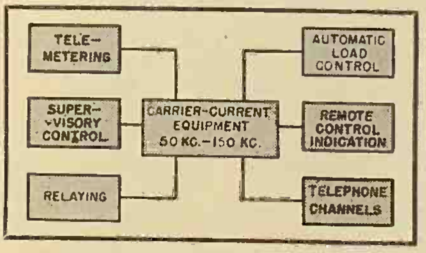
Capacitance coupling has proved to be the most economical method of obtaining small amounts of low voltage energy from high potential sources. Its application in combination with potential devices makes possible a wider field of application for carrier-current communication, relaying, telemetering, and supervisory control than was previously afforded through the use of the conventional potential transformer. The coupling is obtained directly from the high tension line without the use of auxiliary units. Fig. 1 shows a potential transformer.

The capacitor can be used simultaneously as a channel for the carrier frequencies and to supply a potential network, suitable for the operation of relays and indicating instruments.

The individual capacitor units consist of a wet process porcelain housing with fitted ends designed to effectively seal the interior. This housing contains the capacitor element which consists of a large number of individual sections connected in series. Capacity values vary from .0007 μfd. to .004 μfd.

The number of high voltage coupling capacitors that are required in any station depends upon the number of lines to which the coupling is required and the

Fig. 2. Some applications of power line carrier equipment.



400 MC. PERFORMANCE OF TYPE 5840 PENTODE

By NORMAN B. RITCHEY

Application Lab., Sylvania Elec. Prod. Inc.

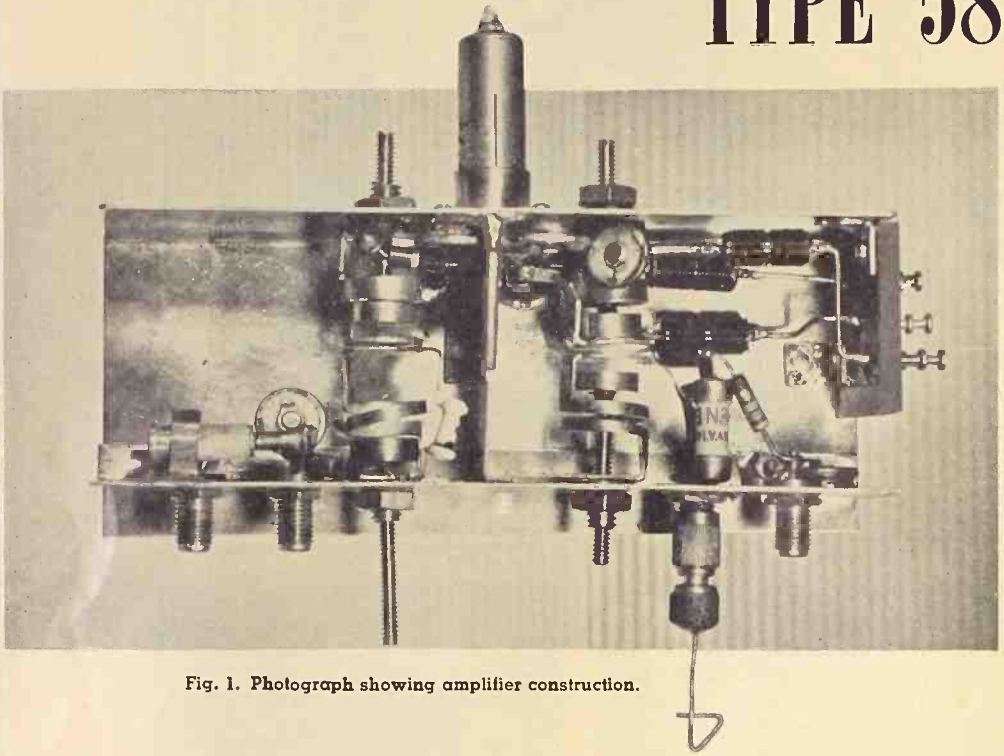


Fig. 1. Photograph showing amplifier construction.

Tests on this pentode give clues to the design and construction of satisfactory 400 mc. r.f. amplifiers.

IN evaluating the performance of an ultra-high frequency amplifier pentode it must be borne in mind that tubes at these frequencies can no longer be considered as high impedance devices. An analysis of amplifier performance must necessarily include the tube impedances. Input impedance has been shown by theoretical considerations and low frequency measurements to vary inversely as the square of frequency. However, at the higher frequencies (above 100 megacycles) measurements at this laboratory have shown an input resistance at the tube terminals which is always less than that predicted by the square law theory. A further investigation of the discrepancy led to a study of the effects of tube leads. The points external to the tube at which the measurements are made are necessarily separated from the points at which the measurement is desired. The inductance and distributed capacitance of the tube leads have a considerable effect at u.h.f.

An example may show the nature of this effect. Assume a pentode with several cathode leads for separation of input and output returns. The grid and one cathode lead inductance are then in series and may be combined as L. Then an approximate equivalent circuit for the input of an r.f. pentode may be drawn as shown in Fig. 2A, where G and K are the actual structure and T-T' are the external terminals to which the measuring equipment is connected. This circuit transforms to that shown in Fig. 2B, where R'_{input} is the resistance which will be seen by the measuring equipment.

It can be shown that:

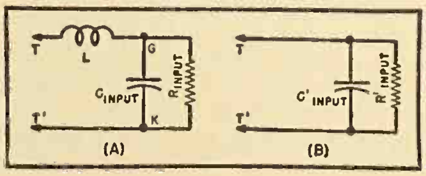
$$R'_{in} = R_{in}(1 - \omega^2 L\,C_{in})^2 + \frac{\omega^2 L^2}{R_{in}} \quad . \quad (1)$$

For typical values of L encountered in subminiature tubes the term $\omega^2 L^2/R_{in}$ may be neglected at 400 megacycles and the equation reduces to:

$$R_{in}' \cong (1 - \omega^2 LC_{in})^2 R_{in} \quad . \quad . \quad . \quad . \quad (2)$$

which at a given frequency is of the form $R'_{in} \cong KR_{in}$. This step down factor at 400 megacycles is approximately 0.4 for the structure used in the tube type 5840.

Fig. 2. (A) Approximate equivalent circuit for the input of an r.f. pentode. This circuit transforms to that shown in (B).



This step down or transformation explains a measurement vs. theory conflict which has been bothersome for some time, but is of somewhat academic interest inasmuch as it does not have any significance as far as tube gain is concerned. A tube amplifies the voltage present at the electrodes rather than that at the terminals, hence the value of input resistance desired for gain calculations is that at the structure. This is the value obtained by square law extrapolation from lower frequency data and it has been noted all along that this is the only value which will explain the actual measured gains. Output impedances were evaluated by a Q meter method.

The pertinent tube data are tabulated below.

Conditions:

$E_f = 6.3$ volts
$E_{bb} = 100$ volts
$E_{c2} = 100$ volts
$R_k = 150$ ohms

Measured:

$G_m = 5000$ μmhos
$r_p = 230{,}000$ ohms
$C_{gp} = 0.015$ $\mu\mu$fd. max.
$C_{in} = 4.2$ $\mu\mu$fd. (cold)
$C_{out} = 4.0$ $\mu\mu$fd.
$\Delta_1 C_{in}$ increase of 0.5 $\mu\mu$fd. due to expansion of parts when heater is lit with the grid at cut-off.
$\Delta_2 C_{in}$ increase of 0.9 $\mu\mu$fd. due to space charge with normal currents through tube.
Input resistance approx. 437 ohms at 400 mc.
Output resistance approx. 5000 ohms at 400 mc.
Subminiature socket capacitances—1 pin to all others 0.8 $\mu\mu$fd.
Allowance of 1.5 $\mu\mu$fd. for stray wiring capacitance.

The impedance values given above lead to one of the basic precepts of u.h.f. amplifier design. The low values of input resistance encountered in even the best pentodes of conventional structure make the amplifier problem one of power amplification. Presuming a constant input impedance to a following stage, the optimum first stage design is that which will develop the maximum voltage at the second grid. In a resistive load the maximum voltage coincides with maximum power, leading to power amplifier considerations and, thus, impedance matching into and out of the tube. This impedance matching is of the utmost importance in realizing maximum gain through the stage.

When considering a single stage, as in this case, the amplifier stage must be operated into a dummy load resistance which simulates the actual input resistance loading of the grid which normally follows, in order that the measured gains will actually indicate useful and realizable gain figures.

The tube impedances are quite low and are the limiting factor on circuit selectivity. However, it is still desirable to maintain a comparatively high Q in the tuned circuits. Iron core inductances have too low a Q at 400 mc. to be satisfactory and silver slug tuned coils have a very limited tuning range which makes for critical tolerances on tube capacitances. Something new in the way of u.h.f. tuned circuits was dictated. A variable L-C tuner developed by V. H. Aske of this laboratory filled this need. A sketch of this unit is shown in Fig. 4. This tuner is composed of a capacitor plate, an inductive loop and a movable brass slug so arranged as to increase or decrease the capacitance and inductance simultaneously.

The advantages of this unit are wide tuning range, more nearly constant L-C ratio, uniformly high Q over the range, and freedom from the spurious modes and multiple paths of the butterfly construction. The tuner is small in size and lends itself to subminiature construction.

Returning to the design of a subminiature amplifier, a block diagram of the basic laboratory circuit is shown in Fig. 3. The impedance levels at various parts of the circuit are shown. Without inquiring into the nature of the input circuit it can be seen that some voltage gain is possible due only to the impedance transformation. This voltage gain is proportional to the square root of the impedance ratios as in any transformer. In actual application this gain would depend on the source impedance and would be present only in the first stage of an amplifier.

The gain figure of primary interest is the actual power gain from the grid of the first stage to the grid of the following stage. A basic formula can be developed for this gain if the assumptions of matched impedances and tuned circuit impedances much higher than tube impedances are allowed. Both assumptions are realizable conditions.

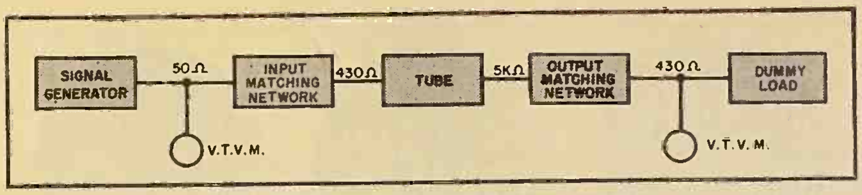


Fig. 3. Block diagram for a single stage amplifier.

From the Norton equivalent circuit the tube gain can be calculated as the product of G_m and output circuit impedance.

$$\text{Tube Gain} = G_m Z_L \quad . \quad . \quad . \quad . \quad . \qquad (3)$$

If matched conditions at the output are assumed, the dummy load simulating the input of the next stage must be stepped up by the output network to match the output impedance of the tube. The load seen by the tube becomes the output impedance in parallel with itself or $R_{out}/2$. Thus the grid-to-plate tube gain is equal to:

$$\frac{G_m R_{out}}{2} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \qquad (4)$$

However, it can be seen that this voltage gain is not actually attainable because the output is not taken at the plate but across the dummy load (or following grid). The voltage at the following grid is stepped down by the output network.

Then the realizable tube gain is:

$$\frac{G_m R_{out}}{2}\sqrt{\frac{R_{in}}{R_{out}}} \quad . \quad . \quad . \quad . \quad . \qquad (5)$$

which reduces to:

$$\frac{G_m \sqrt{R_{in} R_{out}}}{2} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \qquad (6)$$

The over-all voltage gain as measured in the indicated circuit would then be the product of input tuner gain and realizable tube gain. This equation was developed from a voltage analysis but may also give a measure of power gain inasmuch as the impedance level is the same at input and output. From a power standpoint the input tuner contributes nothing and the power gain is given by the square of Eqt. (6) above.

The bandwidth to be expected cannot be evaluated as readily because of the loading effects of the tube impedances and the capacitances inherent in the

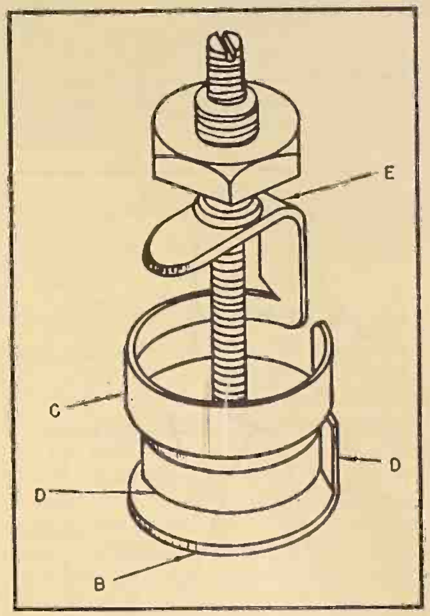


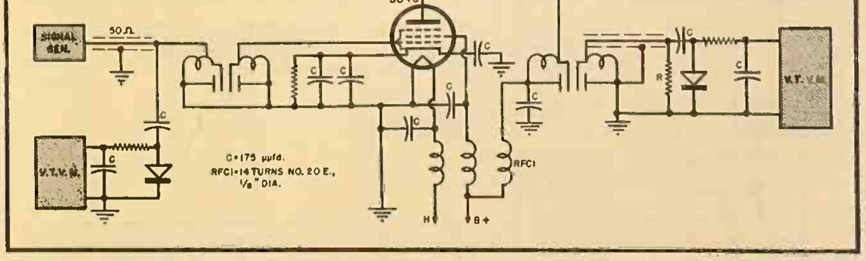
Fig. 4. Drawing of the u.h.f. tuner.

matching networks. However, an approximate calculation of bandwidth may be arrived at if several assumptions are allowed. In practice many of these assumptions cannot be met, but they are useful in simplifying the analysis and make possible the determination of at least the order of magnitude of the expected bandwidth.

1. Assume that the input circuit bandwidth is sufficiently broad that its effect on the over-all response is negligible.
2. Assume equal Q's in the primary and secondary circuits of the output network.
3. Assume equal division of the total shunt capacitances; half on the primary and half on the secondary of the output network.
4. Assume critical coupling in the output network.

Under these conditions the gain-

Fig. 5. Circuit diagram of the 400 megacycle amplifier.



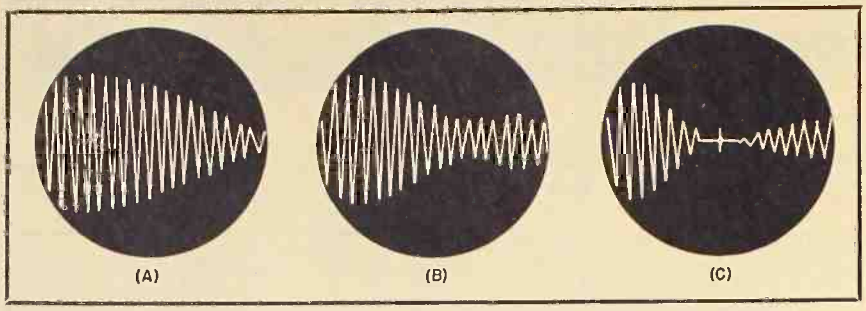


Fig. 6. Impedance match presentation for (A) unterminated delay line, (B) terminated by amplifier with mismatched input (C) amplifier tuned for matched input.

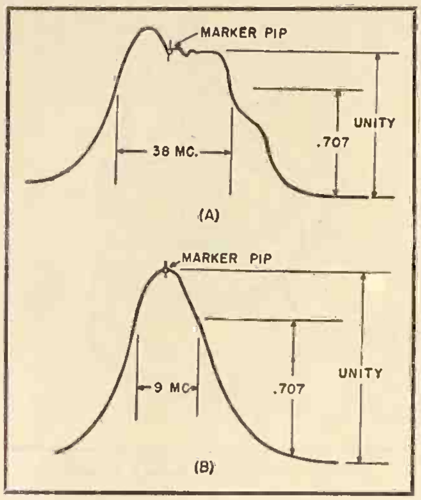


Fig. 7. (A) Input bandwidth for type 5840 at 400 mc. (B) Output bandpass characteristic for type 5840 in a single stage 400 mc. amplifier.

bandwidth product for a double tuned circuit may be shown to be $G_m/\sqrt{2\pi}\,C_T$. Substituting the values for the type 5840 gives a gain bandwidth product of approximately 89 x 10^6 when the socket and circuit capacitances are included. Using the grid-to-plate tube gain as calculated from input and output resistance considerations gives a bandwidth of approximately seven megacycles between -3 db. points.

Any analysis of this nature is not complete without some investigation of the amplifier stability. In the consideration of theoretical safe gain for a u.h.f. amplifier, it must be borne in mind that the feedback path is no longer confined to the grid-plate capacitance. At u.h.f., feedback occurs through many paths due to capacitances, and self or mutual inductances of the leads. A detailed discussion of the grid-plate feedback reaction is found in the basic paper of M.J.O. Strutt and A. Vander Zeil in the Proceedings of the I.R.E. for August, 1938. It will be sufficient here to point out the conclusions derived from this article.

