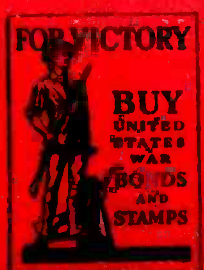
COMMUNICATIONS

★ RADIO ENGINEERING
★ LAB. U-H-F S-R RECEIVER
★ VIDEO AMPLIFIER L-F CORRECTION
★ A MERCURY VAPOR TUBE POWER SUPPLY
★ CLOSED-FORM STEADY-STATE RESPONSE ANALYSIS
★ HIGH FIDELITY SYSTEMS
★ TRANSMISSION LINES AS REACTORS



APRIL

1943

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SELECT *for* SERVICE

NEW HYTRON TUBES

Type	Description†	Price
836	Half-wave, high-vacuum rectifier	$11.50
837	12-watt, r.f. pentode	7.50
954	Sharp cut-off, acorn pentode	5.00
955	Acorn triode	3.00
956	Remote cut-off, acorn pentode	5.00
1616	Half-wave, high-vacuum rectifier‡	5.75
1625	25-watt, r.f. tetrode (12-v. heater)	3.50
1626	5-watt, triode oscillator	2.50
E1148	3.5-watt, u-h-f triode	2.25
VR105-30	Gaseous voltage regulator	1.25
VR150-30	Gaseous voltage regulator	1.25

OTHER POPULAR HYTRON TUBES*

Type	Description†	Price
2C25	15-watt, medium-mu triode	$3.00
2C45	7.5-watt, triode (modulator)	2.50
10Y	15-watt, general-purpose triode‡	1.50
801A/801	20-watt, general-purpose triode‡	2.50
HY61/807	25-watt, r.f. beam tetrode	3.50
841	15-watt, high-mu triode‡	2.25
864	Non-microphonic voltage-amp. triode	1.00
HY24	2-watt, power triode‡	1.50
HY31Z	30-watt, high-mu twin triode‡	3.50
HY65	15-watt, r.f. beam tetrode‡	3.00
HY69	40-watt, r.f. beam tetrode‡	3.95
HY75	15-watt, u-h-f triode‡	3.95
HY1148	(2C24) 1.8-watt, u-h-f triode‡	2.25
HY615	3.5-watt, u-h-f triode	2.25

*This is not a complete list.
‡Instant-heating filament.
†For complete characteristics consult Government specifications. Wattage ratings indicate maximum plate dissipation.

On this list of tubes which have recently joined the growing legions of Hytron types already marching on to Victory, you may find just the ones you want for your War equipments. Whether you choose the tiny "acorns" or the husky 1616 rectifier, you will discover the same high quality and design refinements which have made other Hytron tubes famous. If you place your orders well in advance, you will also be pleased by Hytron's on-schedule deliveries. Not too infrequently, deliveries are made from stock.

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HYTRON RADIO TUBES

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LEWIS WINNER, Editor
F. WALEN, Assistant Editor
A. D'ATTILIO, Assistant Editor

We See...

UNIVERSITY STATIONS HAVE BEEN an effective proving ground for women operators and technicians, according to the result of a recent survey. Utilizing the student bodies as a source of supply, it has been possible to study the abilities of a wide variety of applicants. While many of those selected were majoring in science, there were general-course students, who in aptitude tests showed a scientific inclination. The former were selected for the more complicated duties of the station and the latter for simple studio work. Both groups were given basic training, of course, following the ESMWT plan.

Many stations have also developed corps of women technicians using the university system, with both technical and general groups, as a source of supply. The results have been gratifying in most respects. These women technicians are serving stations in practically all capacities. In some instances, even involved engineering details have been successfully completed by them. Looks like the women are really going to town!

EVERYONE WILL BE WATCHING THAT NEW 90-day plan, which the four f-m stations in Philadelphia are putting into effect. In this unusual plan approved by the FCC, the stations will operate under a cooperative method of rotation. This will keep one station, out of the four participating, on the air each day from 3 P.M. to 11 P.M. Each station will have one regular day of the week for its programs, while remaining days will be rotated.

The stations are also planning to pool all of their spare parts and equipment which will be distributed to each station as the need arises. A committee of coordinators, composed of a representative from each station, will advise and coordinate the programs to achieve the best possible program service under the plan. This unusual step, which will conserve materials, power and manpower, is vital to our wartime program. Congratulations to the broadcasting stations, who evolved the plan.—L. W.

COMMUNICATIONS

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APRIL, 1943

VOLUME 23 NUMBER 4

COVER ILLUSTRATION

A *radioman-SPAR* member of the U. S. Coast Guard, at work in a radio laboratory. This division, organized but a short while ago, is already receiving the plaudits of the nation for its outstanding work.
(Courtesy Chicago Public Relations Office, U. S. Coast Guard)

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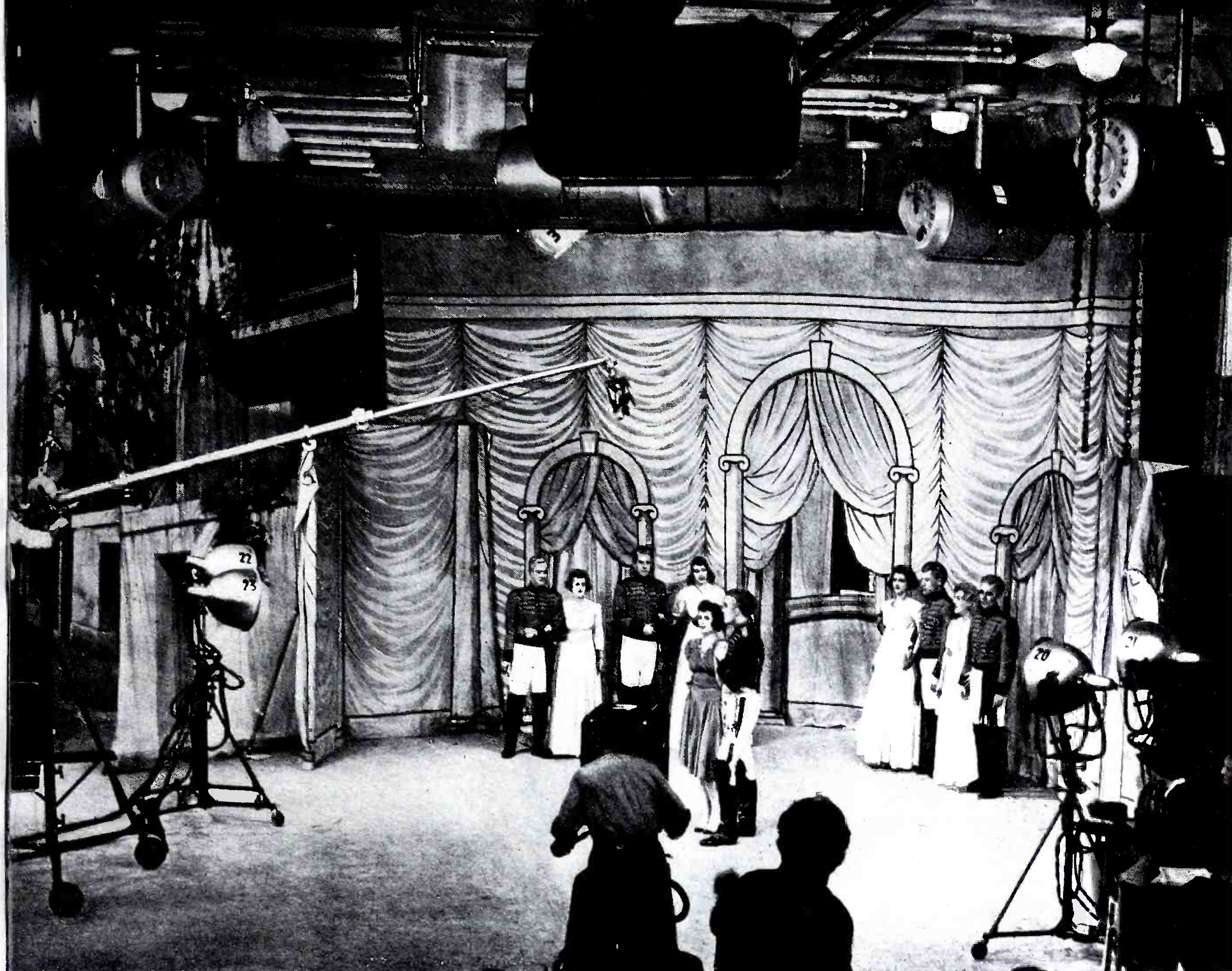

Bright but cool lights now simplify live-talent programming at G-E Television Station WRGB

Studio lighting bright as daylight . . . and cool

An Example of General Electric's Complete Service to Radio and Television Broadcasters

Once television actors had to work under hot studio lights that brought beads of perspiration through make-up, wilted costumes, caused discomfort, hindered good production.

Today at General Electric's proving-ground Television Station WRGB, Schenectady, made-up faces stay cool, costumes don't wilt. Excessive studio air conditioning is no longer necessary. For the handicap of high heat from studio lights is on its way to being completely licked by G-E developed, mercury-filled capillary lamps. They give illumination of daylight intensity with less than half the heat of the noonday sun. Water-cooled, and three to a unit, they have a light efficiency more than double — and heat output one-tenth — that of incandescent lamps. And one control panel aims a dozen of them, noiselessly, independently, anywhere in the studio.

These lamps that give cool studio lighting are another example of the bold research that will enable G-E electronics engineers to build improved cathode-ray scanning and picture tubes, cameras, transmitters, and other equipment for post-war television.

All this so that television may more quickly find its proper place in the peacetime scheme of things as a vital medium of public entertainment and education. . . . *Electronics Department, General Electric, Schenectady, New York.*

Tune in on Frazier Hunt and the News every Tuesday, Thursday, Saturday evenings over C.B.S. On Sunday night listen to the "Hour of Charm" over N.B.C. See newspapers for time and station.