It is shown by Strutt and Vander Zeil that the grid-plate impedance is reactive and of the form:

$$Z_{gp} = \frac{1}{j\omega\ (C_{gp} - A\,\omega^2)} \quad . \ . \ . \ . \quad (7)$$

where C_{gp} is the low-frequency, grid-plate capacitance and A is a constant dependent upon tube construction. It can be seen immediately that for large values of ω there exists the possibility that the effective grid-plate reactance may appear inductive. There will also be some value of ω at which the tube can exhibit self-neutralization. Therefore, it is evident that grid-plate capacitance as usually measured on a 456 kc. bridge cannot be taken as a criterion of stability above some high frequency determined by the tube construction and the constant A. In subminiature pentodes of the 5840 construction, the frequency of self-neutralization is above 200 megacycles. At higher frequencies the feedback increases again but is inductive in nature. The feedback impedance at any one frequency above the crossover point can be represented as an equivalent negative capacitance. This equivalent negative capacitance is now a function of frequency and at 400 megacycles is approximately equal in magnitude and opposite in sign to the low frequency grid-plate capacitance.

It was this coincidence which led to the use of the conventional safe gain formula as an approximation at 400 megacycles:

$$\text{Theoretical Safe Gain} = \sqrt{\frac{K\,G_m}{2\pi f\,C_{gp}}} \ . \quad (8)$$

where K is a function of the number of stages and is 2 for 1 stage, 1 for 2 stages, 0.86 for 3 stages and 0.67 for 4 stages.

The application of this reasoning to multiple-stage design has led to regeneration indicating that such safe gain calculations are overly optimistic.

For the type 5840 this formula gives a safe figure of 14 for single-stage voltage gain. This figure is very close to the previously calculated grid-to-plate tube gain, Eqt. (4), before the output step down, indicating that regeneration may be encountered if care is not exercised in the amplifier layout and wiring.

Amplifier Layout and Wiring

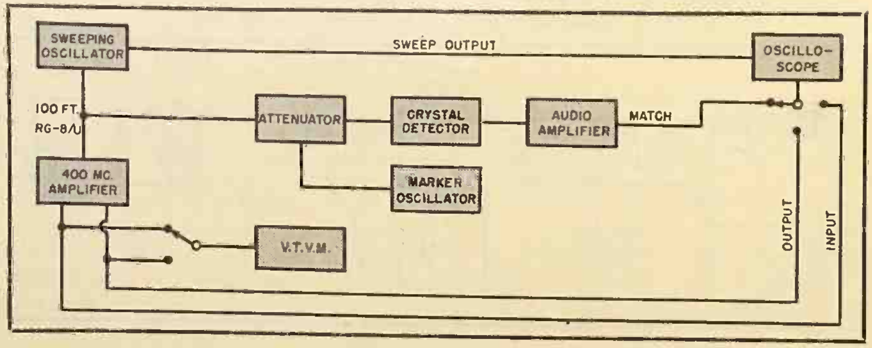
The u.h.f. tuners used in the matching networks are more conveniently coupled by capacitances than by mutual inductance. The coupling network reduces to a pi-type with the parallel elements composed of tuners and the series element a coupling capacitance. The coupling capacitance was accomplished by mounting the tuners end-to-end. The tuner on the movable bracket can be shifted to change the coupling capacitance. A pi-type network of this nature can be made to match impedances differing by ratios greater than 10 to 1, which is the order of match required in this application.

The circuit design for use at 400 megacycles is conventional in other respects. Due care was exercised in the arrangement of parts to keep all leads short, to isolate input and output circuits, and to utilize separate grounds for the input and output cathode returns. The tube type 5840 is provided with three cathode leads, and it has been found that proper connection of all three leads is required to avoid instability in the amplifier. Two cathode leads were bypassed to the input side of the compartmented chassis and one cathode lead was bypassed to the output side of the chassis. The tube basing is such as to make this possible with a minimum of lead length.

All leads carrying r.f. were made of ⅛-inch x 0.005-inch silver plated copper ribbon. Silver mica condensers of the button type were used for all bypassing. A circuit diagram for the sin-

(*Continued on page* 30A)

Fig. 8. Block diagram for determining the input conditions of the amplifier.



STRAIN GAUGE LOAD RINGS

Typical load ring with cover partially removed.

By ALVIN B. KAUFMAN

Circular load rings have some advantages over straight links in force measuring equipment.

IN considering the design of force measuring equipment there is always the question of whether to use a straight link design or circular load ring. Each possesses certain characteristics making its use desirable for some applications and not others.

Strain gauge links may be fabricated to allow direct meter operation* and electrically may have two active and two dummy gauges mounted on the link. Physically they are constructed long and narrow and are capable of being used in very close quarters.

A load ring has the mechanical disadvantage of requiring more room in one plane than a link, but may have a high safety factor with the simple addition of a bolt fixed so that if the link breaks, the bolt will assume the tension or compression load. It also possesses an electrical advantage in as much as all four legs of the strain gauge bridge are active gauges. Those on the inside of the ring have the opposite *dR* sign than those on the outside of the ring. This may be seen clearly by examining the forces imposed on a ring when an attempt is made to pull it out straight, from points 180 degrees apart. As the ring straightens out, the outer surface is compressed as it must become shorter, while the inner surface lengthens to meet the outside length.

The design of a circular load ring need not be any more complicated than that of a link, once the correct formula has been determined. The mathematics leading up to these formulas is complex and may be inspected by those mathematically inclined, in "Formulas for Strain and Stress" by Roark, or in other texts. The formulas used herein will be basic and applicable to the job at hand.

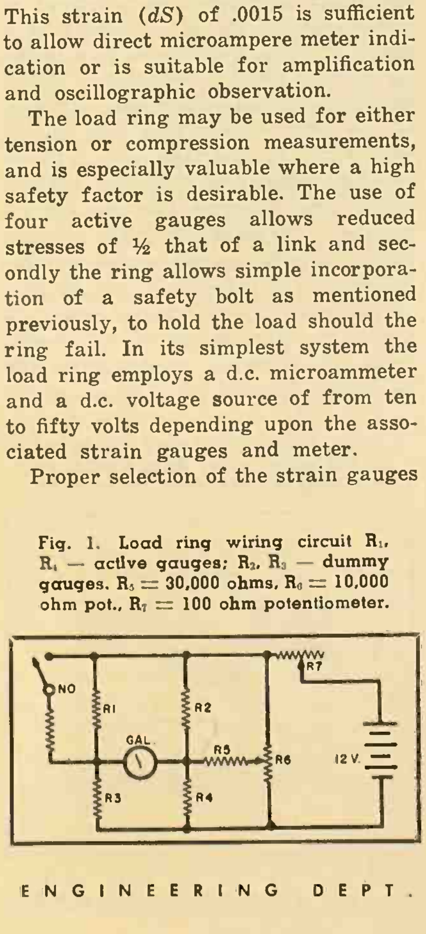
Unlike a link, a force ring will not indicate bending or side loads from any plane. This does not require any special placement of the gauges, or wiring, but occurs due to the construction and natural phenomena surrounding this device.

Generally speaking, rings and links should preferably be designed so that the maximum expected load will give a stress of about 15,000 lbs. per sq. in. (p.s.i.) in aluminum or 45,000 p.s.i. in steel. There values allow the use of practically any steel or aluminum alloy. Where these p.s.i.'s are exceeded, special steels or heat treated steel or aluminum alloys might be necessary. This value of stress will result in a strain of about 15x10^{-4} in these materials. This strain (*dS*) of .0015 is sufficient to allow direct microampere meter indication or is suitable for amplification and oscillographic observation.

The load ring may be used for either tension or compression measurements, and is especially valuable where a high safety factor is desirable. The use of four active gauges allows reduced stresses of ½ that of a link and secondly the ring allows simple incorporation of a safety bolt as mentioned previously, to hold the load should the ring fail. In its simplest system the load ring employs a d.c. microammeter and a d.c. voltage source of from ten to fifty volts depending upon the associated strain gauges and meter.

Proper selection of the strain gauges will vary depending upon the use of a direct indicating meter or an amplification system. In the direct indicating system it is imperative to have the galvanometer internal resistance equal the parallel series resistance of the bridge, for maximum indication, whereas with an amplifier practically any value may be satisfactory. This parallel-series resistance, where all four legs are essentially the same value, may be assumed to equal the value of resistance of one leg of the bridge. Even with full strain on the ring these resistance values will not vary more than approximately one per-cent, therefore this condition is met satisfactorily. Regardless of the gauge used, maximum voltage output or galvanometer indication will occur with maximum current through the bridge. This is apparent when it is noted that the output voltage is equivalent to *dR* (the total change in bridge leg resistances) times the current through a gauge (one-half total bridge current). This $I_g dR_g$ will, with 75 to 120 ohm gauges and 10 volt bridge supply, supply about 30 millivolts at a strain of 15x-10^{-4} in./in. With a one milliampre meter this would supply half scale deflection, or about 500 μa. output with the conventional 50 ohm, 1 ma. meter. It would be preferable to use a microammeter and less strain to secure a full scale indication, as will be discussed further along.

Fig. 1. Load ring wiring circuit R_1, R_4 — active gauges; R_2, R_3 — dummy gauges. R_5 = 30,000 ohms, R_6 = 10,000 ohm pot., R_7 = 100 ohm potentiometer.



As has been indicated in many texts, a mismatch of two or three to one of a load to source (or galvanometer to bridge) will affect the maximum transfer of power or current by approximately 20 to 30%. However, another factor enters the discussion where this situation exists on a bridge. As an example, changing strain gauges from 100 to 200 ohm gauges, when used with a 100 ohm meter, would cause a meter indication decrease of about 30%. Were the meter a 400 ohm meter a decrease of approximately 60% would

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*Kaufman, Alvin B., "Strain Gauge Link," RADIO-ELECTRONIC ENGINEERING, March, 1950.

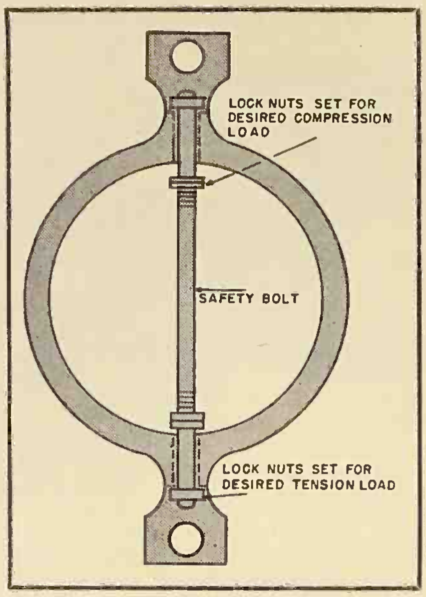


Fig. 2. Safety bolt installation.

occur, where the bridge d.c. voltage is not altered. This is due to a second factor entering into the situation. If the bridge leg resistance values are doubled, then the total bridge circuit and leg currents must fall to one-half their previous values. This in itself will not cause a change of output voltage, for as the current falls to one-half, the dR_g increases to twice the previous value. Thus the dE_o is independent of gauge resistance values, assuming the same gauge factor and bridge voltage. However, $I_g = dE_o/Z_o$ and for a given bridge voltage as the Z_o of the bridge increases the galvanometer current will fall off. As Z_o is equivalent to R_g plus the galvanometer resistance, this formulation is correct. Of course the bridge voltage could be raised to twice the previous value, before exceeding the gauge wattage rating. This is only satisfactory up to the point where excessive wattage will not be dissipated in the gauges. The major reason for pointing out the above is to eliminate design errors where the bridge supply voltage cannot be altered for technical or economic reasons, and to indicate that loose thinking must not prevail on this matching problem. With amplification of the signal this problem is not critical.

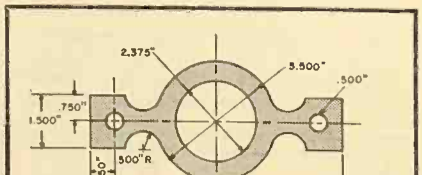
The wattage capabilities of the different strain gauges depends mainly upon the thermal conducting surface upon which they are mounted. With thin aluminum or steel a valve of ½ watt is satisfactory. With heavier aluminum, copper etc., a higher value, possibly one watt, is satisfactory. Where many gauges are mounted closely, as a dummy plate or load ring, it is preferable to play safe and use the lower wattage value.

Calculation of the ring dimensions for specific direct meter indications will be discussed after formulas are presented for the calculation of amplifier-type load rings. The amplifier load ring, with a strain of 15×10^{-4} in./in, will in general be satisfactory for direct μa. meter operation (where meter *R* equals gauge *R*) giving such high readings that the meter must be shunted or a series resistance added to cause on scale operation for full ring load conditions.

In designing load rings there are two criteria that must be remembered. The formulas are correct only for thin rings with the average ring diameter at least five times the thickness of the ring. Any lower ratio causes the formulas to be usable, but not very accurate. Secondly, the end fittings cause the stress to be less than the calculated value. For all reasonable designs, this calculated stress will be within a few per-cent of the measured value. For design purposes, however, it is advisable to calculate for slightly higher stresses than required (possibly 5%) and make calibration runs against precision measuring equipment.

A typical view of a load ring is shown in Fig. 4 and the photographs accompanying the article. The stress (p.s.i.) set up in the ring at section *x-x* is indicated by the formula $S = 1.09\ WR/bd^2$ where *W* is the applied load in pounds, *R* is the average radius, *b* is the width of the ring, and *d* is the thickness; all dimensions measured in inches. Having already decided that a stress of 15 to 45,000 p.s.i. is a working value, then it is desirable to factor the formula and obtain the thickness of the ring, *d*, assuming that radius and width of the load ring will be selected empirically and with the aid of the tables accompanying this article. Thus $d = \sqrt{1.09\ WR/Sb}$. The average radius will be to the center of this thickness, and therefore when the ring is machined it will be necessary to subtract or add one-half of this amount from the *R* radius to determine the machining i.d. and o.d. This is not too critical where the ring thickness is a small percentage of the radius, otherwise it can cause a high error where neglected. Technically the average radius is not to the center of the ring's mass, but is close enough to make other complicated calculations unnecessary. Cable or rod attachment requires machined bosses on the ring, 180° apart. The ring should not be drilled for bolts or other attachments as this will weaken the ring and cause premature failure at relatively low loads. The bosses should be the same width and thickness as the width of the ring at the root. The boss height is not critical, but its radius into the outer surface of the ring should be one-eighth inch or more; preferably the thickness of the ring. The width of the ring should be one-quarter inch or more to minimize any side load causing destructive bending. These side loads would not be indicated to any great extent by the recording equipment, but might build up to the point of destroying the ring. Installation of the strain gauges should be at the stress points *x-x*, two on the inside of the ring and two on the outside of the ring. The inner pair of gauges will have the same sign of *dR* but opposite that of the outer pair. The gauges are wired into a conventional Wheatstone bridge as shown on the basic schematic drawing. Of course the outer and inner gauges are across from each other so that their positive and negative dR_g will add together and give increased bridge output voltage rather than cancel out (being equal but opposite in sign).

Fig. 3. Mechanical design of a typical 10,000 pound ring. Steel should not be plated where gauges are installed.



The strain gauges are installed on the ring by degreasing the area to be used and then gluing with nitrocellulose or bakelite cement according to the type of gauges used. Full installation details and a price list of SR-4 strain gauges will be supplied by *Baldwin-Southwark Locomotive Co.*, Philadelphia, Penna., upon request. It is imperative when ordering gauges to be certain that they will fit the load ring. When the ring is but one-quarter of an inch wide, special small gauges are required; for wider rings, one inch or over, practically any strain gauge will fit.

As mentioned previously the ring force indications are not appreciably affected by bending or side loads, and this same configuration also automatically causes the bridge to be temperature compensated. With all gauges mounted on the ring any temperature change affects all four legs equally and the dE_o remains constant. A slight variation might be noted where a galvanometer is used, but as any bridge resistance change would be a small percentage of Z_o, the error is negligible.

The bridge output voltage may be

Typical load ring showing the active and dummy gauges.

calculated from the formula $dE_o = 2R_g\, I\, K\, dS$. This formula applies only to a bridge with four active legs; where R_g is the resistance of one gauge, I is one-half the total bridge current (ie. gauge current), K is the gauge factor (supplied by mfr.), and dS is the strain in inches/inch. Strain $= P/AE$ where P is the load in pounds, A is the area, and E is the modulus of elasticity. Then P/A is stress or p.s.i. With the stress already agreed upon at 15,000 for aluminum and 45,000 for steel, then strain calculates out very close to .0015 in./in. A simplification of the above formula is $dS =$ Stress/E. This dS figure used in the dE_o formula will give the expected output voltage of the ring under full load. The galvanometer current that may be expected for this same load is dE_o/Z_o where dE_o is the bridge's unloaded voltage output as determined above, and Z_o (where all four bridge legs are the same value) equals the resistance of one bridge leg plus the galvanometer's internal resistance, *i.e.*, $Z_o = R_g + R_{gal}$.

The use of two active and two dummy gauges with the ring, while lowering the dE_o by one-half, has certain mechanical advantages. This loss of gain is not critical, in any case, where amplification is used. It allows installation of all four gauges on the inside of the ring where they may be wired together and connected to a receptacle mounted on one of two shields installed to cover the hole in each side of the ring. These covers then protect the gauges and wiring and by holding the disconnect receptacle provides quick connection facilities. The covers may be secured at one end only, the other end being slotted to allow free movement of the ring. For this type of ring bridge the radius will be the inside radius of the ring, not the average radius used in calculations with the four active gauge ring. The dE_o for this bridge is $dE_o = R_g\, I\, K\, dS$. The installation of dummy gauges at the boss ends will not allow the installation of a safety bolt.

When the load ring is to be designed specifically for use with an indicating meter, then another series of calculations is necessary to insure that sufficient strain and stress exist in the ring to cause full scale deflection of the galvanometer. The required resistance change for full scale deflection of the galvanometer is:

$$R_M = \frac{I_b XX - 2\,(I_g XX) - 3\,(I_g\, R_g X)}{I_g\, R_g + 2(I_g X) + I_b X}$$

where I_b is the total bridge current, X is the resistance value of one leg of the bridge, I_g and R_g are respectively the galvanometer full scale current (in amperes) and internal resistance. This formula, however, gives the change in resistance of one leg of a bridge with only one active leg, the other three dummies. As our bridge has *two* to *four* active legs and it is necessary to know the *change* in resistance of these two or all four legs for this indication, then dR_g (per gauge) $=$

$$\frac{R_{static}\text{-}R_M}{\text{No. of Active Gauges.}}$$

Knowing dR_g, it is a simple matter to find the required strain on a gauge: $dS = dR_g/R_g K$ where K is the gauge factor and dR_g is the change in gauge resistance, with R_g the static resistance value of the gauge.

Gauge factor is the ratio of gauge resistance change to strain:

$$K = \frac{dR_g/R_g}{dS}$$

This factor is supplied by the manufacturer and need not be determined. Normally the gauge percentage change of resistance is several times that of the strain and it is quite common to find gauge factors of 1.8 to 2.1.

After calculating dS, stress must be determined so that it may be substituted into the previous formula and the correct load ring thickness determined. Stress may be calculated from the formula: $S = dS\,E$ where S is stress (p.s.i.), dS is strain, and E is the modulus of elasticity.

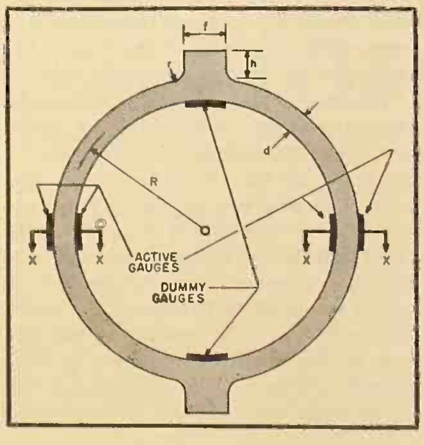
Where this value of S turns out to be smaller than the value given in the tables, then the table design information can be used, the ring putting out more dE_o than required. If this S value is higher than the 15 or 50,000 p.s.i. used for the rings calculated in the tables then it will be necessary to calculate the new thickness required for your given configuration.

Microammeters are suited perfectly for galvanometer use in a direct indicating system. Full scale should be 100 to 500 μa. with an internal resistance of 75 to 240 ohms being preferred. This value of internal resistance allows matching of the meter to any of the three most common strain gauge resistance values of 75, 120, or 240 ohms. Of course, as previously indicated, the meter should equal the gauge resistance or higher bridge supply voltage will be required to counter the reduction of signal due to the mismatch. Even where the meter impedance matches that of the bridge output the higher resistance gauges and meter will give less output for a given strain and bridge voltage. This is seen by $I_g = dE_o/Z_o$. For a given strain and gauge factor, dE_o will remain constant regardless of strain gauge values (at a fixed bridge voltage). However Z_o varies directly as the gauge resistance and the galvanometer resistance. Thus for a given bridge voltage the 240 ohm gauges and associated matching meter will give about one-third the scale indication for a given strain that the 75 ohm gauges would produce. All of this, of course, assumes the use of a 0-100 μa. meter of varying internal resistance. If the bridge voltage were increased twice (maximum allowable for $P = E^2/R$) thus doubling dE_o, the output would still be but a maximum of two-thirds that of the 75 ohm strain gauge bridge.

A most suitable instrument for the direct indicating system is the *General Electric* DP-9 0-100 μa. meter, with 75 ohms internal resistance. Its accuracy

(*Continued on page* 30A)

Fig. 4. Electromechanical design.



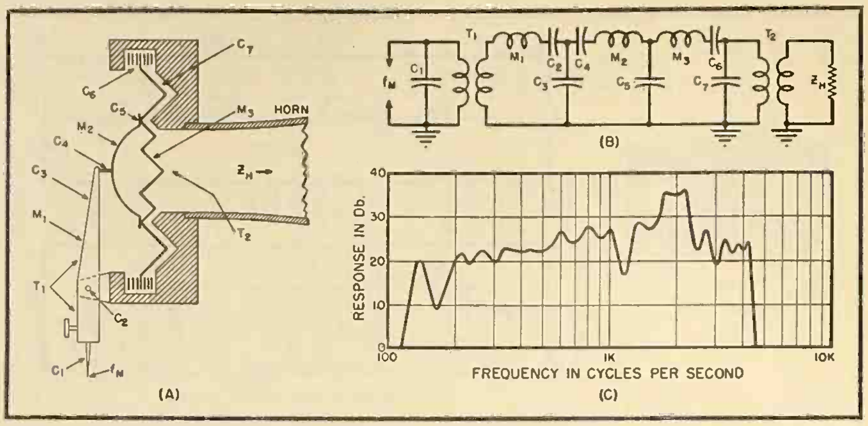


Fig. 5. (A) Mechanical details of orthophonic phonograph. (B) Equivalent electrical circuit. (C) Response of reproducer to standard test record (constant amplitude below 700 c.p.s., constant velocity above 700 c.p.s.).

made this unit an outstanding advance over previous accoustical types.

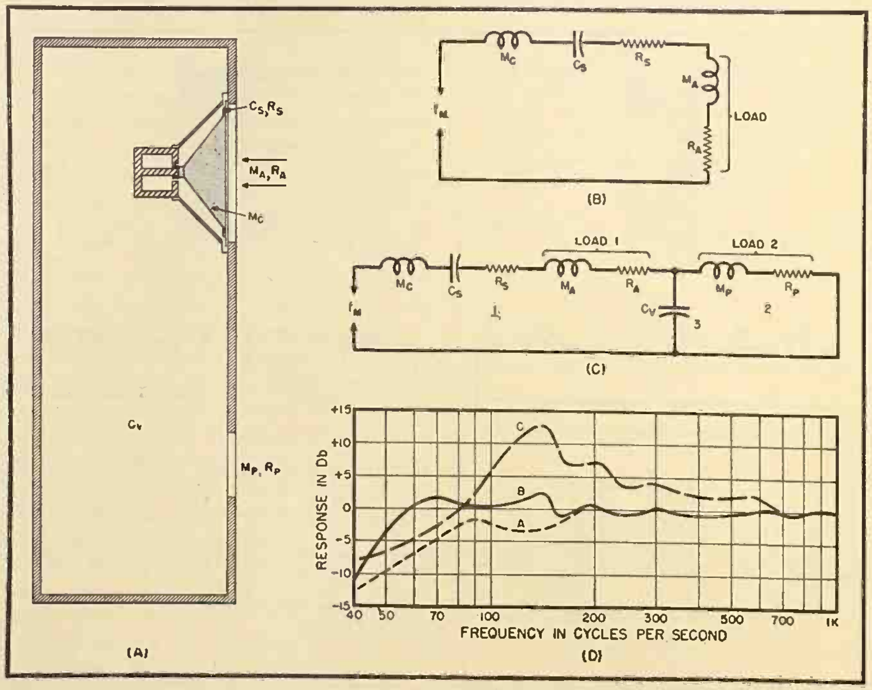
These same principles are also utilized in the design of loudspeaker cabinets. For example, the acoustical phase inverter (or "bass reflex") system consists of a direct radiator loudspeaker mounted in a completely closed cabinet which has an opening to couple the cabinet volume to the air, as shown in Fig. 6. The equivalent circuit in B makes it easier to understand the operation of this system. The effect of the cabinet is to couple the parallel-resonant circuit consisting of C_V (the compliance of the cabinet volume), M_P (the mass of the air in the opening), and R_P (the mechanical resistance of the air load on the opening) to the back of the loudspeaker.

The phase of the velocities on the two sides of the cone differs by 180°. From the circuit it can be seen that the currents in the branches 1 and 2 may differ by as much as 180° when there are positive reactances and no resistances in branches 1 and 2, and a pure reactance in branch 3. The phase angle will be reduced as resistance is introduced. In loudspeaker systems the mechanical resistance is small compared to the reactance, and the constants can be chosen so that the phase angle between the velocity of the cone and the port is very small. The type of frequency response which is obtained with such a system is shown in D. For comparison, there are also shown the response of the loudspeaker mounted in an infinite wall (to eliminate the effects of interference between radiation from the front and the back of the cone), and the response obtained from the conventional open-back radio console cabinet. These curves show that performance can be obtained which is much superior to the conventional cabinet, without the space requirements of the infinite wall baffle.

In these, as well as in many other cases, the method of electromechanical analogies has made it possible to understand fully the operation of acoustic systems and devices, and to know in advance what the characteristics of any new design will be. After the preliminary design has been put on paper, the equivalent circuit can be drawn from a knowledge of the physical dimensions of the various components and from the properties of the materials used—such as density, elasticity, tensile strength, viscosity, etc. Of course, not all of these factors can be calculated exactly,—especially the frictional losses in the system—but an indication of their magnitude can usually be obtained. The general procedure is to design the unit and analyze its operation first on the basis of the equivalent circuit, then, when a satisfactory design has been arrived at, to contruct the unit and measure its performance. This is then compared with the calculated or measured performance of the equivalent circuit, and any of the circuit components can be corrected to make them correspond with the model. The effects of changes in the mechanical design can be determined simply by changing the values of the appropriate electrical components and observing the changes in the response of the circuit, until the optimum performance is attained. It will then be known what changes are necessary in the mechanical design (for example, certain members made stiffer, others made lighter or heavier, more or less damping introduced, etc.), and the final unit can be built accordingly.

Since its introduction this procedure has been an extremely valuable method of electromechanical and acoustic design. It is an extremely important and practical design tool, since it results in product designs that mean greater dependability and improved performance, and gives precise manufacturing information that permits better control of quality during production. The initial introduction of this method resulted in outstanding advances over previous acoustic designs, and its applications at the present time are making it possible to make continuous progress toward the goal of perfect fidelity of sound reproduction.

Fig. 6. Electromechanical networks in design of bass-reflex (acoustic phase inverter) loudspeaker cabinet. (A) Loudspeaker mounted in a bass-reflex cabinet. (B) Equivalent electrical circuit of speaker alone. (C) Equivalent network of speaker in cabinet. (D) Typical response of loudspeaker in bass-reflex cabinet B, compared with response of speaker alone A, and conventional open-back cabinet C.



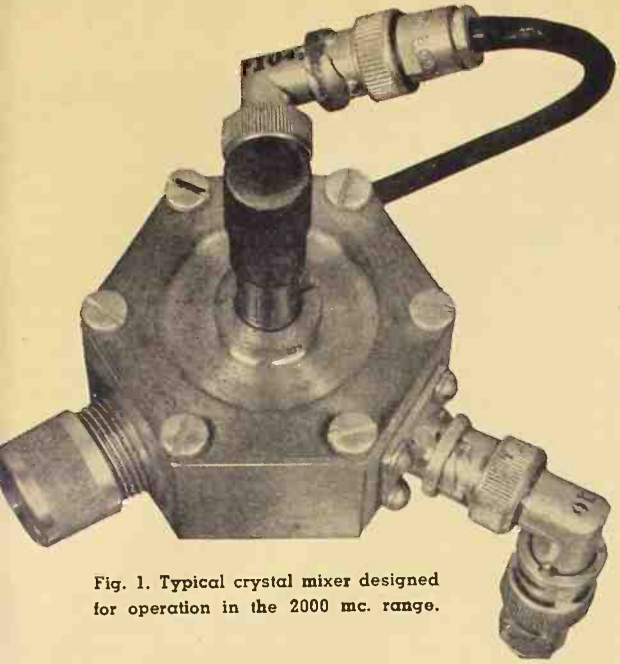
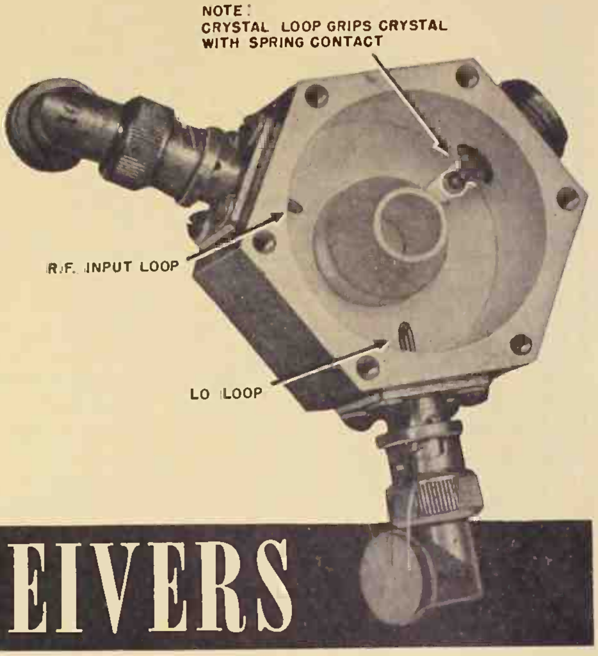


Fig. 1. Typical crystal mixer designed for operation in the 2000 mc. range.

Fig. 2. Interior view showing r.f., local oscillator and crystal loops. The mixer employs a reentrant cavity and micrometer type tuning control.



MICROWAVE RECEIVERS

By J. RACKER

Federal Telecommunication Laboratories

Part 2. A discussion of various types of mixers and local oscillators for use in the microwave range.

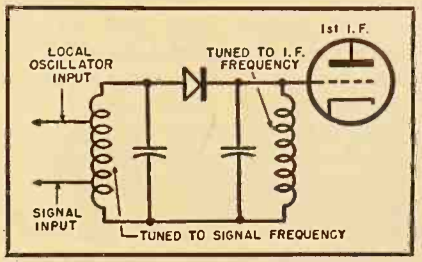
IN the first article on this subject the design of r.f. amplifiers, and the importance of the noise factor in microwave receivers, were covered. In this article we are concerned primarily with the design of the mixer and local oscillator stages. It was previously shown[1] that the over-all noise factor of the receiver is primarily a function of the noise figures of the first several stages and that, at microwaves, little or no improvement can be effected through the use of an r.f. amplifier. Consequently in most commercial receivers, the antenna signal is fed through an appropriate filter directly to the mixer or frequency converter (as it is sometimes called) stage.

Mixer Stage

The technique of converting a high frequency signal to a lower intermediate frequency to obtain greater ease of amplification and frequency selection is particularly important in microwave applications because of the difficulty of amplification. Although the same fundamental principles are employed as at conventional frequencies, converters operating in the centimeter band have very little physical resemblance to those previously used in superheterodyne circuits. It is interesting to note that the crystal detector which was used universally in early radio receivers, then completely abandoned with the development of the vacuum tube, has now reemerged as an important element finding widespread application in microwave receivers.

The mixer, in its simplest form, can be represented by the block diagram shown in Fig. 3. The signal from the antenna is fed to a matching network at the input to the filter. The local oscillator output is placed in series with this signal voltage and the combined output is applied across a rectifying circuit and output network. The effect of the nonlinear rectifying element is to produce harmonics of both input signals, as well as difference and addition signals of both. Generally, the difference between the two fundamental signals is used as the i.f. frequency, though in some cases the second harmonic of the local oscillator is employed. In many receivers a 30 megacycle i.f. is used, though for special applications or for equipment operating at the top band in the microwave range (at about 10,000 mc.) higher i.f. frequencies are utilized.

Fig. 3. Simplified schematic diagram of a crystal mixer.



The conversion factor of the mixer is defined by the following equation:

$$G = E_{if}/E_s \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

where E_s is the amplitude of the input signal

E_{if} is the amplitude of the i.f. signal at the output of the mixer

This factor will be greater than 1 when a triode or pentode mixer is used, due to the amplification of the tube, and less than one when a diode or crystal is employed.