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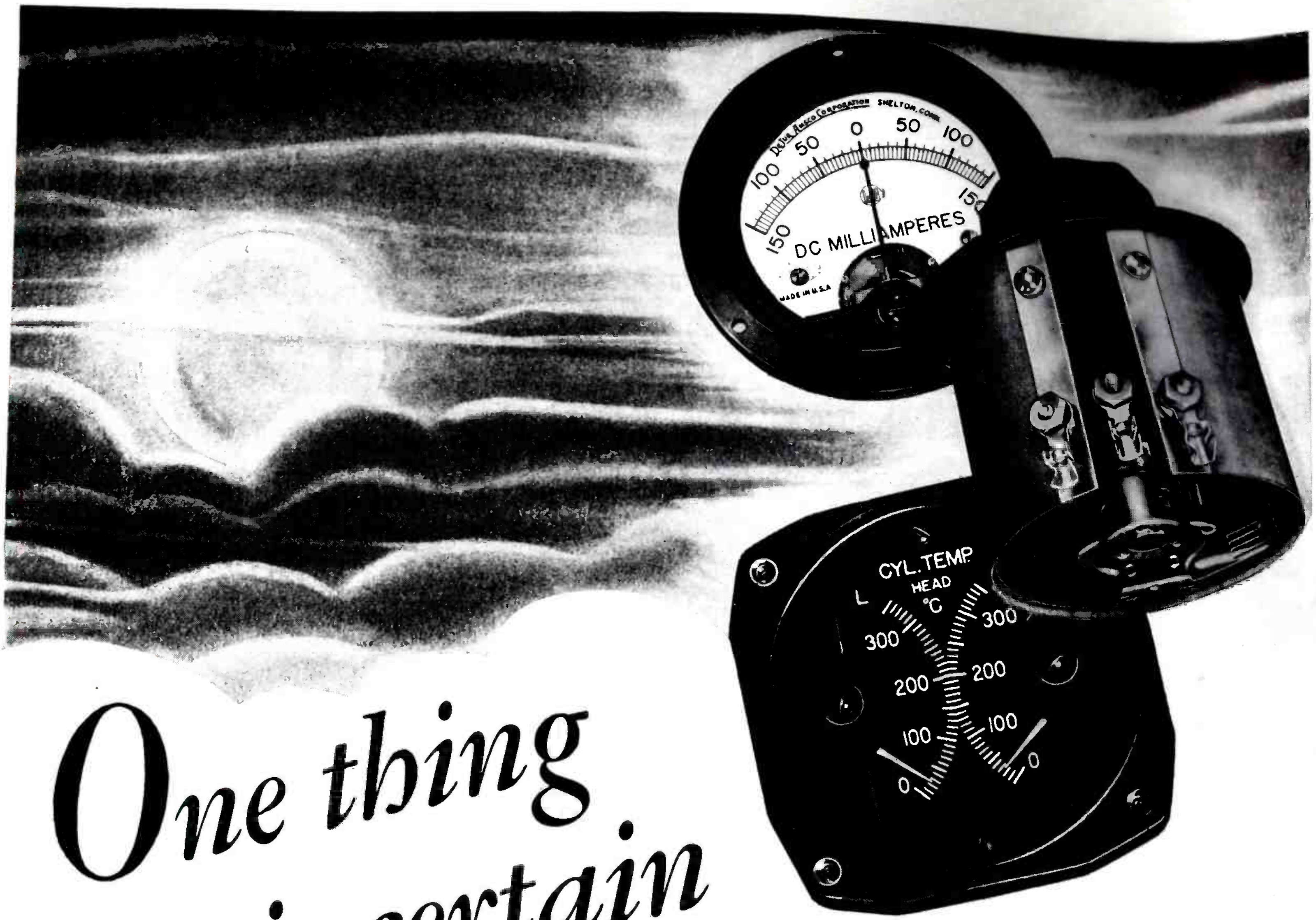

One thing is certain

. . . the American Military Machine must be equipped with superior materials of war. To this end, DeJur Aircraft and Electrical Instruments, Potentiometers and Rheostats are built to the most critical standards of quality and precision. Behind them are twenty-five years of experience and laboratory research. Before them stands the future of the democratic way of life. Neither you nor ourselves can afford to falter now.

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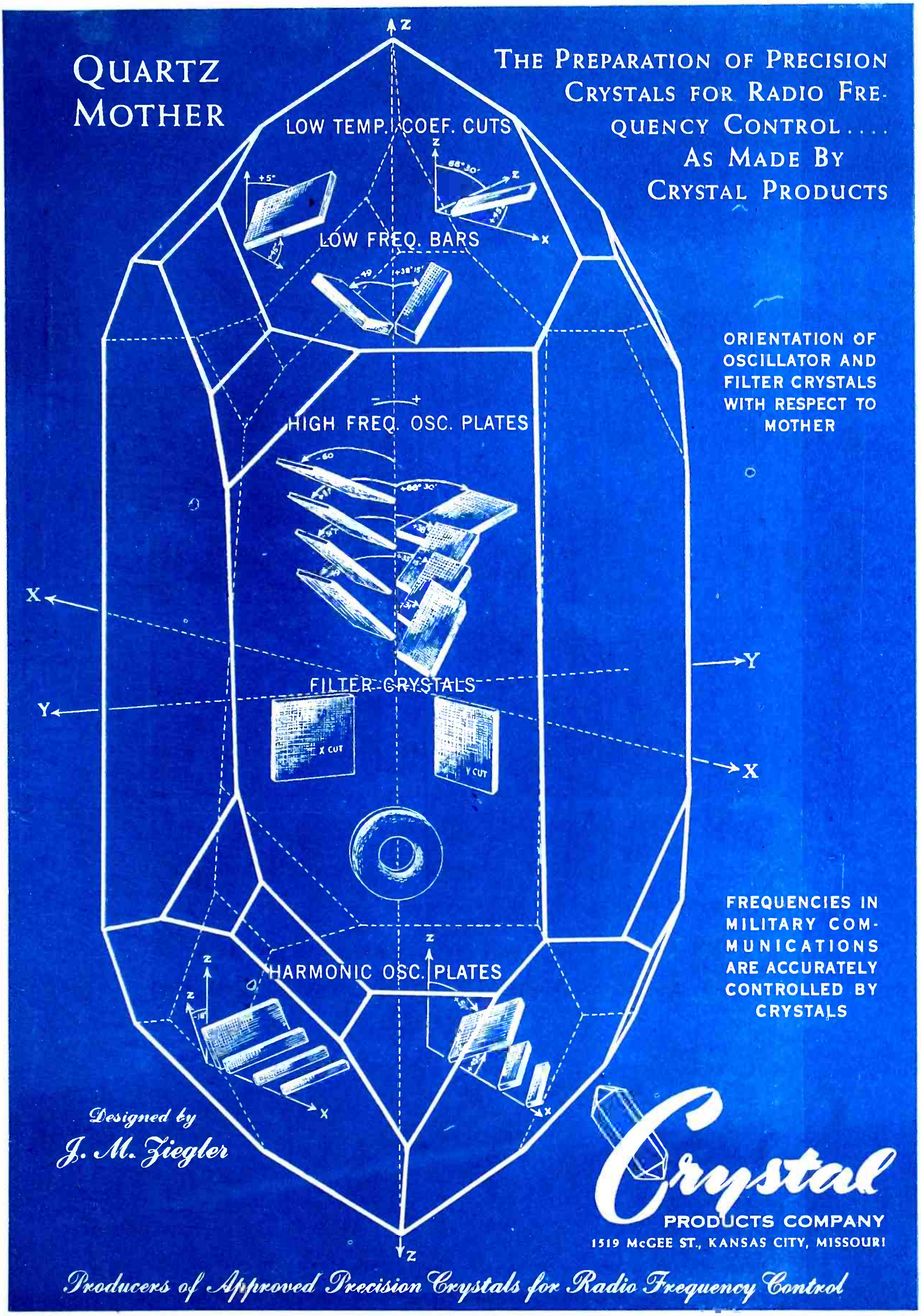

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Hallicrafters are pioneers in FM! Producers of the first general coverage U. H. F. communications receiver to incorporate both FM and AM. Time and research have added much to the performance capabilities of Hallicrafters FM-AM communications receivers . . . wartime experience is adding invaluable engineering advantages . . . all of which will be available to you in your peacetime Hallicrafters communications receiver.

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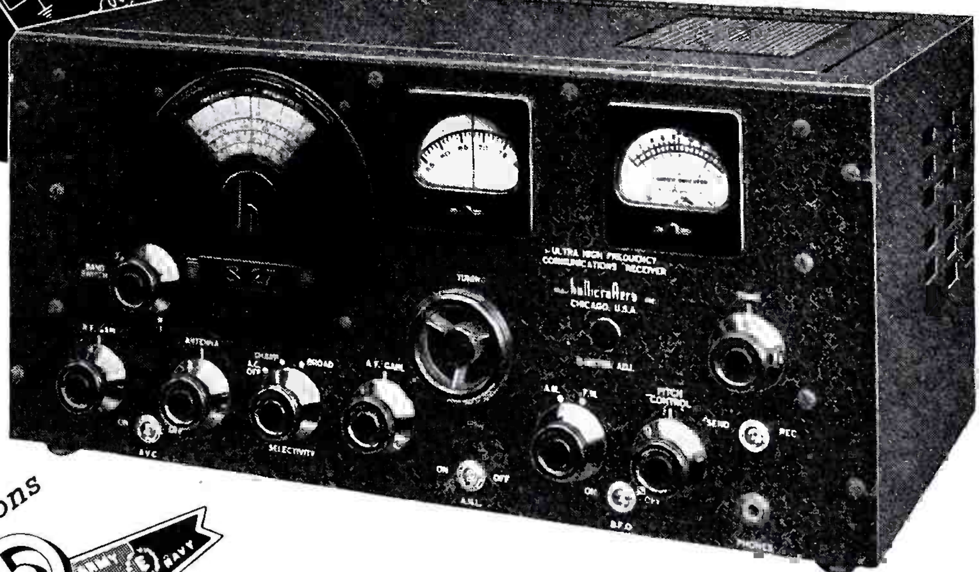

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ARMY NAVY

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If they win ...only our dead are free

These are our enemies.

They have only one idea — to kill, and kill, and kill, until they conquer the world.

Then, by the whip, the sword and the gallows, they will rule.

No longer will you be free to speak or write your thoughts, to worship God in your own way.

Only our dead will be free. Only the host who will fall before the enemy will know peace.

Civilization will be set back a thousand years.