The conversion factor of the mixer can be determined through the use of Fourier series analysis. If the rectifying element has the ideal characteristic shown in Fig. 4A, (i.e. the conductance $g = d_{ib}/d_{eb}$ is constant for positive value of the plate voltage e_b and zero for negative values) and the local oscillator signal amplitude is much greater than the signal voltage (which is readily realized in practice) then the conversion gain can be shown to be [7]:

$$G = \frac{gZ_L \sin n\theta}{\pi n} \quad . \quad . \quad . \quad . \quad . \quad (2)$$

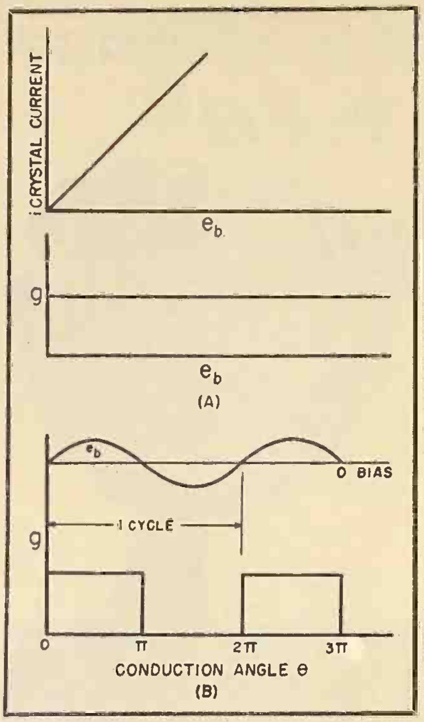


Fig. 4. Curves showing characteristics of an ideal rectifier element.

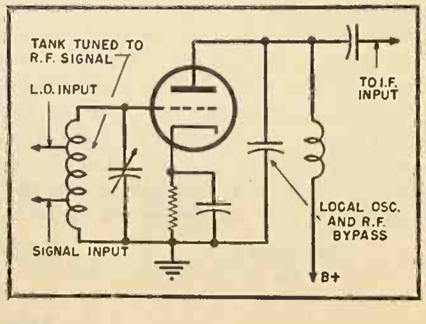
where n is the harmonic of local oscillator used
θ is the conduction angle for one cycle. For example, the conduction angle as shown in Fig. 4B is π
Z_L is the load impedance at the i.f. frequency

Maximum conversion gain occurs when $n = 1$, (i.f. frequency equal to local oscillator frequency plus or minus carrier frequency) and $\theta = \pi/2$ (no bias voltage on rectifier). In this case, the conversion gain becomes:

$$G = \frac{gZ_L}{\pi} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (3)$$

From the foregoing it is apparent that maximum gain is obtained when the fundamental of the local oscillator is used, Z_L is made as large as possible consistent with bandwidth requirements (Z_L varies inversely as the bandwidth so that for wider bandwidth the conversion gain decreases), and no bias voltage on the rectifier is used.

Fig. 6. Equivalent circuit diagram of a typical lighthouse mixer.



There are two carrier frequencies that will provide a given i.f. for a particular local oscillator frequency, f_{LO}, i.e., $f_{LO} + \text{i.f.}$ and $f_{LO} - \text{i.f.}$ One of these frequencies is the signal to be detected, while the other is known as the image frequency. In applications where other units may be operating in the same frequency range, image frequency rejection is an important design factor.

Image frequency rejection is usually effected in the input circuit, which is a resonant coaxial line or cavity—depending upon frequency of operation. This input circuit should also be designed to provide a good match to the antenna—otherwise part of the signal voltage will be lost. This is particularly important when diode or crystal mixers are used, where G is less than 1, since the effective noise figure of the succeeding stage is equal to its noise figure multiplied by $1/G$ (G of mixer). Hence signal loss in the input circuit not only decreases the noise figure of the mixer but affects the following stage by a factor of $1/G$.

Selection of the rectifying element in the mixer is primarily a function of the resulting over-all receiver noise figure. A choice must be made between a tube mixer and a crystal mixer. Although the gain of the crystal mixer is much less than that of a tube mixer, the noise introduced by a mixer is also very much less, so that for frequencies above 3000 mc. an improved noise factor can always be obtained with a crystal mixer. In fact virtually all receivers designed for operation at frequencies above 3000 mc. do use crystals. Another advantage of the crystal is the circuit economy and simplicity attained, which accounts for the fact that most commercial receivers under 3000 mc. also use crystals. These advantages of the crystal must be balanced against its disadvantages of unreliability unless carefully handled (they may require replacement every six months to maintain best performance) and difficulty of antenna isolation.

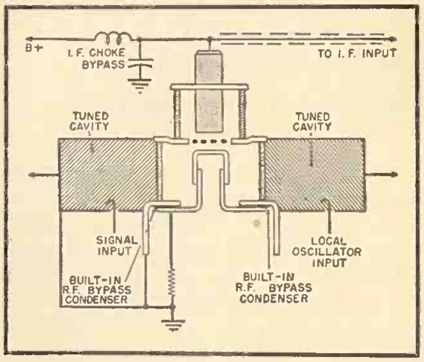


Fig. 5. Diagram of lighthouse mixer.

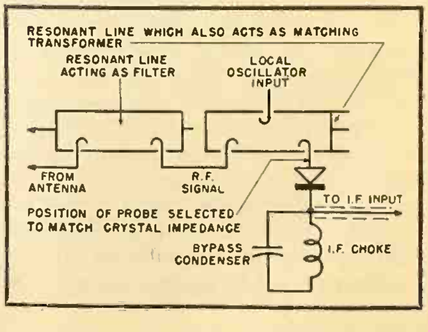
Fig. 5 shows the schematic diagram of a typical lighthouse tube mixer, while Fig. 6 shows the equivalent schematic of this circuit. The tube is biased near cut-off by the self-biasing resistor in the cathode circuit and then driven relatively hard by the injected local oscillator voltage. It should be noted that the voltage drop across the biasing resistor results from the average cathode current, not from the quiescent current. The bias is, of course, not maintained at cut-off with the local oscillator turned off. The increase in average plate current, upon application of the local oscillator voltage, is sometimes used as an indication of correct local oscillator operation and coupling.

The grid-to-cathode resonant line of this mixer is tuned to the r.f. signal. A sufficiently high Q line is used to reject the image frequency (should be about 60 db. down) and low enough Q to pass the bandwidth required. The r.f. signal is coupled to this cavity near its shorted end. A trial and error procedure is usually employed to determine exact position and amount of coupling that will result in a minimum standing wave ratio on the antenna line and maximum i.f. signal in the plate circuit. The local oscillator is loosely coupled into the cavity—usually much further away from the shorted end than the signal input. The plate output is fed directly to the i.f. input through a short (low capacitance) length of coaxial line, with the plate connecting to the inner conductor of this line.

The question sometimes arises as to whether it is preferable to use a given tube as an r.f. amplifier or as a mixer. This question is resolved primarily upon the amount of signal and noise output produced by the tube involved. Although the gain of a tube used as a mixer is only about one-fourth of the gain obtained with the same tube as an r.f. amplifier, the noise output is reduced considerably because current flows only during one-half of the cycle. Therefore, the advantage of using a tube as an r.f. amplifier rather than a mixer is not as great as may be expected. In addition, any advantage of using an r.f. amplifier must be weighed against the inconvenience of tuning r.f. stages.

The most sensitive mixer for micro-

Fig. 7. Simple crystal mixer.



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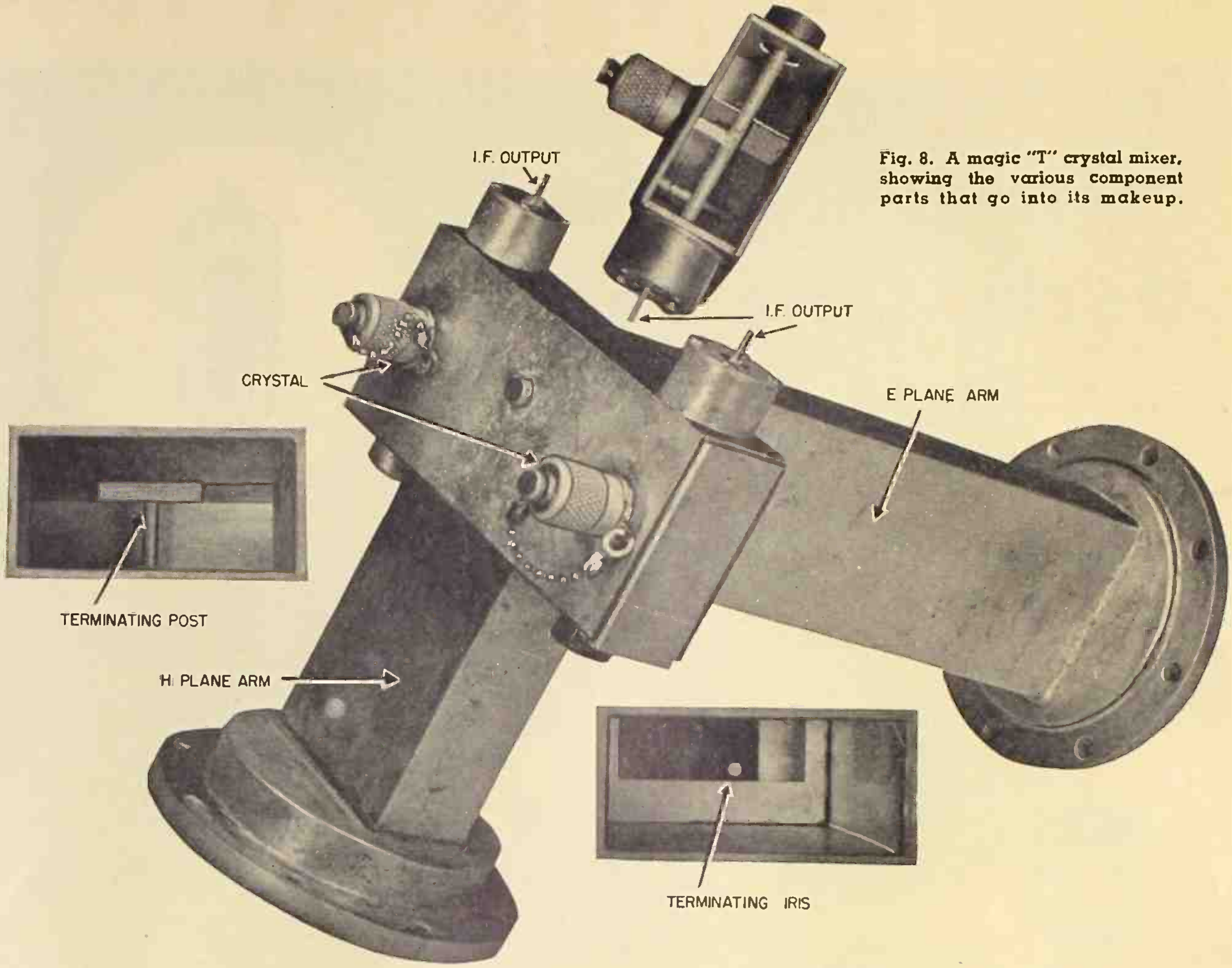
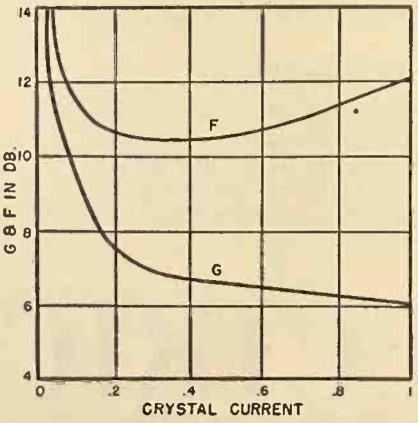


Fig. 8. A magic "T" crystal mixer, showing the various component parts that go into its makeup.

waves above 3000 mc. is the crystal type in which a small contact between a metal whisker and a semiconductor material, such as silicon or germanium, causes rectification of radio frequency signals. For mechanical protection the semiconductor and metallic "catwhisker" are mounted in a cartridge unit, which is presently manufactured so that it is nonadjustable.

Crystals manufactured for use as mixers are numbered in accordance with their frequency range and sensitivity. For example 1N21 crystals operate at about 3000 mc. while the designation of 1N21, 1N21A, and 1N21B indicates poor, medium, and high sensitivity respectively. The initial step in crystal mixer design is to determine which type best meets the individual requirements of the receiver involved.

Fig. 9. Typical crystal current vs. noise and gain curve.



The important characteristics of several typical crystals are given on the following page, Table I.

In designing a crystal mixer the following factors must be considered: 1) Maximum r.f. signal (match between antenna and input), 2) Rejection of unwanted signals (this includes transmitter output, image frequency, and prevention of local oscillator output from reaching antenna), 3) Matching r.f. input to crystal, 4) Minimizing the effects of the local oscillator coupling on mixer cavity matching, 5) Minimizing injection of local oscillator noise on signal, and 6) Matching crystal output to i.f. input with provision for bypassing r.f. and local oscillator fundamental and harmonic frequencies.

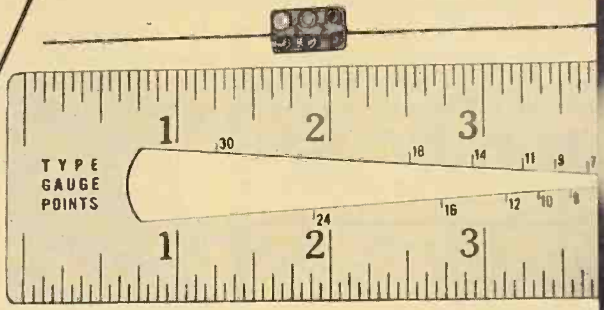
The simplest type of crystal mixer circuit is the one shown in Fig. 7. In this case an antenna filter preselector is used to permit proper image and transmitter output rejection consistent with bandwidth requirements. This filter is usually a simple coaxial or waveguide cavity, depending upon frequency of operation, which is tuned to the r.f. frequency. The antenna input coupling is adjusted to provide maximum output and minimum standing wave ratio.

The output is then fed to the mixer cavity. This cavity, together with the antenna filter, should provide necessary attenuation of the undesired signals. The mixer cavity also acts as a transformer, matching the r.f. input impedance to the crystal. The local oscillator input, usually very loosely coupled, is also injected into the mixer.

Trial and error procedures are generally used to effect the best mixer conditions. The r.f. input loop is adjusted, by rotation and positioning with respect to the short, until proper matching and maximum output are obtained. The crystal loop is usually made so that it holds the end of the crystal as shown in Fig. 1. Again the angle and position of this loop are adjusted to effect matching and maximum power transfer.

Another factor that should be considered in adjusting the crystal loop is the effect of loading on the *Q* of the cavity. As has been indicated in a previous article[4], the bandwidth of the resonant cavity is a function of both the loaded and unloaded *Q*, increasing as the loaded *Q* decreases. If the crystal

Performance PROTECTION



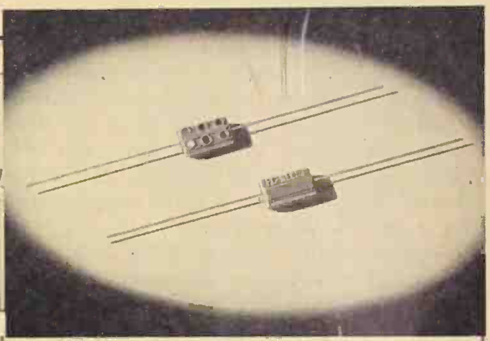


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Fig. 10. Typical crystal designed for microwave mixers for use at frequencies in the 10,000 megacycle range.

loads the cavity to the point where insufficient image rejection (most specifications require 60 db. attenuation of the image frequency) is obtained, then coupling must be decreased.

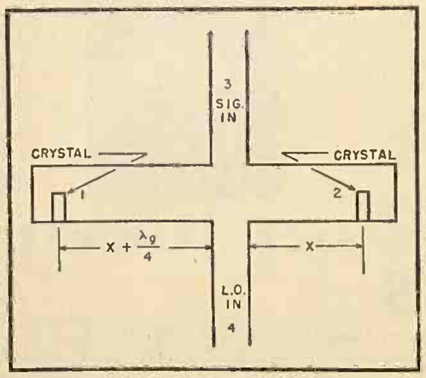
The coupling of the local oscillator is usually adjusted through experimentation. Two factors must be considered in adjusting this loop, i.e. the stability of the oscillator and the operation of the mixer. To achieve maximum local oscillator stability it is desirable to employ loose coupling between oscillator and mixer cavity. Loose coupling also prevents loss of signal down the local oscillator line.

On the other hand, sufficient local oscillator power must be available within the mixer cavity for optimum receiver noise figure. Fig. 9 shows the noise figure, F, and the conversion loss, G, of a typical crystal mixer in terms of crystal current—which is varied by adjusting local oscillator power. As indicated in this figure, considerable improvement in F and G is obtained as the crystal current increases up to about 0.4 ma., after which the noise figure increases slightly with increased current. Hence the loss in sensitivity for a crystal operating less than 0.4 ma. is considerably greater than the loss in sensitivity for a crystal operating above 0.4 ma. For this reason it is standard practice to adjust the local oscillator for a crystal current of about 0.5 ma. The local oscillator coupling is therefore made as loose as possible consistent with the crystal requirements and stable oscillator operation.

Type No.	Maximum Frequency Designed for	Conversion Loss, (Max., db.)	Noise Figure (max)	I.F. Impedance (resistive)
1N25	1000 mc.	8.5	2.5	100-400 ohms
1N21B	3000	6.5	2.0	200-800
1N23B	10,000	6.5	2.7	150-600
1N26	25,000	8.5	2.5	300-600

Table I. Important characteristics of several crystals.

Fig. 12. Balanced mixer so arranged that image frequency wave is transmitted into local oscillator arm.