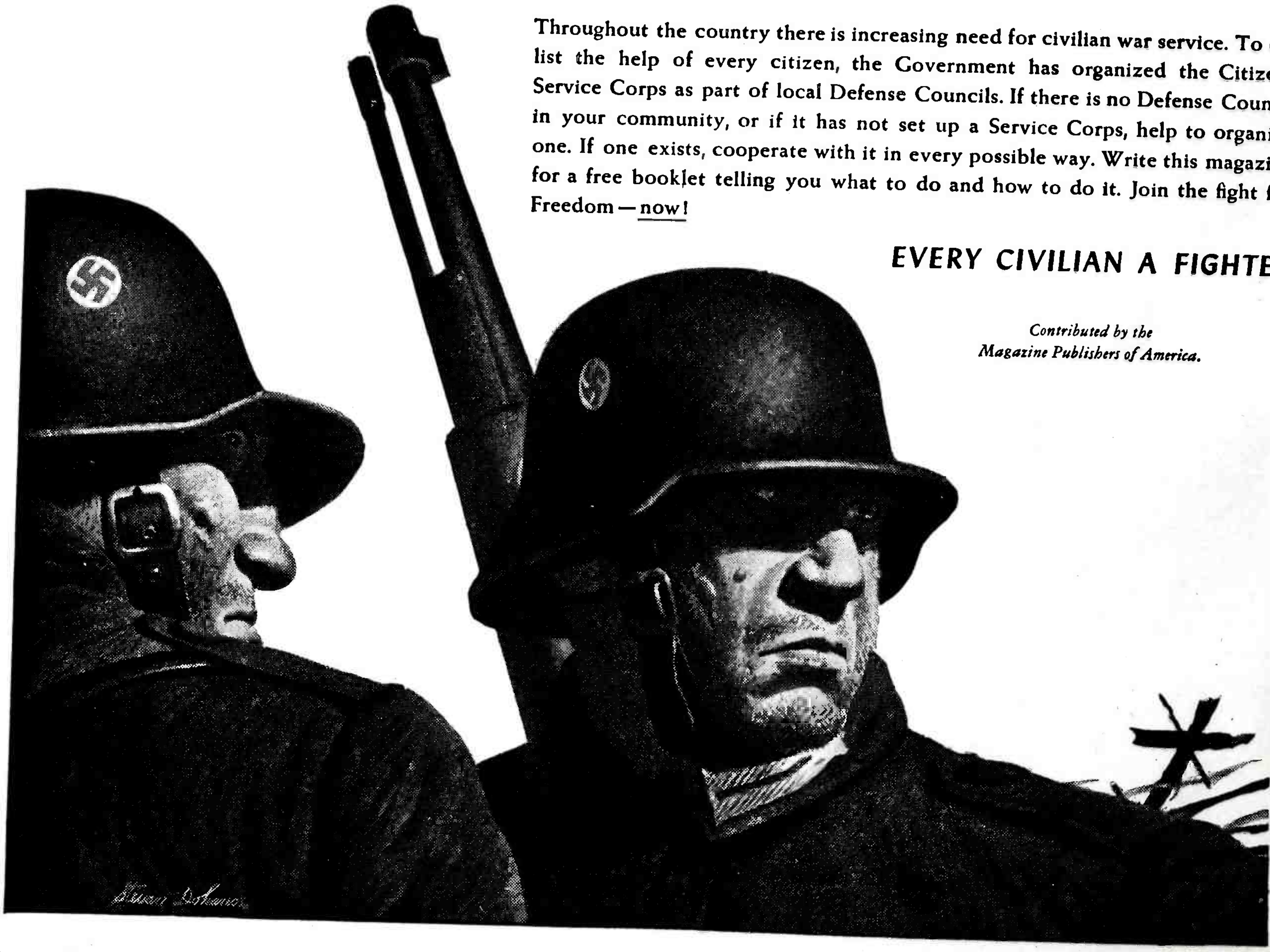
Make no mistake about it — you cannot think of this as other wars.

You cannot regard your foe this time simply as people with a wrong idea.

This time you win — or die. This time you get no second chance.

This time you free the world, or else you lose it.

Surely that is worth the best fight of your life — worth anything that you can give or do.



Throughout the country there is increasing need for civilian war service. To enlist the help of every citizen, the Government has organized the Citizens Service Corps as part of local Defense Councils. If there is no Defense Council in your community, or if it has not set up a Service Corps, help to organize one. If one exists, cooperate with it in every possible way. Write this magazine for a free booklet telling you what to do and how to do it. Join the fight for Freedom — now!

EVERY CIVILIAN A FIGHTER

Contributed by the Magazine Publishers of America.

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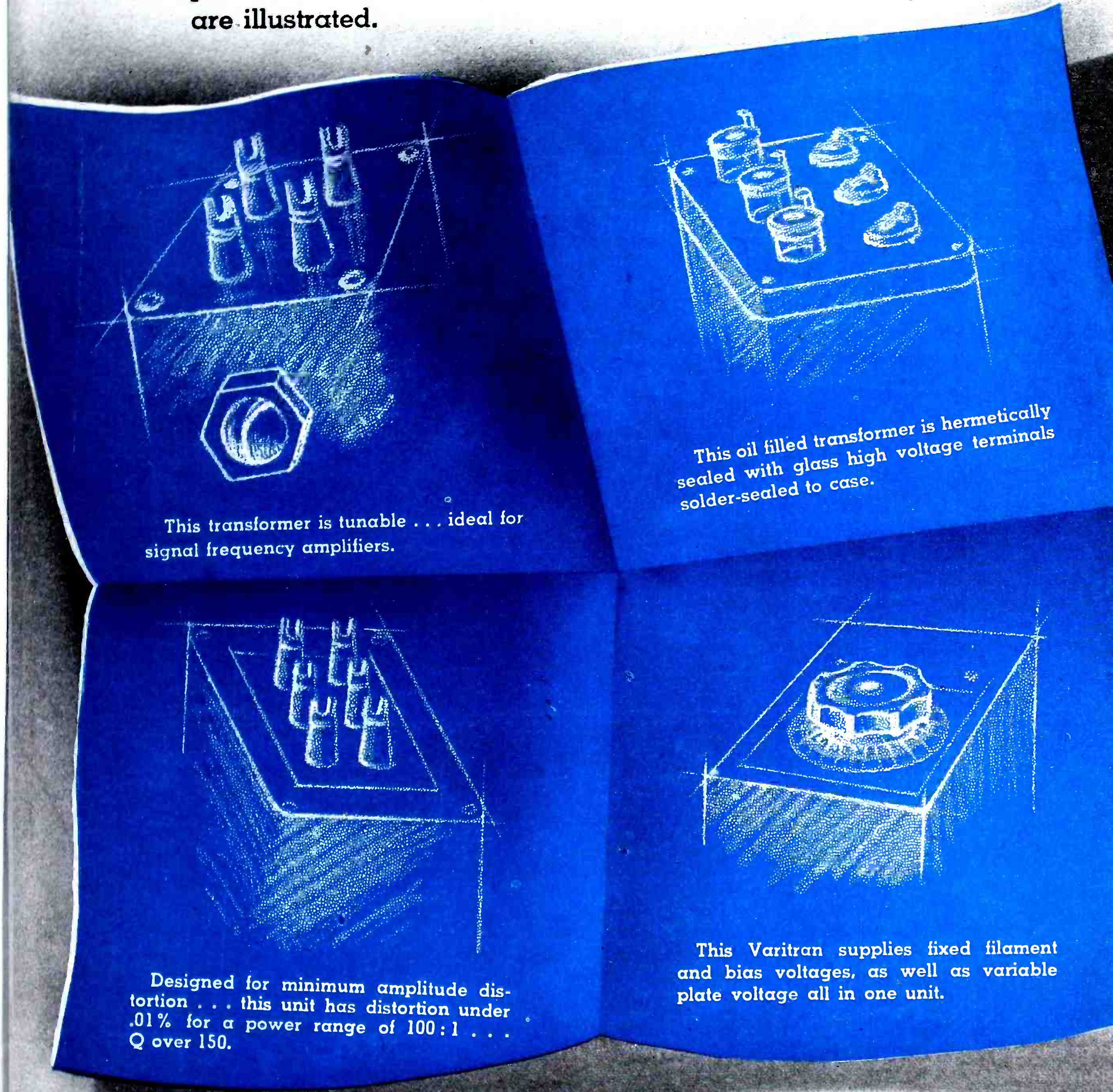
Designs for War . . . Transformers

The requirements in war transformers differ considerably from those of commercial units. The UTC engineering staff has pioneered many of the design features which make possible modern war transformers. A few typical designs are illustrated.



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How far in a FOG?

INDIAN smoke signals were an efficient means of communication over short distances in good weather, but such adverse natural conditions as fog, storms, or high winds rendered them completely ineffective. Today interference with communications is still an important problem — but it is a problem to the solution of which Breeze research and development has made and is making valuable contributions. Breeze Radio Ignition Shielding, designed to shield electrical systems from radiating or absorbing high-frequency interference, makes possible dependable radio communications for America's aircraft and tanks on fighting fronts the world over. Tested under the grueling conditions of wartime operation, the high quality and reliable performance of this equipment reflects Breeze's background of years of experience in the engineering and production of Radio Ignition Shielding.

Breeze CORPORATIONS INC. BREEZE MARK **NEWARK, NEW JERSEY**

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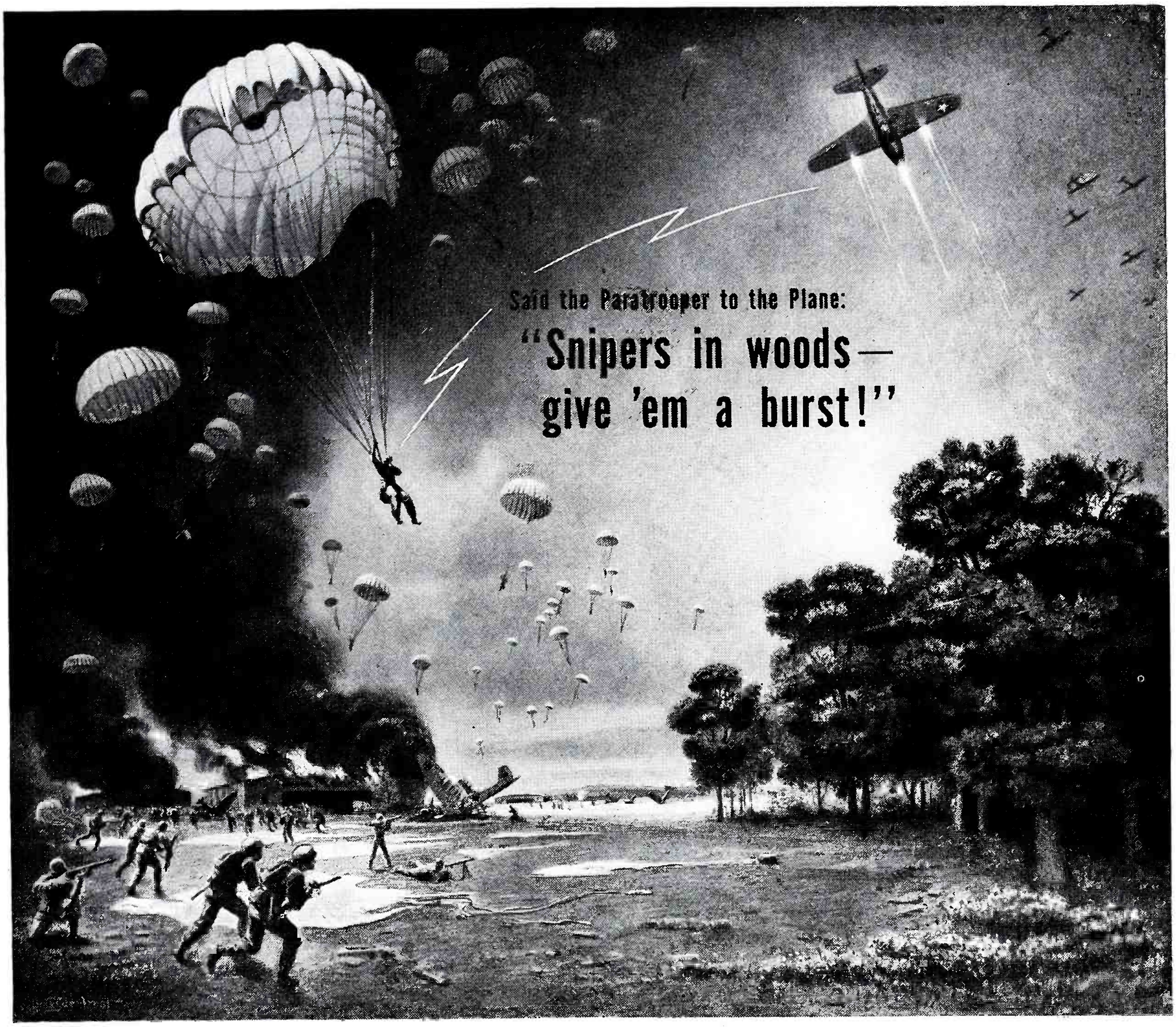

**They work together better . . .
because they can talk together**

From a thousand feet up
The burning airdrome
Looks like
A "pushover" . . .

But
When you get
Right down to earth
It turns out to be
Anything but.

Suddenly the trees
To the right
Start throwing lead—
And your men
Are still hanging
Like clay pigeons
In their harnesses.

* * *

What a break
That you're equipped
With a
Two-way
Radio.

What a break
That you can tell your trouble
To a friendly
Fighter plane.

* * *

Today, communication equipment
Designed and manufactured
By I.T.&T. associate companies
Is helping Uncle Sam's fighting forces
Work together
On land, sea and in the air . . .

Tomorrow, the broad experience
Of I.T.&T.
In the field of communications
Will help build a better world
For every man.

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International small talk . . . It doesn't actually win the battle, but hobnobbing with a foreign buddy is a form of wartime communication that builds international morale. In the picture two Americans and a French soldier tell it with gestures to an English Tommy.

via Electronics . . . International communication is doing more than any other single thing to win this war. Here the talk between soldiers ceases to be "small talk," for global strategy depends upon instantaneous communication of big ideas. Thus the radio transmitting tube becomes the greatest fighting tool ever placed at the disposal of armies.

The same inherent characteristics . . . *high performance, stamina, dependability* . . . that made Eimac Tubes first choice during peacetime have set them apart as the pre-eminent leader during this global war. Just how important and how many jobs they are doing today is a story that will be told once victory is ours. In the meantime rest assured that Eimac still remains a step ahead . . . is still first choice among the leading engineers throughout the world.

Follow the leaders to

Eimac TUBES

EITEL-McCULLOUGH, INC., SAN BRUNO, CALIF.

Awarded for high achievement in the production of war materials.

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COMMUNICATIONS

LEWIS WINNER, Editor

APRIL, 1943

AN U-H-F S-R RECEIVER

Designed For Laboratory Study

by ART H. MEYERSON

New York Fire Department Radio Laboratory

BEFORE we apply theory to practice, one fact must be noted. Super-regeneration is a mode of operation and not a particular method. This concept is important. Super-regeneration may be applied in a number of ways and it is necessary to determine the particular mode employed before attempting design for optimum performance.

There are two ways in which a detector may be self quenched.

(*1*)—By making the time constant of the grid leak and condenser so large as to cause intermittent oscillation.

(*2*)—By having the tube oscillate at both the quench and resonant frequency.

As a practical example, a receiver was designed for portable operation at 100 megacycles, incorporating an r-f stage to prevent reradiation. The receiver was designed so that it could be easily converted from a self-quenched, to a dual-oscillator quench detector. This was done in order to determine the relative merits of the two methods. The circuit was designed more as a standard for comparison of various types of s-r detectors, than for practical use. Provisions were made for variation of those constants which directly affect the operation of the detector.

The following principles were incorporated in the receiver design.

(*1*)—***Tube*** . . . Since operation was to be at u-h-f, any tube with low input capacitance would be satisfac-

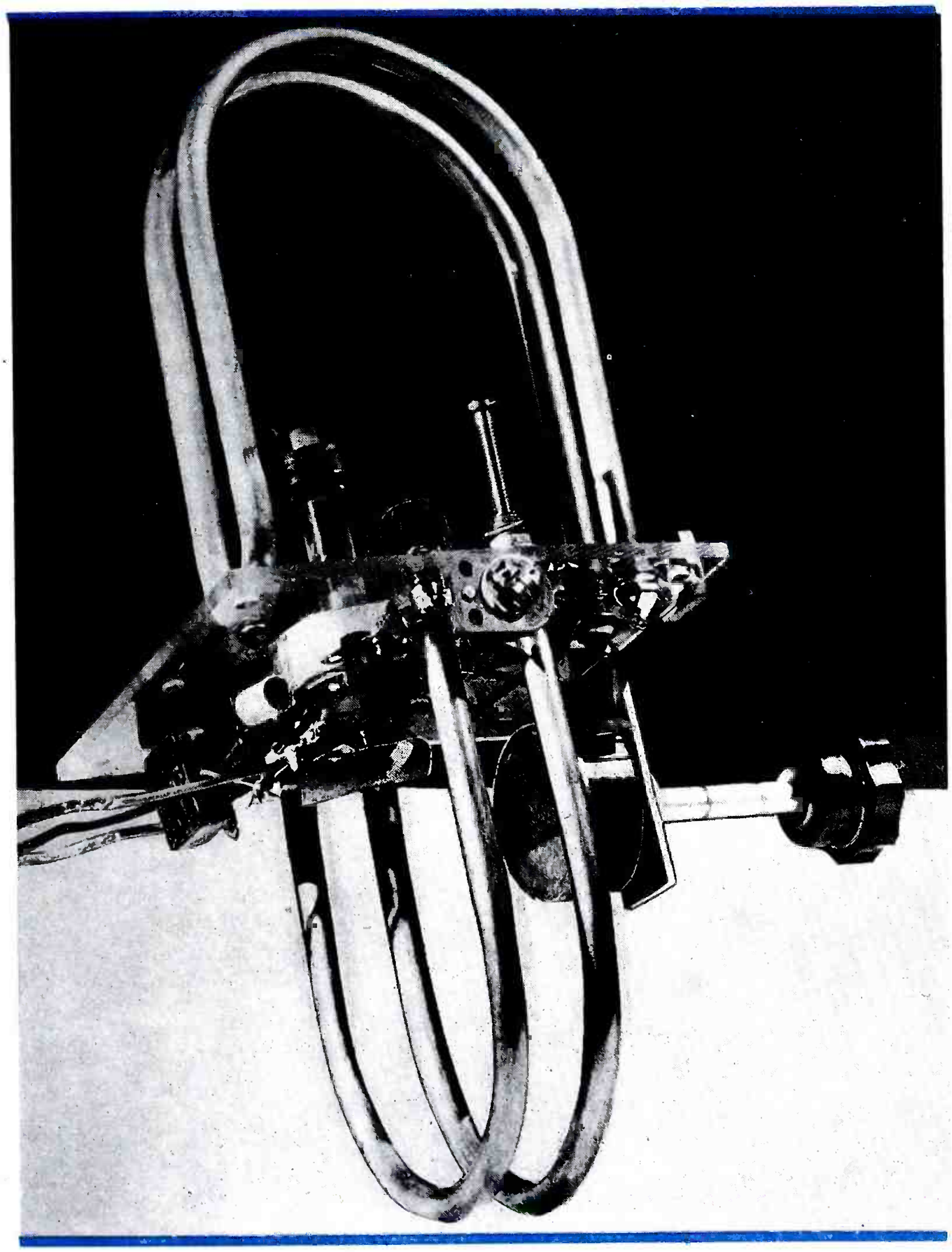

Figure 1
The detector unit. The volume control on the right is used to vary the value of grid resistor. The coil mounted on the polystyrene rod to the rear of the tube is movable, to vary the coupling of the quench oscillator coil. Adjustable screw immediately to the right of the tube, varies the capacitive coupling between the grid and plate at the resonant frequency.

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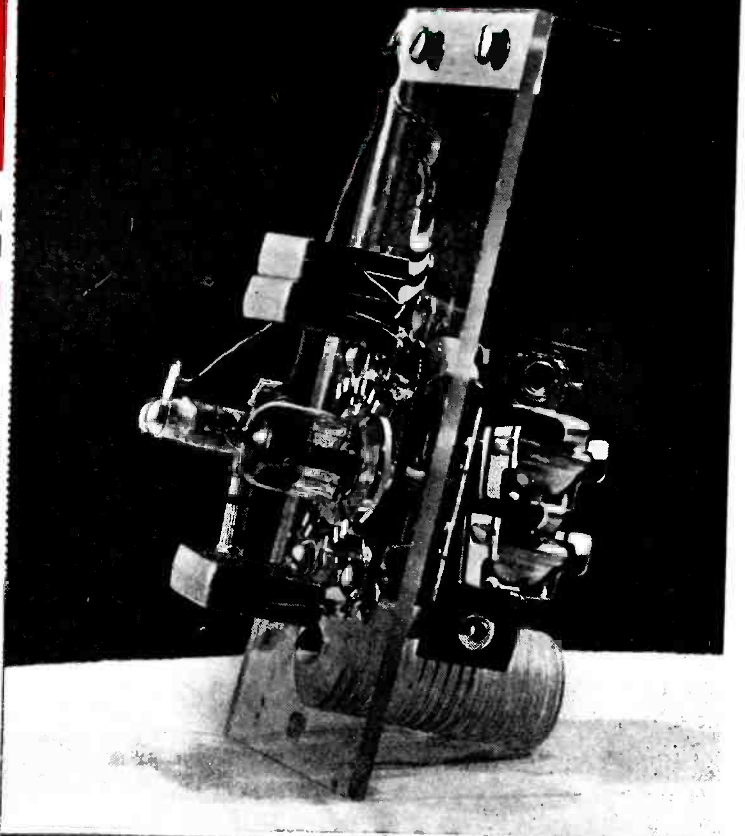

Figure 2

The r-f unit. The components are all mounted on a polystyrene strip, with the exception of the plate loop. This was done to facilitate comparison of various types of r-f stages.