There are two general types of mixer cavities used, the tuned and untuned cavity. Where image rejection is not an important parameter, the cavity can be designed to have a flat response over the entire r.f. band. This will be true when an r.f. amplifier is used, or in applications like radar where no equipment operating at the image frequency is contemplated. In applications where image frequency attenuation is required, the mixer cavity is generally tuned to the r.f. signal. This can be effected at frequencies above 3000 mc. through the use of double or triple stub tuners as shown in Fig. 11. At about 2000 mc. or below a reentrant cavity can be used with a micrometer type probe varying the capacity at the open-circuited end as shown in Fig. 1.

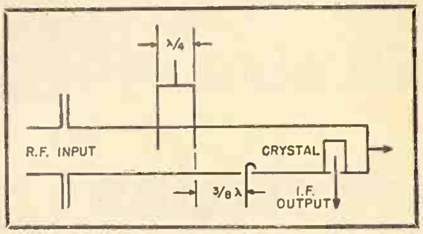


Fig. 11. Use of double stub transformer for varying resonant freq. of mixer cavity.

The output of the crystal is coupled through a coaxial line to the i.f. input. This is generally effected mechanically by making the crystal cartridge the inner conductor and the connector the outer conductor of the coaxial line. This line should be a low capacitance type to minimize loading of the i.f. input. It can be noted from the table of typical crystals listed previously, that the matching impedance of the i.f. circuit is given by the manufacturer. Provision must also be made in the output circuit for a local oscillator and r.f. bypass capacitor. This capacitance should be sufficiently large to effectively short the *LO* and r.f. energy and small enough not to absorb too much i.f. power. It can be obtained mechanically by placing a dielectric of appropriate thickness and dielectric constant around the crystal cartridge. The first i.f. amplifier is designed—using conventional techniques—to provide the best noise figure.

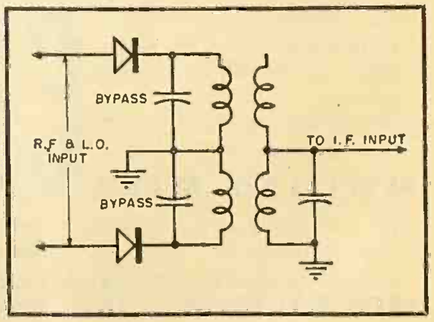
The disadvantage of the mixer described above is that some of the r.f. signal may be transmitted down the *LO* line, some of the local oscillator energy may be radiated through the r.f. input to the mixer, and noise generated in the local oscillator—klystrons are particularly noisy—is injected into the mixer. To overcome these disadvantages, a balanced mixer circuit, shown in Figs. 8 and 13, employing a "magic T" hybrid guide can be used.

The magic T is a four arm circuit consisting of a section of line, usually a wave guide, with two other lines connected perpendicularly to it. Each of the adjoining wave guides makes, with the original guide (arms 1 and 2), an ordinary T-junction. The section formed by the branch in the broad side (arm 3) of the main wave guide and the original guide is called an E-plane T-junction, and behaves essentially as a series-connected circuit. The arm in the narrow plane (arm 4), called an H-plane T-junction, acts as a parallel circuit.

There is no direct coupling between arms 3 and 4 in a magic T that is properly matched. The two crystals are placed in arms 1 and 2. The r.f. power fed into arm 3 will go into arms 1 and 2 with a 180 degree phase difference. The power fed into arm 4 will also go into arms 1 and 2, in this case with no phase difference. However, none of the power transmitted down arm 3 will go into arm 4, or vice versa. If the r.f. input is applied through arm 3 and the local oscillator power through arm 4, the signal appearing across the two crystals will consist of the *LO* power and sidebands with the same phase, and the r.f. signal 180 degrees out of phase. The i.f. signal therefore appears out of phase in the two crystals. If the two i.f. output terminals of the mixer go into a push-pull input circuit to the

(*Continued on page* 25A)

Fig. 13. Simplified representation of a balanced crystal mixer.



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In order to meet rigid government specifications on these developments—and at the same time produce them economically—HELIPOT* has perfected unique manufacturing facilities, including high speed machines capable of winding extreme lengths of resistance elements employing wire even less than .001" diameter. These winding machines are further supplemented by special testing facilities and potentiometer "know-how" unsurpassed in the industry.

So if you have a problem requiring *precision potentiometers* your best bet is to bring it to The HELIPOT Corporation. A call or letter outlining your problem will receive immediate attention!

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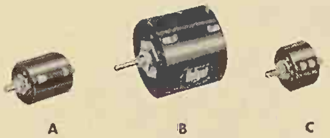
In this panel are illustrated standard models of HELIPOT multi-turn and single-turn precision potentiometers—available in a wide range of resistances and accuracies to fulfill the needs of nearly any potentiometer application. The Beckman DUODIAL is furnished in two designs and four turns-ratios, to add to the usefulness of the HELIPOT by permitting easy and rapid reading or adjustment.



MODELS A, B, & C HELIPOTS
A—10 turns, 46" coil, 1-13/16" dia., 5 watts—resistances from 10 to 300,000 ohms.
B—15 turns, 140" coil, 3-5/16" dia., 10 watts—resistances from 50 to 500,000 ohms.
C—3 turns, 13-1/2" coil, 1-13/16" dia., 3 watts—resistances from 5 to 50,000 ohms.
—Ask for Bulletin 104—

MODELS D AND E HELIPOTS
Provide extreme accuracy of control and adjustment, with 9,000 and 14,400 degrees of shaft rotation.
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Feature both continuous and limited mechanical rotation, with maximum effective electrical rotation. Versatility of designs permit a wide variety of special features.
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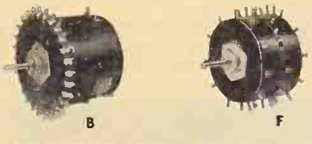
MODELS R AND W DUODIALS
Each model available in standard turns-ratios of 10, 15, 25 and 40 to 1. Inner scale indicates angular position of HELIPOT sliding contact, and outer scale the helical turn on which it is located. Can be driven from knob or shaft end.
R—2" diameter, exclusive of index.
W—4-3/4" diameter, exclusive of index. Features finger hole in knob to speed rotation.
—Ask for Bulletins 104 and 114—

The versatility of the potentiometer designs illustrated above permit a wide variety of modifications and features, including double shaft extensions, ganged assemblies, the addition of a multiplicity of taps, variation of both electrical and mechanical rotation, special shafts and mounting bushings, high and low temperature operation, and close tolerances on both resistance and linearity. Examples of potentiometers modified for unusual applications are pictured at right.



MULTITAPPED MODEL B HELIPOT AND 4-GANGED TAPPED MODEL F
This Model B HELIPOT contains 28 taps, placed as required at specified points on coil. The Four-Gang Model F Potententiometer contains 10 taps on each section. Such taps permit use of padding resistors to create desired non-linear potentiometer functions, with advantage of flexibility, in that curves can be altered as required.

3-GANGED MODEL A HELIPOT AND DOUBLE SHAFT MODEL C HELIPOT
All HELIPOTS, and the Model F Potentiometer, can be furnished with shaft extensions and mounting bushings at each end to facilitate coupling to other equipment.
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Electron-Optical Mapping in a MAGNETRON

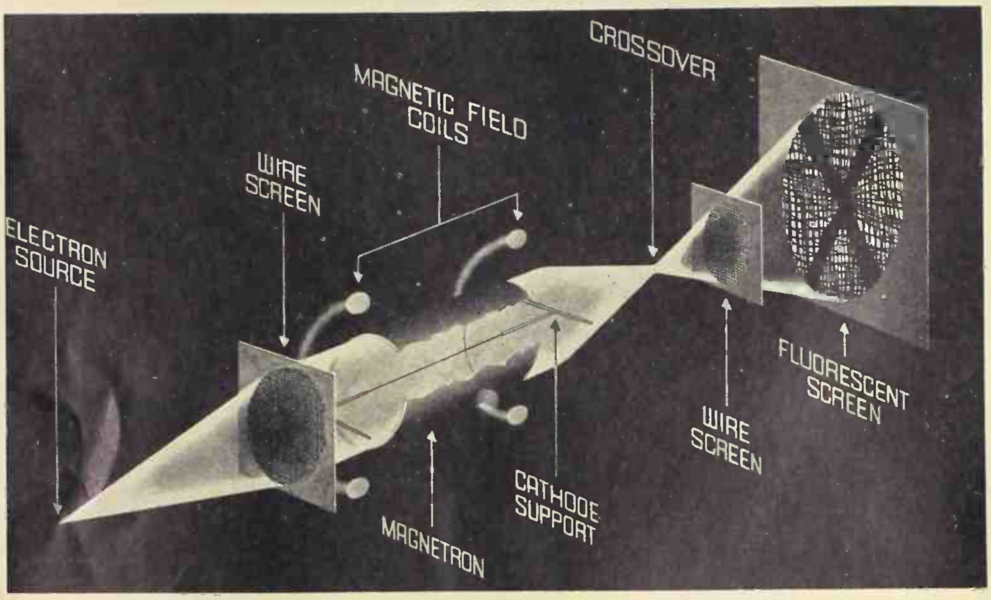


Fig. 1. Perspective drawing illustrating the electron-optical technique for experimentally mapping the electric-field distribution within a magnetron.

This new technique developed at NBS is useful for experimentally mapping electric and magnetic fields.

AN accurate, sensitive technique* for experimentally determining the electric-field distribution and space-charge density within a magnetron has been developed by D. L. Reverdin at the National Bureau of Standards. The new method, which is also well suited to investigations of electron-optical lenses, gas discharge, and other space-charge problems, is a modification of the electron optical shadow technique[1] recently developed at the Bureau for the quantitative study of minute electric and magnetic fields. A magnetic lens is used to produce shadow images of two fine wire screens placed at either end of the magnetron in the path of an electron beam. Then, from the distortion in the shadow network caused by deflection of the electron rays as they pass through the magnetron field, the radial electric field is computed and the space-charge distribution obtained.

The magnetron is a vacuum tube widely used for generating power at microwave frequencies. In its elementary form, it consists of a cylindrical cathode—ordinarily a straight wire filament—and a coaxial cylindrical anode. The tube functions under the joint action of an externally applied magnetic field and the electric field created by applying a d.c. potential between the anode and cathode.

The high space-charge density within a magnetron is known to have an important bearing on performance. However, very little is actually known concerning the electric-field distribution and space-charge configuration within the tube. Although the problem has been investigated theoretically[2] by many workers, the formidable mathematics involved have not permitted an exact solution, and the various simplifications of the theory that have been suggested have led to widely divergent results. Attempts at direct measurement have also proved unsuccessful because the very critical symmetry of the field under study was disturbed. A promising approach to the problem has now been provided by the method developed at the Bureau. This technique has been used to map the charge distribution within a cut-off or steady-state magnetron. Further application to oscillating magnetrons should lead to a much better understanding of their operation and should yield information of considerable value to the engineer who is interested in designing improved types of magnetrons or in predicting the performance of existing types.

The Bureau's method uses an electron beam as a probe but keeps the charge density of the probe beam small compared to the space charge in the magnetron. Thus, the field under study is

(*Continued on page* 29A)

** This work was carried out in connection with a doctoral dissertation submitted by D. L. Reverdin to George Washington University, Washington, D. C., in February 1950. Dr. Reverdin, formerly a guest worker at the National Bureau of Standards, has now returned to Switzerland.*

Fig. 2. Specially designed C-R tube for quantitative study of electric-field distribution and space-charge density in a magnetron.

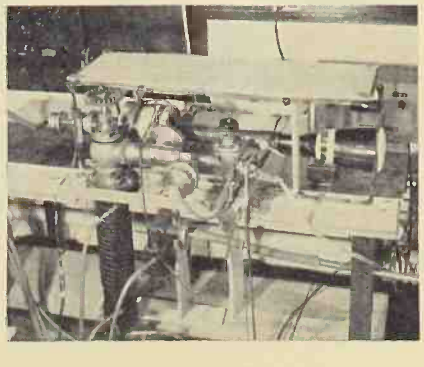
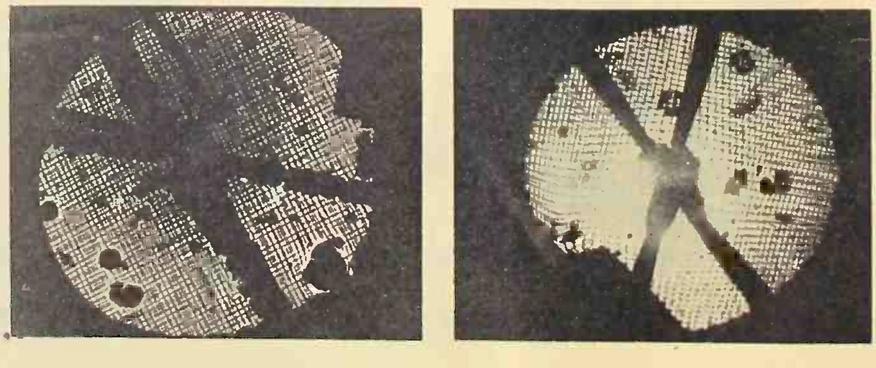
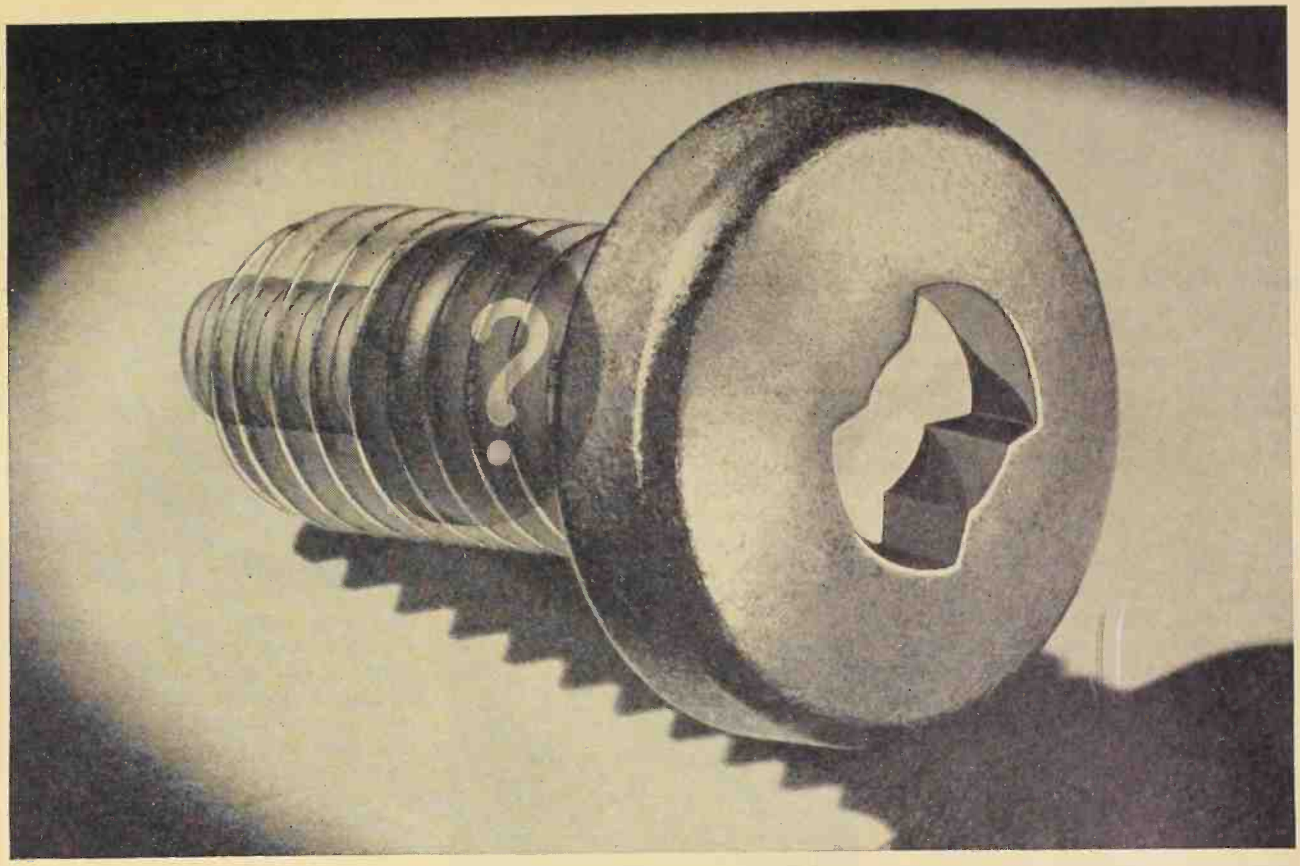


Fig. 3. Shadow patterns obtained by the new technique. With no field, the shadow pattern formed by the two screens is undisturbed (left). The presence of a field within the magnetron distorts this shadow pattern considerably (right).