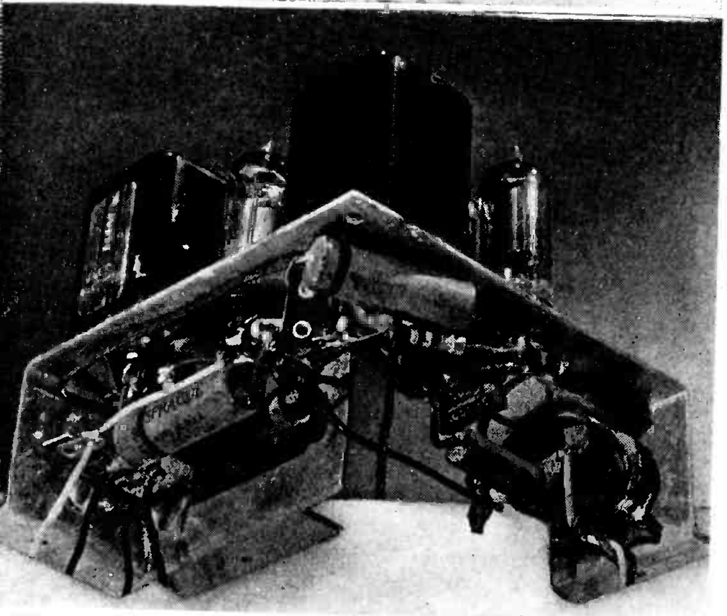

Figure 3

The audio unit. This consists of a 1S5 driver and two 1S4's in push pull, capable of delivering 550 milliwatts. The close construction was necessary for portability.

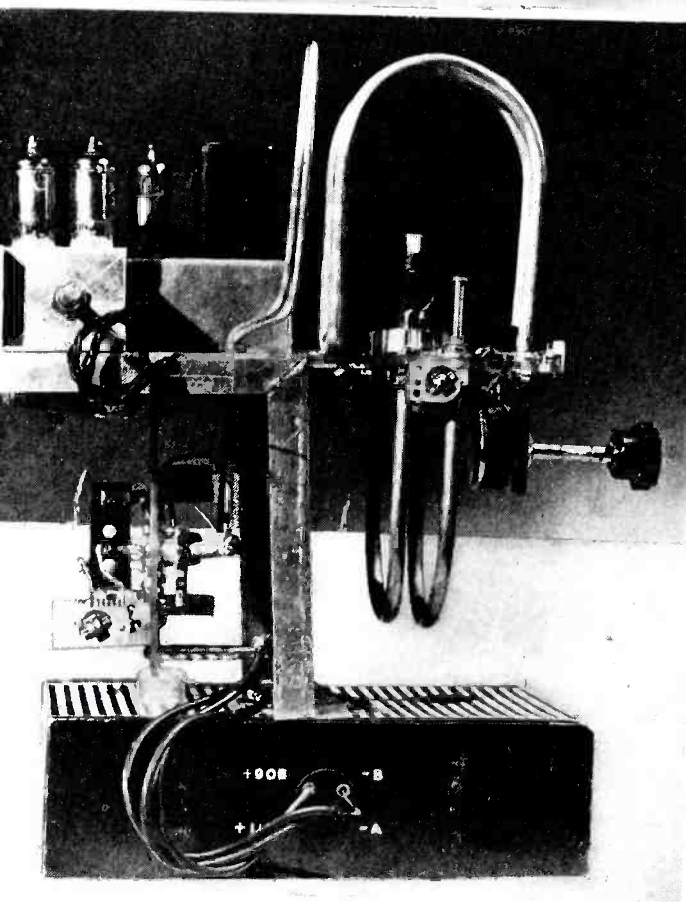

Figure 4

Here is the complete unit assembled. The coupling loop is adjustable for variable coupling between the r-f and detector stages. The weight, assembled, is less than 12 pounds, with an operating life in excess of 40 hours.

tory. For this receiver, a triode was used since it is less critical to control than a pentode.

(*2*)—***Circuit*** . . . Obviously high-Q circuits would be best. Resonant lines were used since they not only have a higher Q but also have better frequency stability than coils at u-h-f.

(*3*)—***B voltage*** . . . This was kept a constant, with the provision that the tube operate over a B range of voltage depending on the voltage regulation of the supply circuit. This is important since the design of the s-r detector may be so critical as to cause the receiver to "plop out" for slight voltage variations. For portable operation, the detector must be so designed as to operate to the end life of portable B batteries, which is usually 65 volts.

(*4*)—***R-F stage*** . . . The design of the r-f stage is not critical. Its value as far as gain or selectivity is concerned is practically nil. The main purpose behind its incorporation in the design was to prevent re-radiation. However, it was important to determine its loading effect on the detector, since this would influence the percentage of feedback.

The design of the audio system will not be discussed here, since it does not bear directly on the detector design.

Some of the components were made variable, so as to facilitate the study of their effect on super-regeneration. These were . . . (*1*)—The value of grid resistor. (*2*)—The percentage of feedback at resonant frequency. (*3*)—The quench frequency. (*4*)—The percentage of feedback at quench frequency.

All tests were conducted at maximum and minimum plate voltage.

For the particular specifications involved, the circuit shown in Figure 5 was used. This is a conventional feedback oscillator employing resonant lines as circuit elements. The receiver was constructed so that the three components . . . r-f stage, s-r detector, and the audio unit . . . were each self-contained. This was done to permit interchange of various types of units for study. The entire unit was designed for portable operation from self-contained batteries, so as to permit field tests. Incidentally the removal of the quench frequency transformer converts the detector to self-quenching.

The coupling between the grid and plate circuits of the detector was constructed, so as to allow a variation of capacitive coupling through a range of from 3 to 40 mmfd.

The quench frequency transformer coupling was made adjustable by hav-

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ing the plate coil slide on a polystyrene rod toward the grid coil.

The grid resistor was a variable 10 megohm potentiometer. And the r-f stage was variably coupled to the detector. All tests were conducted at 100 megacycles.

First trials were made with the self-quenched type of detector. This was done by shorting out the grid coil of the q-f transformer, using the plate coil as an r-f choke.

Self Quenched Detector

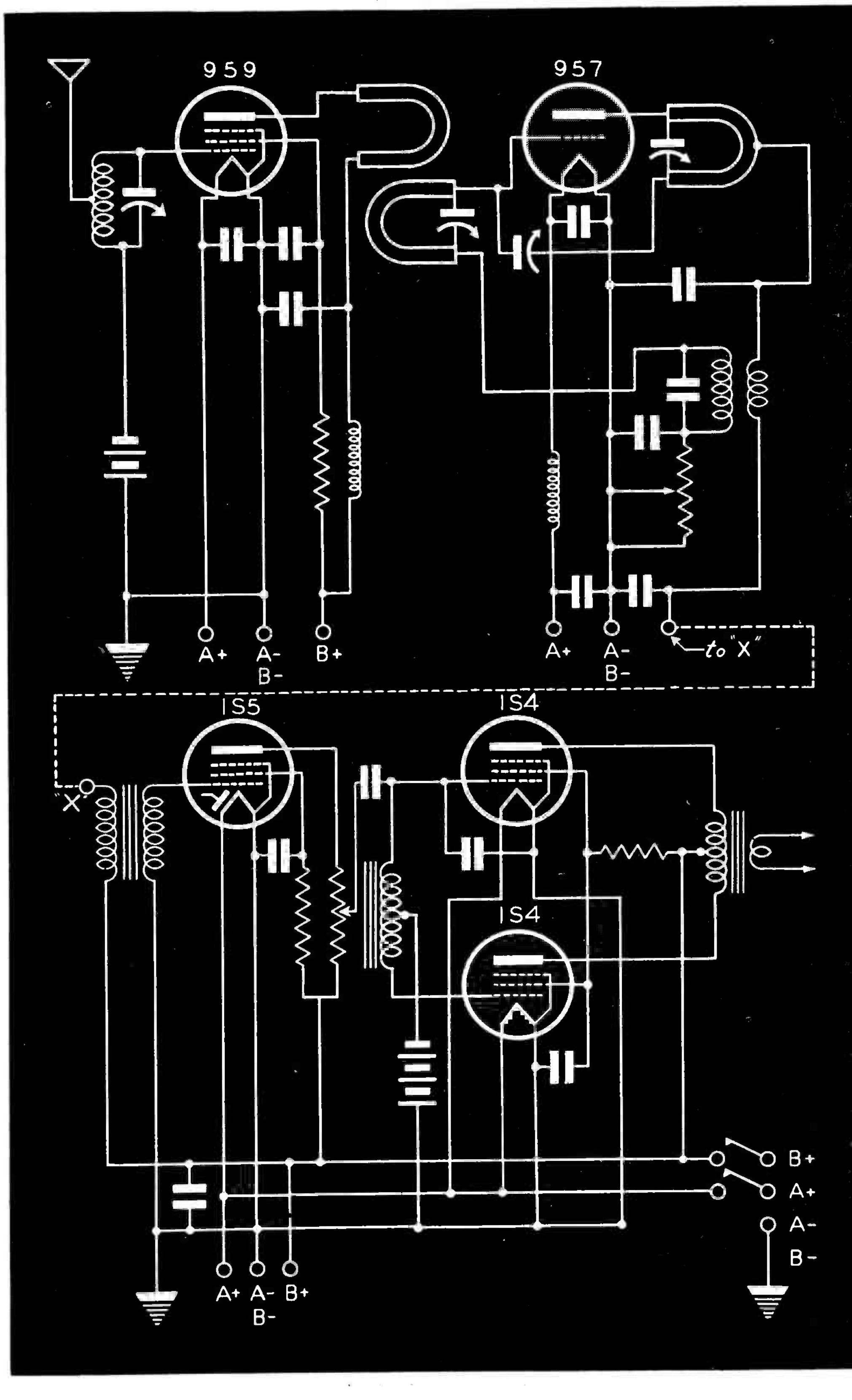
In the self-quenched detector, the effect of a signal is to increase the frequency of quench. This can readily be determined by adjusting the value of r-c in the grid circuit, so that the quench frequency comes within the audible range. Introducing an external signal into the grid circuit will increase the pitch. This increase will be rapid for small values of signal and slower as the signal increases in amplitude. The change in quench frequency affects the bias voltage, since grid current will flow more often as the frequency increases. Thus variations in signal amplitude will be reflected in the grid bias voltage and be reproduced in the plate current.