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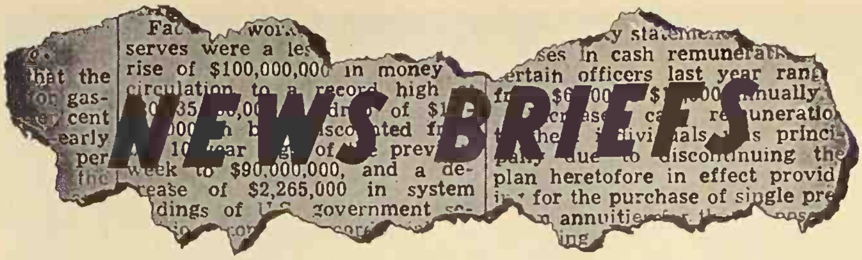


NEWS BRIEFS

RADIO PROPAGATION RESEARCH

Cheyenne Mountain in Colorado has been selected as the site for a series of ultra-high frequency radio experiments to be conducted by the Central Radio Propagation Laboratory of the National Bureau of Standards.

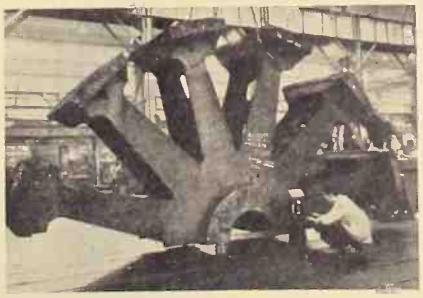
The powerful transmitters to be used by the NBS scientists will beam ultra-high frequency radio signals from this natural "skyhook" 9300 feet above sea level and out to the eastward over the flat rolling plains toward Kansas. This basic research program into the behavior of extremely high radio frequencies is designed to provide scientific information for the allocation and use of these frequencies for such systems as high frequency television and aircraft navigation and communication.

Suitable space for the operation of these transmitters for an extended period has been leased by The Broadmoor Hotel, Inc., owners of the Cheyenne Mountain Lodge.

RESISTANCE WIRE STRAIN GAUGES

The Baldwin Locomotive Works of Philadelphia solved the problem of weighing a huge 35-ton steel casting by means of resistance wire strain gauges.

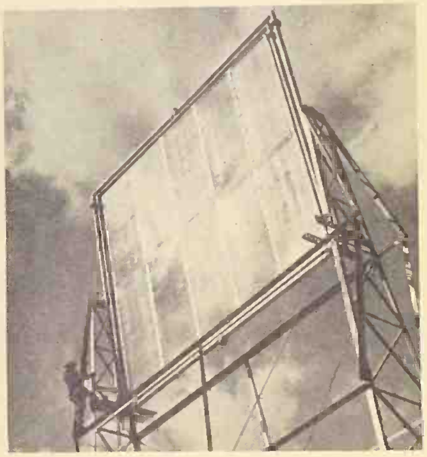
The gauges are bonded to short vertical columns incorporated in three load cells of 50,000 lb. capacity each, which were developed for accurate weighing of all kinds. The casting shown is a semi-rotor to be shipped to a hydro-electric plant in Canada.



REFLECT MICROWAVES

In the recent test on microwaves for communication between a generating plant of the *Pennsylvania Electric Company* and a substation 12 miles away, a 20-foot-square aluminum reflector like the one shown was used by *Westinghouse Electric Corporation* to provide a continuous path for the microwaves between terminal stations where line of sight was not available.

Since there was no direct line of sight between the two stations, an aluminum reflector was mounted on a mountain top to redirect the microwaves much as a mirror reflects light. Perforations in the sheet reduce its wind resistance without impairing its efficiency, and as the microwaves used have a wavelength of about one foot and the openings are only one inch wide, the waves do not pass through.



The system operates in both directions and is capable of sending seven voice conversations simultaneously.

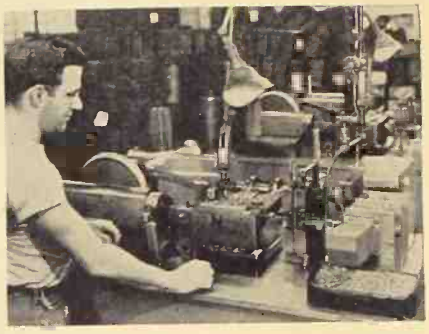
MICROWAVE REFRACTOMETER

George Birnbaum of the National Bureau of Standards has developed an instrument which measures and records small differences in frequency between two resonant cavities. In its present form, the new instrument can be adjusted over a wide band of microwave frequencies for measurement of dielectric constants of lossless gases and changes in the dielectric constant of such gases and very low-loss liquids and solids.

According to reports, the microwave refractometer should be readily adaptable to manufacture as a field model since the microwave components and electronic circuits are straightforward and compact. It has direct application in several fields of scientific research and industrial production, providing a convenient method for continuous monitoring of impurities in gases or liquids and for rapid testing of small solid samples. It could also be used as an ultramicrometer and to measure the thermal expansion of cavity materials.

METAL PIN PRODUCTS

The Parts Division, *Sylvania Electric Products Inc.*, Warren, Pa., is now producing rolled metal pin products with diameters from .040" to .070" and from ¼" to 1½" in length by specially designed high-speed pin roller machines, for electron tube leads, precise hinge pins, terminal boxes, toy train track pins, and a wide range of similar products.



Production facilities permit mass-production of small metal pins of different compositions; with square or chamfered burr-free ends; according to close specifications; with or without grooves or other shaping.

SUBMINIATURE ELECTRONIC DEVICES

A laboratory investigation in which components consisting of one to three subminiature electron tubes and associated equipments were studied was described in a paper delivered to a Symposium on Improved Quality Electronic Components recently sponsored jointly by RMA, IRE, and AIEE in cooperation with NBS and other government agencies.

W. Wheeler, W. H. Hannahs, and J. A. Caffiaux of *Sylvania Electric Products Inc.*, prepared the paper and referred to these specialized components for miniaturization as "unit sub-assemblies." The basic design was developed to facilitate quick repair and maintenance of electronic circuits by means of plug-in units. It also permits hermetic sealing or the embedding of complete sub-assemblies in solid blocks of plastic.

NEW SYLVANIA PLANT

The first step in *Sylvania Electric Products Incorporated's* program to expand its radio tube manufacturing facilities will be the immediate construction of a new plant in Shawnee, Oklahoma. When completed around the first of next year, the new plant will have a capacity of more than a million radio tubes per month.

C. W. Hosterman, formerly manufacturing superintendent of the *Sylvania* plant at Huntington, West Virginia, will be manager of the new plant which will cover 50,000 square feet and will employ approximately 500 people. This will be the first *Sylvania* plant in the Southwest; other plants are located in Massachusetts, New York, Pennsylvania, Ohio, and West Virginia.

Microwave Receivers

(*Continued from page* 20A)

i.f. amplifier, the noise originating in the local oscillator will be cancelled out.

The role of the signal-input and local oscillator (*LO*) input arms can be interchanged. Since there is no direct coupling between the signal-input circuit and *LO* input circuit, no decoupling of the local oscillator is required. Only enough local oscillator power to drive the two crystals is required. If more local oscillator power is available a matched dissipative attenuator can be used in the *LO* arm. The load admittance presented to the local oscillator is relatively constant because the crystals are positioned in the guide arms to provide a good match to the wave guide. The fact that the mixer can be operated with relatively low *LO* power, without danger of loss of signal through interaction between r.f. input and *LO* line, is one of the many advantages of the magic T mixer.

It is possible to feed the image frequency wave into the local oscillator dissipative attenuator by making one of the crystal arms a quarter of a wavelength longer than the other as shown in Fig. 12. The result of adding a quarter of a wavelength to one side of the magic T is that, if image frequency waves of equal amplitude are developed by the crystals, their phase relationship as they converge on the junction is such that they are transmitted entirely down the local oscillator arm of the magic T.

The following experimental procedure can be used to design a magic T balanced mixer. First the size of the wave guide to be used is determined based on the type of wave guide used for the r.f. input. Two sections of slotted wave guide, each shorted at one end, (one quarter of a wave longer than the other when image rejection is desired) are selected—the exact length of the line is primarily a function of mechanical convenience. A crystal is placed a short distance from the shorted end of the line and an r.f. signal is fed across the open end of the line. The crystal position is then varied, by sliding it down the slot, until standing waves, as measured by a slotted line standing wave meter, are minimized. This is done for both sections.

The two sections are then joined and one input line inserted at the center. This line must be properly terminated (as shown in Fig. 8) by a post or a window (also called iris). These elements have previously been described.[4] A number of trial and error tests must be made to determine the best matching network over the entire band. It is relatively simple to match for one frequency but the problem is to obtain good matching over the complete r.f. band—otherwise some method of variable tuning must be provided which would be very complex. After proper termination for this arm is determined, it is removed and the other input arm is properly terminated by a similar procedure. Then both arms are placed in the circuit and standing waves minimized in all arms over the band. It should be noted that a capacitance should be provided at both crystals for r.f. and *LO* return. The output of the crystals is fed into a balanced circuit as shown in Fig. 11.

Local Oscillator

The type of local oscillator used will depend upon the power requirements of the mixer, the stability desired, and frequency of operation. Klystron and lighthouse oscillators are used for most receivers, with the lighthouse type preferred—when they can be used—because they are less noisy. The design principles of these oscillators are essentially the same as those employed for the oscillators described in the articles covering transmitter design [2,3]. Automatic frequency control circuits previously described are used to maintain the necessary oscillator stability. When a balanced mixer is used (low *LO* power required) it is sometimes advisable to employ a crystal oscillator and operate at a fifth or sixth harmonic.

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* * *

NEW PRODUCTS

MEASURING INSTRUMENT

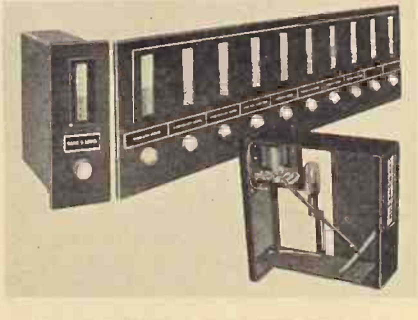
The Analascope, announced by *Analytical Measurements Inc.*, 585 Main St., Chatham, N. J., is a new instrument providing a convenient dependable means for measuring and continuously showing any phenomena that can be translated into electrical impulses.

Results are displayed on a long persistence screen of a 5″ cathode-ray tube and allows continuous observation of non-recurrent phenomena at sweep speeds of from .01 to 5 seconds. A two decade precision potentiometer calibrated with a self-contained standard cell provides direct measurements from .001 pH to 15 pH and 0.1 millivolts to 1.5 volts.

An input attenuator of 1000 megohms provides overlapping ranges of 1 micromicroampere to 10 milliamperes and 1 millivolt to 100 volts per centimeter deflection.

INDICATING RECEIVER

Panalarm Products, Inc., 7218 North Clark St., Chicago 26, Illinois, has developed a pneumatic indicating receiver with illuminated direct reading dial.

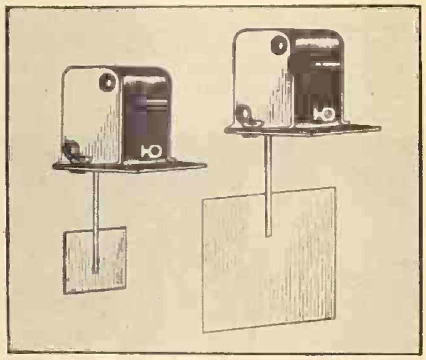


An exceptionally sensitive response to changes in levels in tanks, towers or vessels, or changes in valve position corresponding to variations in impulse pressures from any 3-15 lb. transmitter, from any remote point, is visually simulated on the graduated dial. A solid red band on this translucent scale dynamically depicts every dip and rise in levels.

Receivers can be equipped with positive "high" and "low" alarm mercury switches, adjustable over the entire range. These may be supplied to operate a standard *Panalarm* signalling unit, mounted in the gauge.

AIR FLOW SWITCH

Coral Designs, division of the *Henry G. Dietz Co.*, P. O. Box 248, Forest Hills, New York, N. Y., has an improved vane type air flow switch for use in forced air cooling of electronic equipment. The quantity of air required for forced-air-cooled tubes is specified for various types of service and often for various power levels.



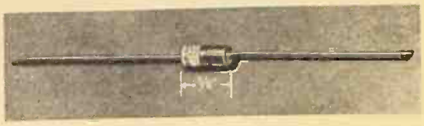
The Cat. 103-A Vane Type Air Flow Switch is designed to operate a control relay to guard against tube failure in the event of blower failure or air-passage obstruction and guards against air flow failure. It will operate on a minimum velocity of 500 feet per minute. Electrical ratings of 5 amperes at 250 volts a.c. are Underwriters' Laboratories approved.

A data sheet describing this switch will be sent upon request.

OXIDE RECTIFIER

A copper oxide rectifier has been designed by *Bradley Laboratories, Inc.*, 82 Meadow St., New Haven, Conn., to obtain an extremely high reverse resistance of over one megohm per plate.

Featuring vacuum processed gold contacts, CX18 has 2½″ leads. The rectifier barrel is only ⅝″ long and ¼″ in diameter, and is rated up to 5 milliamperes d.c. The unit is intended for circuits in which very low leakage and maximum stability are essential.



INFRARED SPECTROMETER

The Perkin-Elmer Corporation, Glenbrook, Conn., has announced its Model 12-C Infrared Spectrometer with many improvements, including mobility, all-metal cabinet, faster recorder, and new wavelength drive.

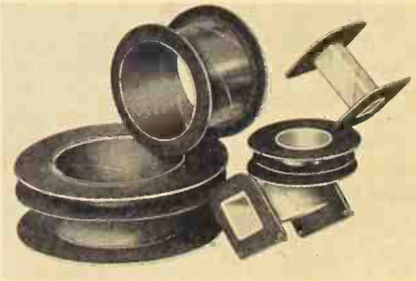
Controls and recorder of the new cabinet are readily accessible to the operator in either a sitting or standing position, and rugged construction of spectrometer, recorder, and cabinet permit factory assembling and shipment as a unit. Over-all dimensions of the caster-mounted cabinet are 54″ long x 46″ high x 30″ wide. The improved wavelength drive has simplified, completely reversible speed-change controls.

All improvements will be available for delivery early in September; most of them will also be obtainable as accessories for existing *Perkin-Elmer* infrared spectrometers.

COIL BOBBINS

Precision Paper Tube Co., 2045 W. Charleston St., Chicago 47, Ill., is now manufacturing locked-flanged precision coil bobbins made with a new plastic-coated core on which the flanges are securely locked. Coils can be wound on these vulcanized fibre bobbins to closer tolerances, and fewer rejects are assured.

According to the manufacturer, the new plastic-coated core makes *Precision* coil bobbins 15 to 20% stronger, yet light in weight. It also increases the bobbins' insulating qualities and moisture resistance, and has excellent heat-dissipating qualities. Spiral-wound, this core can be made any size, any shape, and to any ID or OD with very close tolerances.

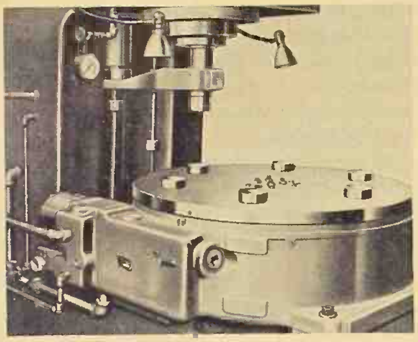


Flanges can be supplied in any desired shape, and with any combination of slots or holes for terminal wiring, and can be furnished flat, recessed, or embossed. Bobbins can be made in any shape or size to buyers' specifications.

Full information and free samples may be secured by writing the manufacturer.

AUTOMATIC INDEX TABLE

A 33" automatic Index Table with 24" work-circle is now being manufactured by *The Denison Engineering Co.,* 1160 Dublin Road, Columbus 16, Ohio. This index table can be used on either *Denison's* 35-ton Multipress or other makes of equipment.



Hydraulically powered, this table provides variable speeds for any preselected indexing rate from 10 to 70 indexes per minute, and positions the dial with an accuracy of plus or minus .002". When installed on the Multipress, it is powered by the pumping unit of the press through the control system of the press. It can also be operated by a small auxiliary pumping unit when used with other than hydraulic equipment.

An important feature of this table is said to be a positive locking device that holds the dial firmly in place at each station. The connecting linkage between driver and locking mechanism releases the locking pin as the table prepares to rotate through each cycle.

RECORDING TAPE

The *Minnesota Mining & Manufacturing Co.*, 900 Fauquier Ave., St. Paul 6, Minn., has announced that its splice-free "Scotch" Brand Sound Recording Tape, ¼" x 2400 ft., is now available on aluminum N.A.B. approved hubs or reels.

This tape, No. 111, is individually boxed in a sturdy, cloth-hinged, telescope-type box with four solid corners to provide maximum protection at all times. Current production of No. 111 has very high temperature resistance, permitting operation at temperatures as high as 200°F. and shows no oxide rub-off on the heads.

Output uniformity is considerably better than ¼ decibel within the reel. Uniformity from reel to reel is near ¼ db.

SILICON LAMINATION

Now available from the *Thomas and Skinner Steel Products Company,* Indianapolis 7, Indiana, is a new oriented silicon lamination which permits reduced transformer stacks with resultant savings in materials and weight. According to the manufacturer, transformer designers will obtain greater efficiency with the normal number of E's and I's, or equal efficiency with fewer laminations.

These laminations are available in standard E and I designs and sizes as well as for special applications. A bulletin giving full data and curves is available upon request.

POWER SUPPLY

Model 315 voltage regulated power supply announced by *Kepco Laboratories, Inc.*, 149-14 41st Ave., Flushing, New York, features one regulated B supply, one regulated C supply, and one unregulated filament supply.

The B supply is continuously variable from 0 to 300 volts and delivers from 0 to 150 milliamperes. The C supply is continuously variable from 0 to 150 volts and delivers 5 milliamperes. A.c. output is 6.3 volts, 5 amperes, center-tapped, unregulated.

This unit is mounted in a sturdy cabinet, and is also available for standard relay rack mounting on special order.

SCREEN PROCESS PRINTER

Mech-Tronics Equipment Co., Box 510, Silver Spring, Md., has announced the availability of its Model 20 Screen Process Printer which automatically applies wiring and capacitors to cylinders up to 1¼ inches in diameter.

Screen sizes can be accommodated up to the dimensions of the 5" x 9" frame. A flat plate printer, Model 30, is available on a 60 day delivery basis.

Additional information can be had by writing to *Mech-Tronics* for Information Bulletin No. 21.

Television

41 MC. TV I.F.

Mr. D. W. Pugsley of the *General Electric Co.* presented a paper at the 1950 Cincinnati Television Conference devoted largely to the practical aspects of a 41 mc. i.f. for TV receivers. This frequency is in accordance with the new R.M.A. Standard of 41.25 mc. for the sound and 45.75 mc. for the picture carrier.

The 41 mc. i.f. utilizes three 6BC5's in a stagger tuned circuit to give a gain of 69.4 db., as compared with 66.4 db. for the old 20 mc. i.f. using 6AU6's. Also the bandwidth is increased from 3.0 to 3.25 mc. Other advantages include greatly reduced image interference, reduction of interference from diathermy and radio amateur equipment, and the elimination of interference between receivers due to local oscillator radiation. The 41 mc. i.f. costs slightly more to build, but it was felt that the increased performance justified the increased cost.

Mr. Pugsley concluded his paper with a recommendation that all TV receiver manufacturers convert to the 41 mc. i.f. on the basis of *General Electric's* experience.

TV WAVEFORM MONITOR

Polarad Electronics Corporation, 100 Metropolitan Avenue, Brooklyn 11, N. Y., now has available a portable television waveform monitor designed for waveform analysis and amplitude measurement of video signals in television circuits.

Model TO-1 is a companion instrument to the Portable Picture Monitor, Model 102-MPS, and may be used as a general purpose instrument in many applications. According to the manufacturer, it has a wide frequency response, high sensitivity, excellent synchronizing capability, precision calibrating circuits, and unusually large symmetrical horizontal expansion. Visual presentation is on a 5-inch cathode-ray tube.

REMOTE VIEWING BY TV

The Remote Control Engineering Division of the Argonne National Laboratory has developed a method for working with radioactive materials at a distance by remote control and stereo-television. Directed by Dr. H. L. Hull, the study, design, and development of this method is an important step in the development of equipment and techniques needed by scientists in order to protect themselves from radiations emitted by many of the materials used in the atomic energy program.

The method involves manipulation by the use of various types of remote control devices while the operations or manipulations are viewed by use of stereo, or three-dimensional, television. In certain operations, where the intensity of radiation is quite high, a system of manipulation is required in which the experimenter is at a considerable distance from the material being handled. In such a system, the conventional methods of viewing are not practical, and a viewing system involving the use of three-dimensional television has been employed.

The investigation of a three-dimensional television system was carried out by a group of Remote Control Engineers under the leadership of Prof. H. R. Johnston, who is on loan to the Argonne National Laboratory from Northwestern University's Department of Electrical Engineering.

MULTIPLE TV ANTENNA

Plans for a super-antenna project to provide facilities for television stations in the New York area were furthered recently with the signing of a contract between *RCA Victor* Division, the *National Broadcasting Co., Inc.*, and representatives of *Empire State, Inc.* The contract calls for development of a multiple television and broadcasting system atop the Empire State Building.

A specially engineered supporting structure for five TV antennas and three FM antenna systems is contemplated, as well as the installation of two emergency TV antennas on the mooring mast of the building. Work under the contract will be started in the television laboratories of the *RCA* Engineering Products Department in Camden, N. J., as soon as participating stations and networks provide antennas for the project.

It is expected that this proposed focal point of telecasting service will result in better reception for millions of set owners in the Metropolitan New York area.

ANNUAL TECHNICAL CONFERENCE

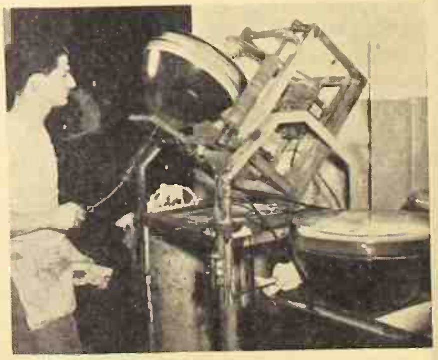
The Fourth Annual Spring Technical Conference of the Cincinnati Section IRE, held recently at the Engineering Society Headquarters in Cincinnati, featured "Television" as its theme. Eight papers by prominent engineers were given, climaxed by the banquet address of *Eastman Kodak Company's* Ralph M. Evans on "Seeing Light and Color."

Also featured were a number of interesting exhibits of new apparatus by manufacturers in the television field.

COATING FOR TV TUBES

At the *Tel-O-Tube Corporation,* East Paterson, New Jersey, 1000 to 1300 television tubes a day are brush-coated with a "dag" colloidal graphite dispersion, as shown in the photograph. The opaque electrically conducting coating improves picture performance by absorbing reflected light, and also serves as a final high voltage anode in the electronic system.

Manufactured by the *Acheson Colloids Corp.*, Port Huron, Michigan, "dag" colloidal graphite is chemically inactive, highly resistant to electron bombardment, and low in photoelectric sensitivity.



TECHNICAL BOOKS

"AN INDEX OF NOMOGRAMS," compiled and edited by Douglas Payne Adams. Published jointly by The Technology Press of Massachusetts Institute of Technology and *John Wiley & Sons, Inc.*, 440 Fourth Ave., New York 16, N. Y. 174 pages. $4.00.

This unusual index serves as an invaluable time-saver in the repeated solution of mathematical formulas as it lists over 1700 nomograms published in well-known periodicals in twenty-one fields of science and engineering.

The book is divided into two main parts: Index A—Key Words, and Index B—Master Index. Index A contains an alphabetical list of key words which are associated with each of the diagrams. Following each entry is the key number permitting reference to Index B where the periodical, date of issue, volume number and page number of nomogram are listed.

In addition, the variables employed in each diagram are shown in a bracket following its title so that the reader will know its content at a glance. This index will be invaluable in directing the reader to the nomogram he will need for the quick and accurate solution of his problem.

"TRANSIENT PERFORMANCE OF ELECTRIC POWER SYSTEMS" by Reinhold Rudenberg. Published by *McGraw-Hill Book Company*, 330 West 42nd St., New York 18, N. Y. 832 pages. $12.00.

This is an enlarged edition in English of a German text on the same general subject prepared by the author. It deals with the transient performance of power circuits with lumped properties and the topics correspond to parts of courses the author has given on electric machinery, apparatus and instruments, power transmission and distribution, and transients proper during the last twelve years at Harvard University.

Mathematical analysis is considered as an auxiliary tool, and emphasis has been placed on the physical understanding of the analysis of an example and the ease of discussion of the result. In order to serve as a reference, the various chapters considering different phenomena are built up as independently as possible.

Many numerical examples, taken from actual practice, are worked into the text and numerous oscillograms and measurements throughout may further bridge the gap between theory and the practice of the transients. They also enable the reader to place immediately on the correct numerical level the results derived from the analysis. The numerous oscillograms and measurements distributed throughout the text, which for the most part stem from the author's industrial activities, may further interlink the practice and theory of transients. A bibliography at the end of the book refers to a number of significant publications for every chapter.

* * *

Electron-Optical

(*Continued from page* 22A)

undisturbed. An electron gun sends the beam axially through the tube. Coaxial coils surrounding the magnetron provide a homogeneous magnetic field for the operation of the magnetron and at the same time act upon the beam as a convergent magnetic lens, bringing it to a focus beyond the tube. Two fine wire screens are placed in the path of the electrons, one just in front of the magnetron, the other just beyond the back focus of the beam. A complex shadow pattern due to the two wire screens is then formed on a fluorescent screen. When the d.c. potential across the magnetron is zero, the pattern is undistorted. However, when an electric field is applied to the magnetron, the shadow network on the fluorescent screen becomes quite distorted; and theoretical analysis of this effect has related the distortion of a given part of the pattern to the intensity of the electric and space-charge fields in the corresponding region of the magnetron.

In practice, photographs are taken of the shadow network, both in the undistorted and distorted form. The changes in the paths of the electron rays as they pass through the magnetron are then determined from measurements of the shadow patterns and the geometrical constants of the system, such as the positions of both wire screens, the magnetron, and the electron source, and the number of meshes per unit length of the wire screens used. From the deflection of an electron ray entering the magnetron at a given radial distance from the center, the strength of the electric field in the corresponding region of the magnetron is computed.

In comparison with previous methods using a pencil beam of electrons but no optical system, this method is much more sensitive and accurate. It also has the advantage of giving a complete field map in a very short time. The principal source of error lies in the uncertainty regarding the configuration of the electric fringe field at either end of the magnetron under space-charge conditions.

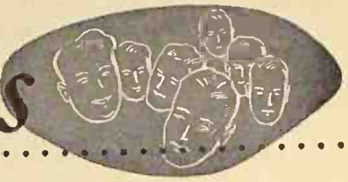
The Bureau's study of the field within a steady-state magnetron indicates that the actual space-charge distribution differs considerably from that predicted by the theorists. A number of different shapes of space-charge configuration were observed which are closely related to the symmetry of the magnetron. A certain lack of sharpness noted in the patterns gave a visual indication of the noise in the tube. This suggests further extension of the method to learn more about the problem of noise in an oscillating magnetron.

REFERENCES

1. *Electron-Optical Shadow Method, NBS Technical News Bulletin, 33,* 106 (1949).
2. *Hull, A. W., Phys. Rev. 18,* 31 (1921); *L. Brillouin, Phys. Rev. 60,* 385 (1941).

* * *

Personals



V. A. CARPENTER has joined *National Electronics, Inc.*, Geneva, Illinois, as Chief Engineer according to an announcement by Mr. John H. Hutchings, vice president of *National Electronics*. Mr. Carpenter has twenty-three years of experience in the electronics field and has been associated with *Western Electric Company, Trimm, Inc.*, and others. Most recently he was vice president of *Continental Electric Company*, Geneva, Illinois.

C. B. DALE, director of research for *Webster-Chicago Corporation*, has been elected vice president in charge of research. Mr. Dale, who joined *Webster-Chicago* ten years ago, has been in charge of the development of all current models of record-changers and magnetic recorders, and has a number of major patents in his name. The election of W. S. Hartford, general sales manager, as vice president in charge of research was also announced by the company.

SAMUEL FREEDMAN, formerly New Developments Engineer for *DeMornay Budd Inc.*, of New York and California, is now Owner and Manager of the recently organized firm of *Technical Products and Services Company*, Santee, California. Author of many books and articles on two-way radio, aviation radio, microwaves, etc., Mr. Freedman will compile informative literature and catalogues on new radio-electronic developments for schools and laboratories.

GILBERT E. GUSTAFSON, vice president in charge of engineering at *Zenith Radio Corporation*, Chicago, Ill., received an honorary degree in electrical engineering from Stevens Institute of Technology at the school's commencement exercises in Hoboken, N. J. Mr. Gustafson is a Fellow of IRE, a member of the ARRL, the Radio Engineers Club, and has served as chairman of many committees in engineering and trade organizations.

BRIGADIER GENERAL TOM C. RIVES, USAF (Retired), has been appointed to the staff of the Commercial Equipment Division of the *GE* Electronics Department at Electronics Park where he will handle special assignments. General Rives, who was awarded the Distinguished Service Medal for his radar work during the war, received his B.S. degree in electrical engineering from Alabama Polytechnic Institute and his M.S. degree from Yale University.

RALPH B. TOMER has been appointed Chief Commercial Engineer of *Hytron Radio & Electronics Corporation* and will supervise the relations between the engineering divisions of *Hytron* and its customers. Mr. Tomer has been associated with the industry for the past 15 years in the capacity of commercial engineer, design engineer, production engineer and industrial engineer. He was also associated with the development of the proximity fuse and the Land Camera.

Strain Gauge

(*Continued from page* 11A)

is one-half of one per-cent and with its critical damping and mirror scale is excellent for the job. The price is also nominal, listing around $60.00.

Amplification systems are beyond the scope of this paper, varying to quite some extent depending upon the transmitted frequency range and associated equipment. Of course it is entirely practical to use a.c. on the bridge and a carrier system of amplification and detection. This is used commonly where numerous channels of different output levels are to be recorded on an oscillograph. The formula for determining dE_o is also quite correct, r.m.s. values being indicated for bridge supply volts and dE_o.

Load rings, like load links, will calibrate out to straight line conversion curves. Load rings are linear in their characteristics, but where over-all high accuracy is desired it is necessary to calibrate the ring with appropriate equipment.

400 Mc. Performance

(*Continued from page* 8A)

gle-stage test amplifier is shown in Fig. 5. The crystal type voltmeter probes were built into the circuit. A vacuum tube millivoltmeter was used in conjunction with the crystal probes.

In tuning the amplifier there are six variables. These are the four tuners and the two spacings between tuner end plates in the input and output coupling networks. Due to interaction of these coupling elements there can be several combinations of tuning which give both resonance at the desired frequency and the desired impedance match; however, the insertion losses of the network are greater at some combinations than at others. In all cases tuning was accomplished for maximum output with a 400-megacycle input, indicating matched impedance at resonance and minimum insertion loss.

Gain measurements were made by a measurement of input and output voltage, using the built-in crystal probes and a vacuum tube millivoltmeter. No claim is made for the absolute accuracy of the voltage measurements, but it was found that at low voltages both crystal probes gave the same indication when connected across the same source. At higher voltages (0.2 volts) the maximum difference between indications on the same source was under 10 per-cent.

An investigation of input matching and standing waves shows that false gains may be measured due to standing waves created by a mismatch in the input circuit. Inasmuch as the input voltmeter probe is physically separated from the input tuner it can read the correct voltage across the tuner only for unity standing wave. When an input mismatch occurs the input power decreases, causing a reduction of output power and voltage, but a condition of input tuning can be found which places the input probe at the trough of the standing wave, giving an indication of low input voltage and a consequent false indication of increased gain. The physical separation of the probe and the point of desired measurement is made necessary by the sizes of the components involved and cannot be avoided and, therefore, accurate gain measurements are possible only for conditions of matched impedances. As this is also the desired condition, the crystal probe voltmeters may be used if means are available to insure the all-important impedance match.

For this reason the input conditions of the amplifier were investigated by the use of the *Kay Electric* Mega-Match equipment. Through the use of this apparatus the conditions of input match are easily observable. A block diagram of the equipment is shown in Fig. 8.

The sweeping oscillator sends a signal down the 100-ft. delay cable and a portion of this signal is reflected by a mismatch at the far end. At any instant the reflected signal at the crystal detector will differ in frequency from the incident signal by the rate of change of the FM signal times the cable delay. The crystal will detect the beat between these two signals to produce a low frequency output. The low frequency is then amplified and applied to the scope. Synchronization is insured by using the sweep voltage of the sweeping oscillator on the horizontal scope plates. The low frequency signal to the scope is proportional in amplitude to the smaller of the two signals into the detector. This is always the reflected signal. For a shorted cable the reflection coefficient is unity and the scope signal is a maximum. When the cable is terminated in its characteristic impedance the reflection is zero and the scope pattern becomes a horizontal line.

Editor's Note: In this and future issues, page numbers for the Radio-Electronic Engineering section will be followed by the letter A to distinguish them from the Radio and Television News section.

The name of Glen Walters was inadvertently omitted as co-author of the articles entitled "Helical Coils as Transmission Lines and Radiators," July, 1950 and "Automatic Antenna Pattern Recorder," June, 1950 Radio-Electronic Engineering section.

PHOTO CREDITS

Pages	
4A	Westinghouse Electric Co.
5A (bottom)	General Electric Co.
6A	Sylvania Elec. Prod. Inc.
15A, 16A, 18A	Federal Telecommunication Labs.
22A	National Bureau of Stds.

Inasmuch as the 400 megacycle amplifier was designed for a 50-ohm input and the cable used was 50-ohm RG—8/U, the reduction of the scope pattern to a straight line with the amplifier on the cable end indicates an impedance match. The amplifier input consists of reactive elements such that a match is possible over comparatively narrow bandwidth. Fig. 6 indicates scope patterns for (A) an unterminated delay line, (B) delay line terminated by an amplifier with mismatched input, and (C) delay line terminated in a properly tuned amplifier for matched input.