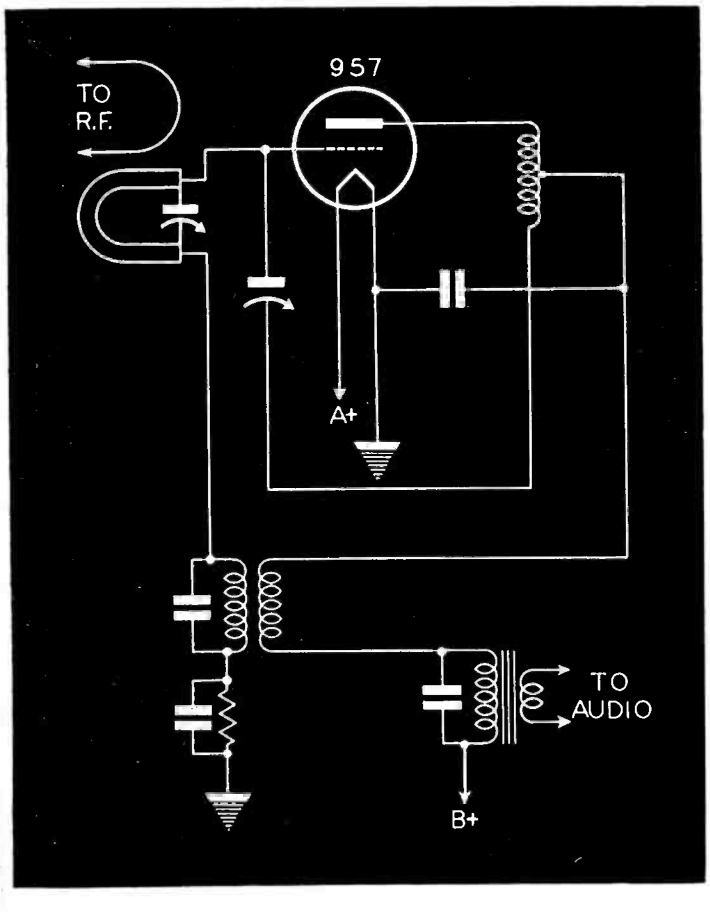
In the self-quenched detector, the quench frequency is determined by the time constant of the grid leak and condenser. The relationship of resistance to capacity is important, for the quench frequency will decrease for any increase in either resistance or capacity. Therefore, in order to increase the quench frequency it is necessary to decrease either resistance or capacity. This poses a problem, when it is desired to increase the quench frequency to some higher value, since, for a given capacity, the maximum quench frequency possible will be determined by the lowest value of resistance which will cause the oscillator to superregenerate. To partly overcome this condition, the grid condenser should be reduced to the lowest possible value. The limiting factor is the input capacitance of the tube used. The

(*Continued on page 67*)

Figures 5 (right) and 6 (bottom)

In Figure 5, we have the circuit used for the field model shown in Figure 4. The tuning range was 75-118 megacycles. Good shielding of the receiver plus the addition of the r-f stage prevented any re-radiation. Figure 6, shows an alternative detector unit. Results are comparable to those obtained with the detector shown in Figure 5. The selectivity is broader with this circuit.





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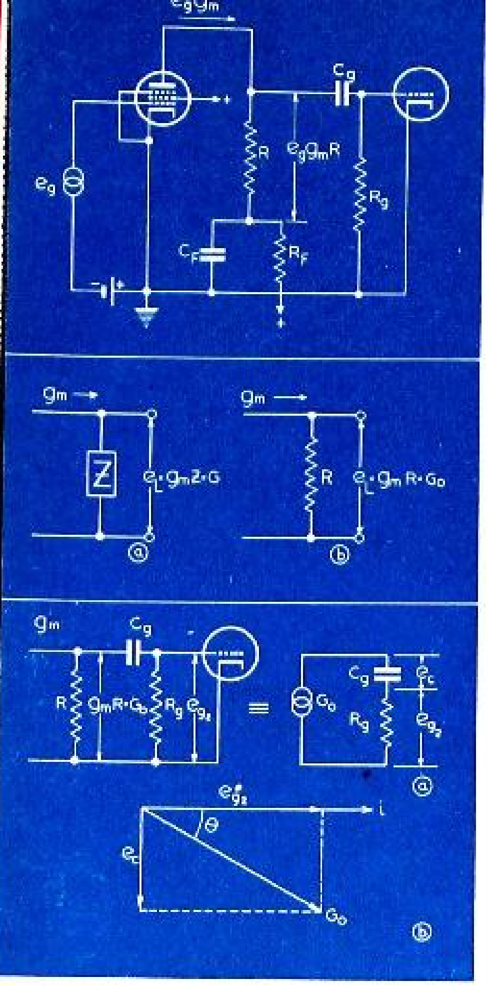

VIDEO AMPLIFIER L-F CORRECTION

An Analysis of Plate Filter Compensation

by WILLIAM A. LYNCH

Instructor, Graduate Electrical Engineering Department,
Polytechnic Institute of Brooklyn

•

Figures 1 (top), 2 (center), 3 (bottom)

In Figure 1, the pentode amplifier stage, showing the low-frequency compensation circuit, R_FC_F. Figure 2 illustrates the equivalent circuit for a pentode amplifier, (a) at any frequency, and (b) in the mid-frequency region. Figure 3 (a), the equivalent circuit, including the grid coupling circuit for the following stage. In (b), the phasor diagram at low frequencies.

•

This paper, specially prepared for COMMUNICATIONS is based on material given in a course by the author in "Cathode Ray Tubes and Circuits."

THE resistance-coupled amplifier exhibits a departure from an ideal characteristic in both the high- and low-frequency regions of its spectrum. At low frequencies, from a few hundred cycles on down, the amplitude vs frequency response falls off and the output voltage assumes a leading phase angle. Where fixed bias and screen voltages are used, these departures may be assumed to be attributable entirely to the grid-coupling circuit acting as a resistance-capacity voltage divider across the load circuit.

The response at low frequencies depends upon the size of the grid condenser and the grid resistor. Increasing the size of either or both improves the low frequency response, and were it not for practical limitations in the choice of these components, there would be little need for low-frequency compensation. It so happens, however, that practical values of R_g and C_g are usually such that the amplifier seriously discriminates against low frequencies lying within the required pass band. It is necessary, therefore, to correct the low-frequency characteristic, and in the case of amplifiers used to transmit visual intelligence, the correction for phase shift is of even greater importance than the maintenance of constant gain.

A simple expedient widely used to effect the correction, is the addition of a resistance R_F in series with the load resistor, R, and across which is shunted a condenser, C_F, as shown in the circuit of Figure 1. This is the familiar plate circuit filter, serving in this case, the two-fold purpose of providing low-frequency compensation, and decoupling. Because of its simplicity and effectiveness, a rather detailed discussion of this single expedient will follow.

Plate filter compensation does not provide perfect correction for all low frequencies, but if the circuit components are properly chosen, the amplifier characteristics can be adjusted to satisfy a rather wide range of response requirements. It is the purpose of this article to outline methods of determining the proper proportionality of the components from a consideration of the mathematical expressions for the response characteristics. The analysis is developed in terms of variables expressed in normalized form, and in terms of two disposable constants which determine the proportionality of the filter elements. The conclusions which are reached serve as a simple and accurate guide to correct design, and are applicable to any amplifier circuit meeting the initial assumptions. These are (*1*) that fixed bias and screen voltages are employed, (*2*) that the magnitude of the plate load impedance is very much smaller than the tube resistance, and (*3*) that the grid resistor of the following stage is very much larger than the magnitude of the plate load. The second and third assumptions are generally met in wide-band pentode amplifiers.

Before proceeding with the analysis of the compensated circuit, it may be well first to examine the characteristics of the uncompensated amplifier at low frequencies. This will provide a basis for comparison, and at the same time will serve to introduce the writer's terminology. The low-frequency equivalent plate circuit for a pentode amplifier is shown in Figure 2. Shunt capacities and high-frequency correction devices have been omitted because they are inoperative in the range of low frequencies under consideration. Th pentode tube is represented as a constant current generator, whose output current, with a unit applied signal voltage, is simply, g_m amperes (where g_m is the grid-plate transconductance; a constant with frequency). The plate resistance of the tube, normally appearing in shunt with the load in the equivalent circuit, has likewise been omitted because of the initial assumption.

The voltage gain at any frequency can be expressed very simply for a pentode as

$$G = e_L/e_g = g_mZ \qquad (1)$$

Where Z represents a general complex load impedance (Figure 2a). Within the mid-frequency region of the spectrum where the reactive elements of the circuit have a negligible effect, the load impedance is purely resistive and is equal to the plate load resistor, *R* (Figure 2b). Throughout this region, the amplifier is inherently linear and the voltage gain is a constant quantity, g_mR. We shall refer to this mid-frequency value of gain as the ***nominal gain*** of the amplifier, and designate it by the symbol, G_0.

In making an analysis of amplifier characteristics, it is desirable to think

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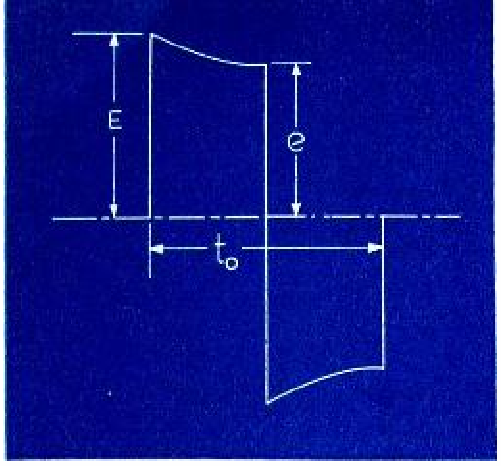
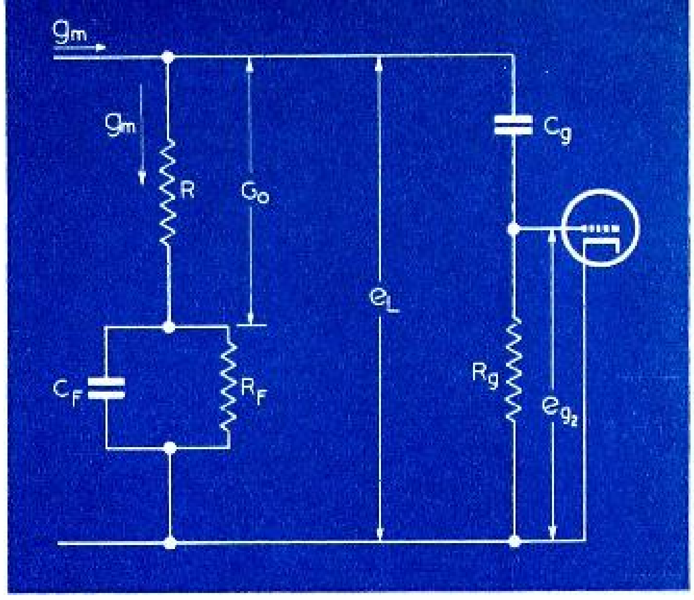


Figures 4 (left) and 5 (right)
In Figure 4, we have the appearance of the voltage e_{g2} with a square-wave input when a small amount of distortion is present. In Figure 5 is the equivalent circuit of the compensated amplifier.



in terms of normalized or relative values of the variables, thereby generalizing the analysis and broadening its applications. With this in mind, we shall use the nominal gain as a reference value and express all values of gain as ratios. By dividing each value of gain by G_0, we obtain the *relative* gain which we shall designate by *H* and define in any of the following equivalent forms . . .

$$H = G/G_0 = g_m Z/g_m R = Z/R \qquad (2)$$

In general, the relative gain is a complex numeric which may be written in any of the equivalent forms of complex notation . . .

$$H = \begin{cases} a + jb \\ |H|\,(\omega)\,\varepsilon^{j\theta(\omega)} \\ |H|\,(\omega)\,\underline{/\theta(\omega)} \end{cases} \qquad (3)$$

$|H|$ is the absolute magnitude of the gain and is defined numerically as

$$|H| = \sqrt{a^2 + b^2}$$

It is equal to unity in the mid-frequency region, and is less than unity for all values of gain smaller than the nominal gain. θ is the phase angle and is found from the following relationship . . .

$$\theta = \tan^{-1}\frac{b}{a}$$

Both quantities are indicated as being functions of frequency, or of (ω). θ is understood to be the *residual* phase angle which remains after subtracting the normal 180° shift of phase between the grid and the plate voltages, from the total phase angle.