The small marker pip shown on the scope for frequency locating purposes is generated by an oscillator set at 400 megacycles and injecting a signal into the attenuator preceding the crystal detector. This equipment thus will give an indication of correct input matching. The sweep amplitude can then be reduced to zero, causing a fixed frequency output from the oscillator at the center frequency (400 mc.) and the gain measured by the v.t.v.m.

The basic setup was modified to make possible the observance of input and output bandpass characteristics on the oscilloscope. This circuit is shown by the alternate switch positions indicated in Fig. 8 and the resulting pass band characteristics are shown in Fig. 7.

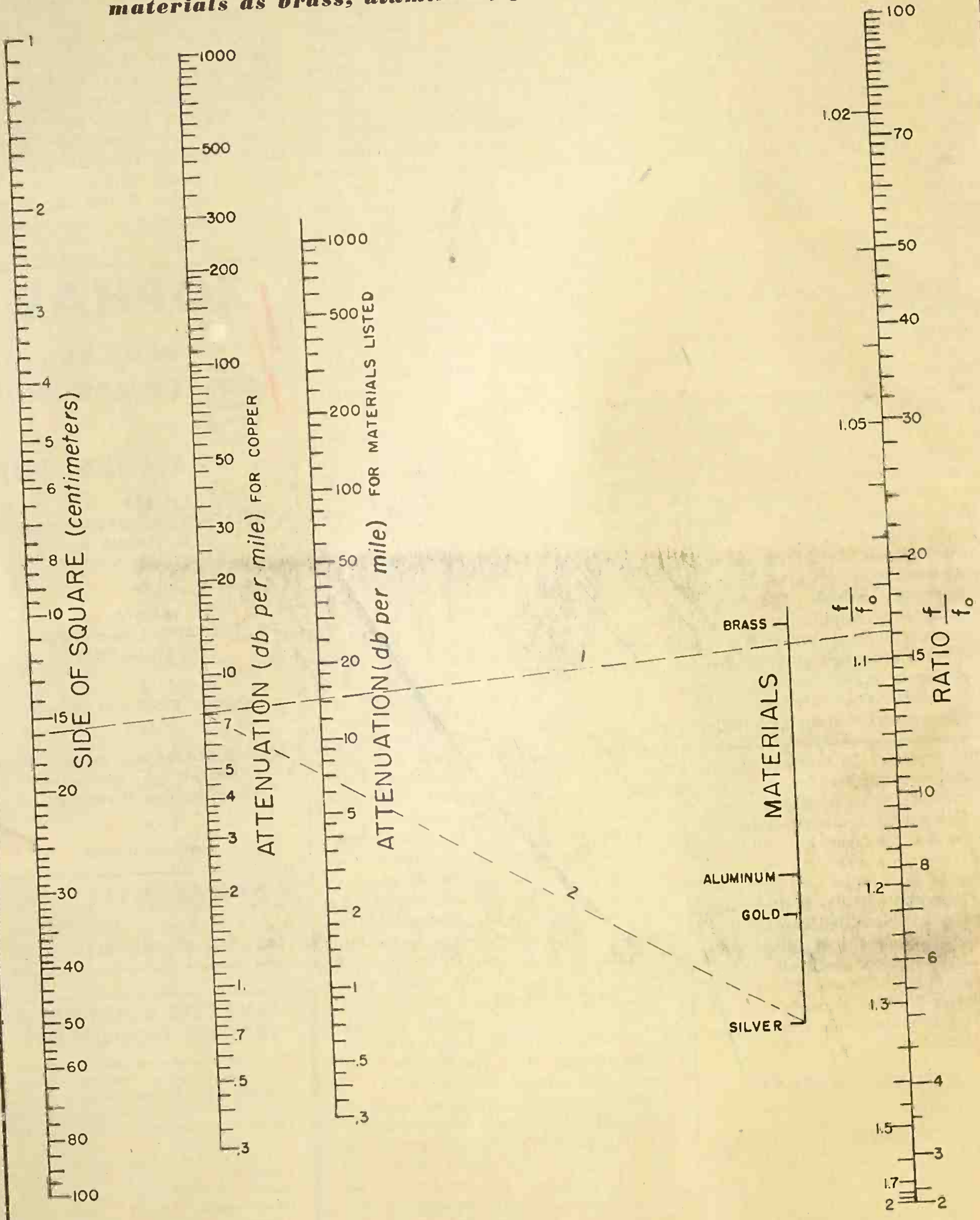
Tubes for receiver amplifier use at 400 mc. must necessarily be considered as power amplifiers because the low interelectrode impedances have comparatively low resistive components. This leads to the need for impedance matching both into and out of the tube.

In a properly designed circuit tuned for proper matching conditions, the type 5840 pentode is capable of considerable gain at 400 megacycles. A load of 437 ohms is used to simulate the loading of a following stage containing an equivalent tube operated in the same manner. The actual tube voltage gain from a 437-ohm level to a 437-ohm level is measured as approximately 4.5 with an over-all bandwidth of approximately 9 megacycles. This corresponds to a power gain through the tube of 20.

These figures on performance are a little better than would be indicated by calculations based on our input resistance measurements. This is not unexpected as input resistance can be shown to be a function of circuitry as well as tube construction. The primary circuit element of interest in increasing input resistance is the unbypassed length of screen lead. The effect of an unbypassed screen lead is essentially regenerative, as shown by Strutt and Vander Ziel. If carried too far, the feedback from screen lead inductance can cause oscillation. This screen lead regeneration can be neutralized at one frequency by resonating out the lead inductance with a bypass condenser selected to cause series resonance. When this is done the measured gains at exact center frequency of 400 mc. are in close agreement with the calculated value, but oscillation and instability results as the unit is tuned to other frequencies only moderately removed from 400 mc. Heavy screen bypassing (175 $\mu\mu$fd.) was used in the amplifier of this test to avoid this critical region.

SQUARE WAVE GUIDE ATTENUATION

Nomograph for determining square wave guide attenuation for the $TE_{0,1}$ mode for wave guides constructed of such materials as brass, aluminum, gold, silver, and copper.



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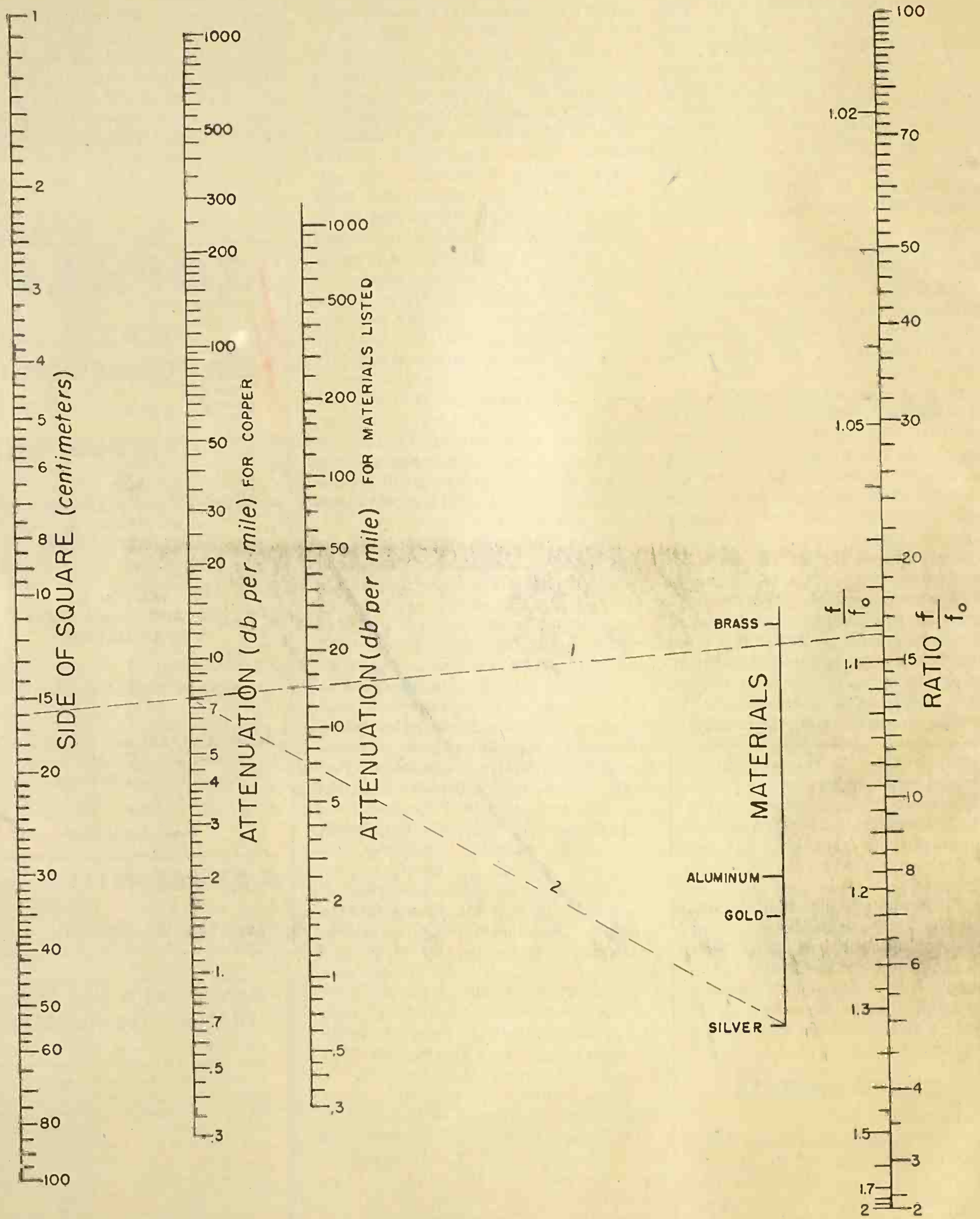
SQUARE WAVE GUIDE ATTENUATION

Nomograph for determining square wave guide attenuation for the $TE_{0,1}$ mode for wave guides constructed of such materials as brass, aluminum, gold, silver, and copper.



Courtesy of Federal Telephone and Radio Corp.

Train at Home in Spare Time for RADIO and TELEVISION

I Send You 18 BIG KITS OF RADIO-TELEVISION EQUIPMENT

My Famous Training System Prepares You in Double-Quick Time for a Good Job or Your Own Profitable Radio-Television Business

Radio-Television is now America's greatest opportunity field! Trained men are needed to fill good jobs and handle profitable Radio-Television Service work. I have trained hundreds of men for success in Radio-Television—and I stand ready to Train you too, **even if you have no previous experience.** My training is 100% practical—designed to give you the knowledge and experience you need to make money in Radio-Television in the shortest possible time. I Train you with up-to-the-second revised lessons—PLUS many big kits of Radio-Television equipment. You actually do over 300 demonstrations, experiments and construction projects. In addition, you build a powerful 6-tube 2-band radio, a multi-range test meter and a complete Television receiver! All equipment is YOURS TO KEEP.

EASY TO MAKE EXTRA MONEY WHILE YOU LEARN

You do all your training with me AT HOME in spare hours. Keep right on with your present job and income while learning—and earn extra cash besides! The day you enroll I begin sending you plans and ideas for doing profitable spare-time Radio-TV work. Many of my Sprayberry students pay for their entire training this way! You get priceless experience and many plans for making extra money. You build all your own Radio-TV Test Equipment from parts I send you—nothing else to buy. Just one more reason why I believe I offer the ambitious man the biggest value in top notch Radio-TV Training available anywhere in America today.

Be Ready for Top-Paying Radio-Television Jobs

Radio-Television is growing with amazing speed. More than 2000 Radio broadcasting stations PLUS an additional 102 Television stations are now on the air. Radio sets and TV receivers are being made and sold in record-breaking numbers. If you enjoy working with your hands . . . if you like to do interesting and varied work . . . if you really want to make good money and work in an industry that has a future . . . YOU BELONG IN RADIO-TELEVISION. But you MUST have good Training to "cash in" . . . the kind of Training that starts you out with basic fundamentals and carries you right through every circuit and problem of Radio-Television Servicing and Repair. In a word . . . that's Sprayberry Training . . . the course backed by more than 20 years of association with the Radio-Television industry!

FREE 3 BIG RADIO-TELEVISION BOOKS

I want you to have ALL the facts about my complete system of Radio-Television Training! Act now! Rush the coupon for my three big Radio-Television books: "How to Make Money in Radio-Television," PLUS my new illustrated Television Bulletin PLUS an actual sample Sprayberry Lesson—all FREE with my compliments. No obligation and no salesman will call on you. Send the coupon in an envelope or paste on back of post card. I will rush all three books at once!

Sprayberry Academy of Radio, Dept. 25-M
111 North Canal St. Chicago 6, Ill.

Mail Coupon Today!

NO OBLIGATION
No Salesman Will Call

SPRAYBERRY ACADEMY OF RADIO, Dept. 25-M
111 North Canal St., Chicago 6, Ill.

Please rush to me all information on your Radio-Television Training Plan. I understand this does not obligate me and that no salesman will call upon me.

Name.. Age.........

Address..

City.............................. Zone..... State..........

Please Check Below About Your Experience

☐ Are You Experienced? ☐ No Experience

VETERANS: My Radio Training is Approved for Veterans.

IF YOU ARE EXPERIENCED in Radio I'll qualify you for Television in 4 to 8 weeks. Rush coupon.

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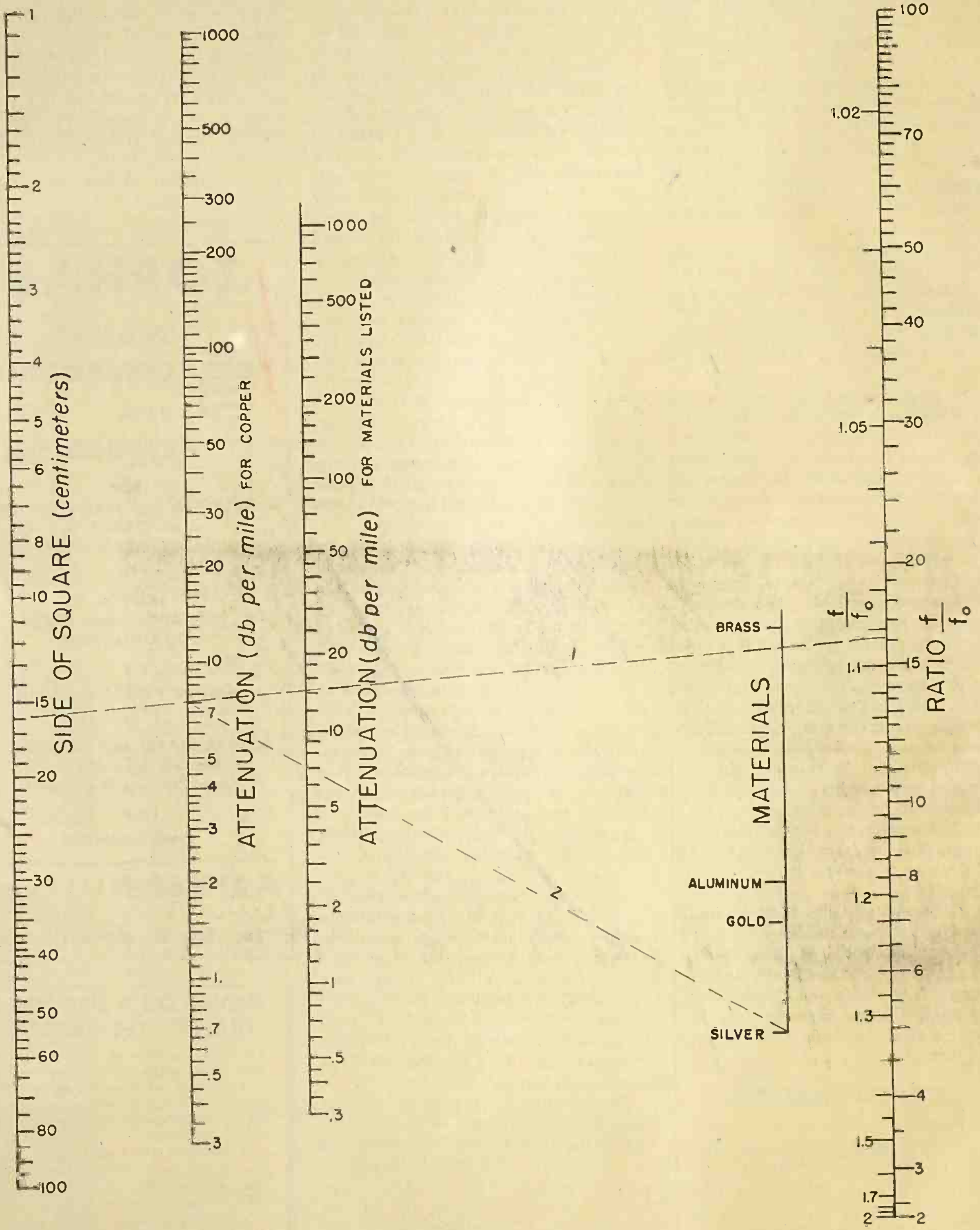

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SPRAYBERRY TELEVISION