The equivalent circuit including the grid-coupling elements for the following stage, is shown in Figure 3, together with the phasor* diagram indicating the phase relationships at low frequencies.

It will be noted that the voltage, e_{g2}, applied to the succeeding grid, leads the input voltage, e_g (the load voltage G_0 and e_g being considered to have no residual phase difference).

As the frequency is lowered from the mid-frequency region, the reactance of the grid condenser becomes appreciable, and increases until at a specific frequency, it becomes exactly equal to the grid resistance, R_g. At this point, due to the fact that the voltage drops across C_g and R_g are equal and in quadrature, the relative gain is 0.707 and the phase angle, 45°. We shall identify this frequency as the *nominal low-frequency cutoff* and designate it by the symbol, f_0.

By definition,

$$R_g = \frac{1}{2\pi f_0 C_g} \qquad (4)$$

hence

$$2\pi f_0 = \omega_0 = \frac{1}{R_g C_g} = \frac{1}{T_g} \qquad (5)$$

T_g represents the time constant of the grid circuit, $R_g C_g$. Similarly, any radian frequency, ω, may be written as $1/T$, where *T* is defined as the *radian period*. This is not to be confused with the usual conception of the period which is the reciprocal of the frequency. We are now prepared to assign a scale of relative values for the independent variable. By converting frequencies to radian periods and dividing each radian period by T_g, we define a *relative period* which is also in ratio form . . .

$$T/T_g = \omega_0/\omega = f_0/f \qquad (6)$$

This is a convenient unit to use inasmuch as the analysis deals largely with *RC* time constants.

When it comes to plotting the amplifier characteristics, the independent variable, which is the relative period, is plotted as the abscissa, while the dependent variable, which is $|H|$ or θ, as the case may be, is plotted as the ordinate. The choice of the relative period, rather than the frequency ratio, for the independent variable, results in locating infinite frequency $(T/T_g = 0)$ at the origin, with frequency decreasing in the positive direction (toward the right). Thus the frequency is zero when T/T_g becomes infinite. While this is the reverse of the conventional representation, it is advantageous for the mathematical methods to be employed, wherein it is required that the impedance function be expanded about infinity.

It has been pointed out that the grid-coupling circuit acts as a voltage divider across the load circuit. The voltage e_{g2} which is transmitted to the grid of the following tube is

$$e_{g2} = G_0 \frac{R_g}{R_g + \dfrac{1}{j\omega C_g}} = G_0 \frac{1}{1 - jT/T_g}$$

The relative gain up to the next grid is then

$$H = e_{g2}/G_0 = \frac{1}{1 - jT/T_g} \qquad (7)$$

Writing this in magnitude and phase angle notation . . .

$$|H|^2 = \frac{1}{1 + (T/T_g)^2} \qquad (8)$$

$$\theta = \tan^{-1} T/T_g \qquad (9)$$

The absolute magnitude of the gain may also be written as a function of the phase angle; a convenient form for purposes of plotting the response curve . . .

$$|H| = \frac{1}{\sqrt{1 + \tan^2\theta}} = \cos\theta \qquad (10)$$

The transient response is found by applying to the input, a unit voltage step, $e_g(t)$; i.e., a voltage which rises instantaneously to a value of 1 volt and remains at that value thereafter. The load voltage is an amplified replica of the step which rises suddenly to G_0 volts. The voltage e_{g2},

*Often incorrectly called a vector.

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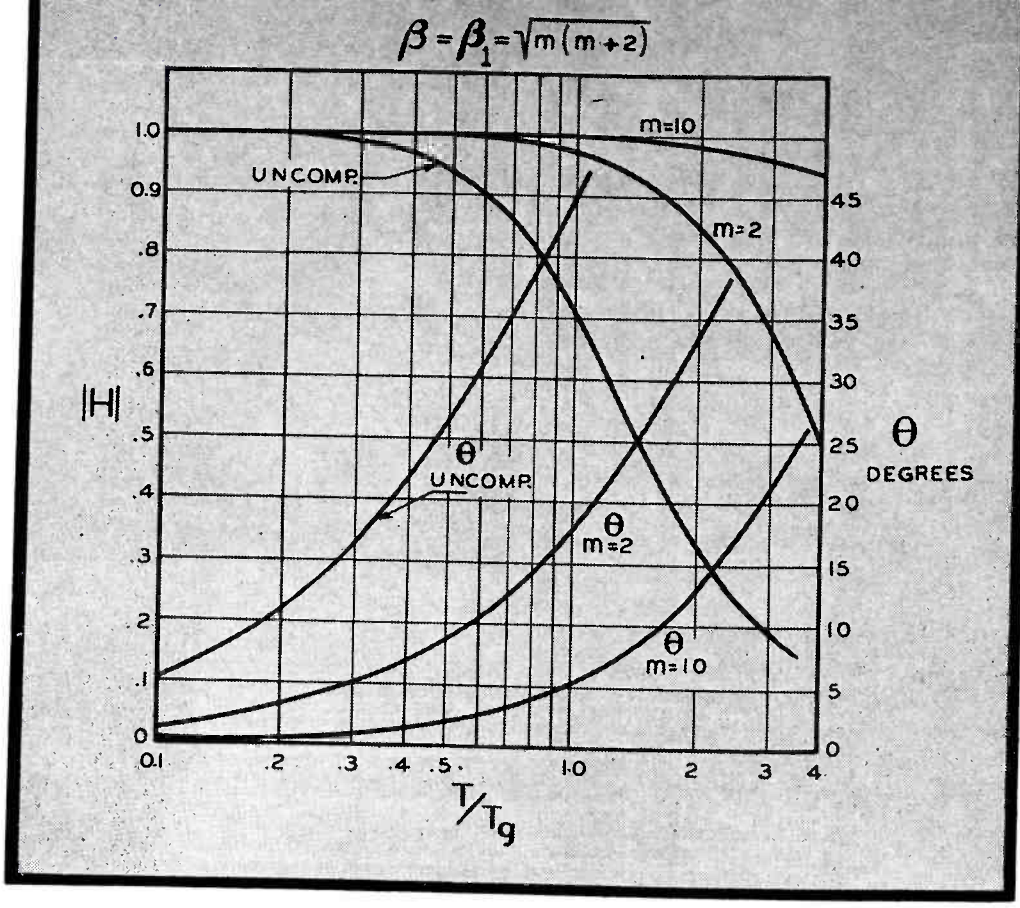
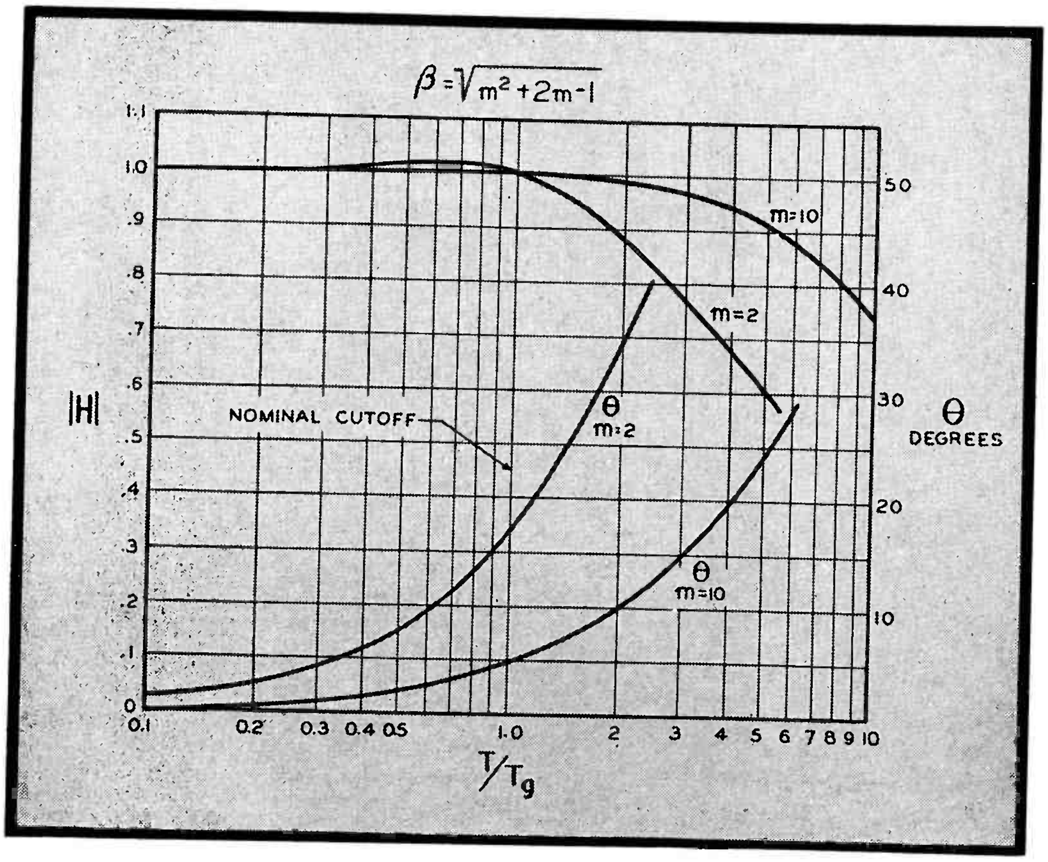


Figures 6 (top) and 7 (bottom)

Figure 6 illustrates the phase and amplitude response curves when $\beta = \beta_1$. The curves for the uncompensated amplifier are shown for comparison. Figure 7 shows the phase and amplitude response curves when the gain is arbitrarily adjusted to unit value at the nominal cutoff.



which is passed on to the following grid, is G_0 minus the voltage drop across the condenser, C_g. If time is measured from the instant the step occurs, the voltage across the condenser will be zero when t is zero, and will build up exponentially as the condenser charges. Consequently, the voltage e_{g2} starts out with a value equal to G_0 when t is zero, and decays exponentially, approaching zero value asymptotically. The voltage e_{g2} is expressed mathematically as follows . . .

$$e_{g2} = G_0 - G_0\,(1 - \varepsilon^{-t/R_gC_g}) \qquad (11)$$

Expressing the transient response to a unit voltage step as a relative quantity, and as a function of time

$$H(t) = e_{g2}/G_0 = \varepsilon^{-t/T_g} \qquad (12)$$

where t is in seconds, measured from the instant of application of the step voltage.

Instead of using a unit voltage step, let us consider the results of applying a unit square wave to the amplifier input. We shall define a unit square wave as one whose voltage excursions above and below the average value are one volt and of equal time duration. Low frequency distortion of the square wave causes the top and bottom of the wave to tilt toward the average-value axis, and while this slope appears to be linear, it is actually exponential, as determined from Equation 12.

It might prove interesting to discover whether the distortion of the square wave at low frequencies is caused principally by phase shift or by loss of gain. Figure 4 illustrates the appearance of the voltage e_{g2} when a unit square wave is impressed across the the input terminals and a small amount of distortion is present. Call the initial voltage rise E and the voltage at the end of the half-cycle, e. If t_0 is the period of the square wave, the fractional deviation, δ, at the end of the time interval $t_0/2$, may be defined as

$$\delta = \frac{E - e}{E} \qquad (13)$$

but $E - e$ represents the voltage rise across the condenser C_g therefore

$$\delta = (1 - \varepsilon^{-t_0/2T_g}) \qquad (14)$$

If $t_0/2T_g$ is a small fraction, as it will be for small deviations, the exponential function in parenthesis may be expanded into an infinite series and the higher order terms neglected. We may then write

$$\delta \cong t_0/2T_g = \frac{2\pi T}{2T_g} = \pi\frac{T}{T_g} \qquad (15)$$

where T is the radian period of the square wave and equal to $t_0/2\pi$. For small angles, the tangent of the angle is approximately equal to the angle itself, expressed in radians, hence

$$\delta \cong \pi\,\theta \qquad (16)$$

Suppose, for example, at a certain low frequency, with a sine-wave input, an amplifier exhibits a phase shift of 1° or 0.0174 radian. If a square wave of the same frequency is now applied to the amplifier, we find from equation 16 that δ is about 5½%, which means that the wave top has decayed 5½% by the end of the half-

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cycle. Now let us determine the value of the relative gain with a sinusoidal input when θ is 1°. From equation 10, it is only necessary to find the cosine of 1°, which is 0.99985. Hence the gain has dropped only 15/1000 of 1%!

The above example illustrates the fact that the achievement of a flat amplitude response over the usable low-frequency region is no guarantee that the square-wave response will be acceptable over the same region. This can also be seen from the characteristics of the trigonometric functions involved. The cosine varies but slightly in the vicinity of zero degrees, whereas the tangent exhibits a high rate of change in the same vicinity. Hence the distortion of the square wave is determined primarily by the amount of phase shift, and for small phase angles, the deviation can be estimated to be roughly 5½% per degree.

The Two Conclusions

Two conclusions may be drawn from this; (*1*) that it is more important to try to prevent phase shift at low frequencies than to strive for flat amplitude response, and (*2*) that the criterion of flat amplitude response is of little consequence in judging performance if the amplifier must satisfactorily transmit a visual signal which is non-sinusoidal.

The Steady-State Response of the Compensated Amplifier

Figure 5 shows the equivalent circuit of the compensated amplifier. The expression for the plate load impedance, which now includes the plate filter network, can be written in the following form, after some algebraic rearrangement . . .

$$Z = R \cdot \frac{1 - j\left(1 + \frac{R_F}{R}\right)\frac{1}{\omega R_F C_F}}{1 - j\frac{1}{\omega R_F C_F}} \qquad (17)$$

This can be simplified by making some substitutions. First, let $m = R_F/R$, and $\beta = T_F/T_g$. These are the two disposable constants mentioned earlier which determine the proportionality of C_F and R_F. Then substitute T_F for R_FC_F, the time constant of the filter network, and T for $1/\omega$. The overall relative gain is then obtained by multiplying Z/R by the grid dividing ratio, (Equation 7). The result may be written in magnitude and angle notation as follows:

$$|H|^2 = \frac{1 + \left(\frac{1+m}{\beta}\right)^2 \left(\frac{T}{Tg}\right)^2}{1 + \frac{1+\beta^2}{\beta^2}\left(\frac{T}{Tg}\right)^2 + \frac{1}{\beta^2}\left(\frac{T}{Tg}\right)^4} \qquad (18)$$

$$\theta = \tan^{-1} \frac{\frac{\beta - m}{\beta}\left(\frac{T}{Tg}\right) + \frac{1+m}{\beta^2}\left(\frac{T}{Tg}\right)^3}{1 + \frac{1+m+m\beta}{\beta^2}\left(\frac{T}{Tg}\right)^2} \qquad (19)$$

Note that the above expressions reduce to those derived for the uncompensated amplifier, when *m* is zero and β is unity.

These two equations are not particularly useful as they stand, but they provide a basis for determining suitable relationships between the constants. The equations are equivalent to the usual expressions derived for this type of compensation, only they have been developed in such a form that rather simple notions of mathematics can be applied to obtain critical values of the constants, *m* and β. By establishing these critical values, we can determine the limits within which the constants should be varied for a specific range of response characteristics and also establish the effect of the direction of variation.

Let us determine first the values of the constants to produce a flat amplitude response. Referring to Equation 18, the expression for the amplitude

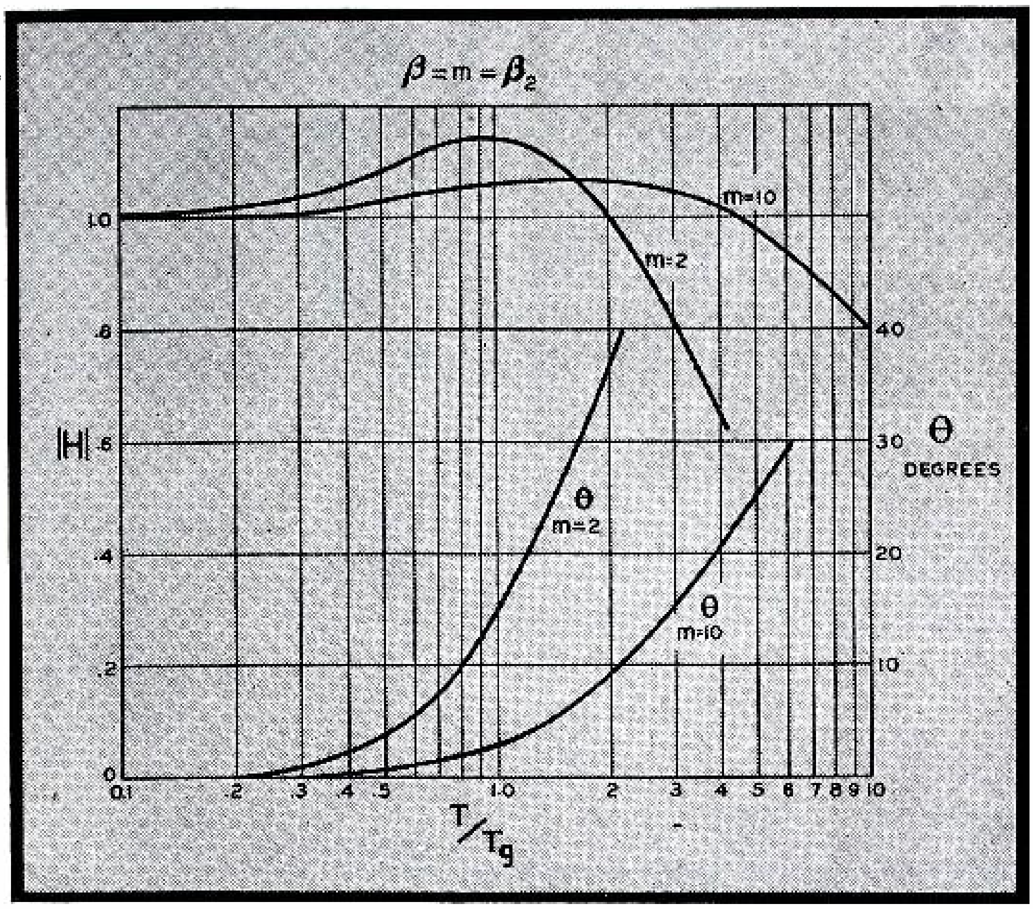

Figure 8
The phase and amplitude response curves when $\beta = \beta_2$.

characteristic, we note that when $T/T_g = 0$, $|H| = 1$, the relative gain in the mid-frequency region. What are the conditions necessary for $|H|$ to remain constant with frequency? Equation 18 is in the form of a quotient of two algebraic polynominals in ascending even powers of the variable, and with constant terms equal to unity. If the coefficients of like powers of T/T_g in the numerator and denominator are made equal, it is seen that the expression becomes a constant equal to unity for *all* values of the variable. Mathematically, this means that the derivative of $|H|^2$ with respect to $(T/T_g)^2$, i.e. the slope of the curve of the function, is a constant equal to zero.

There is no difficulty in equating the coefficients of $(T/T_g)^2$, but what of the fourth-power term in the denominator? To meet the condition for constant gain, this term must vanish which means that $1/\beta^2$ must equal zero, or β must be infinite. Hence the ideal flat response cannot be achieved with this type of compensation except in the practically impossible case where T_F is infinite. We can, however, approach the fulfillment of the conditions by making β as large as possible so that $1/\beta^2$ approaches zero.

Since we are able to equate only the coefficients of $(T/T_g)^2$ the derivative vanishes only at the origin, and the response curve does not remain

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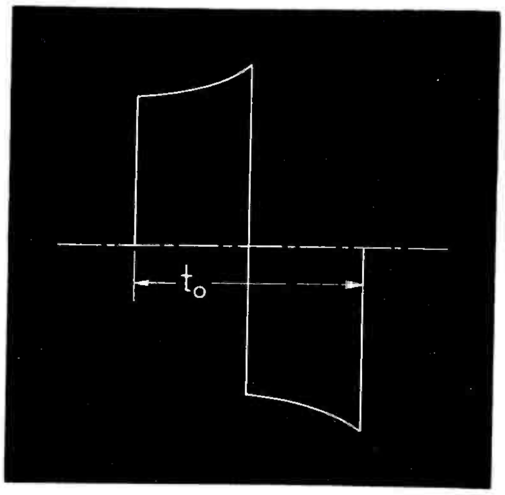
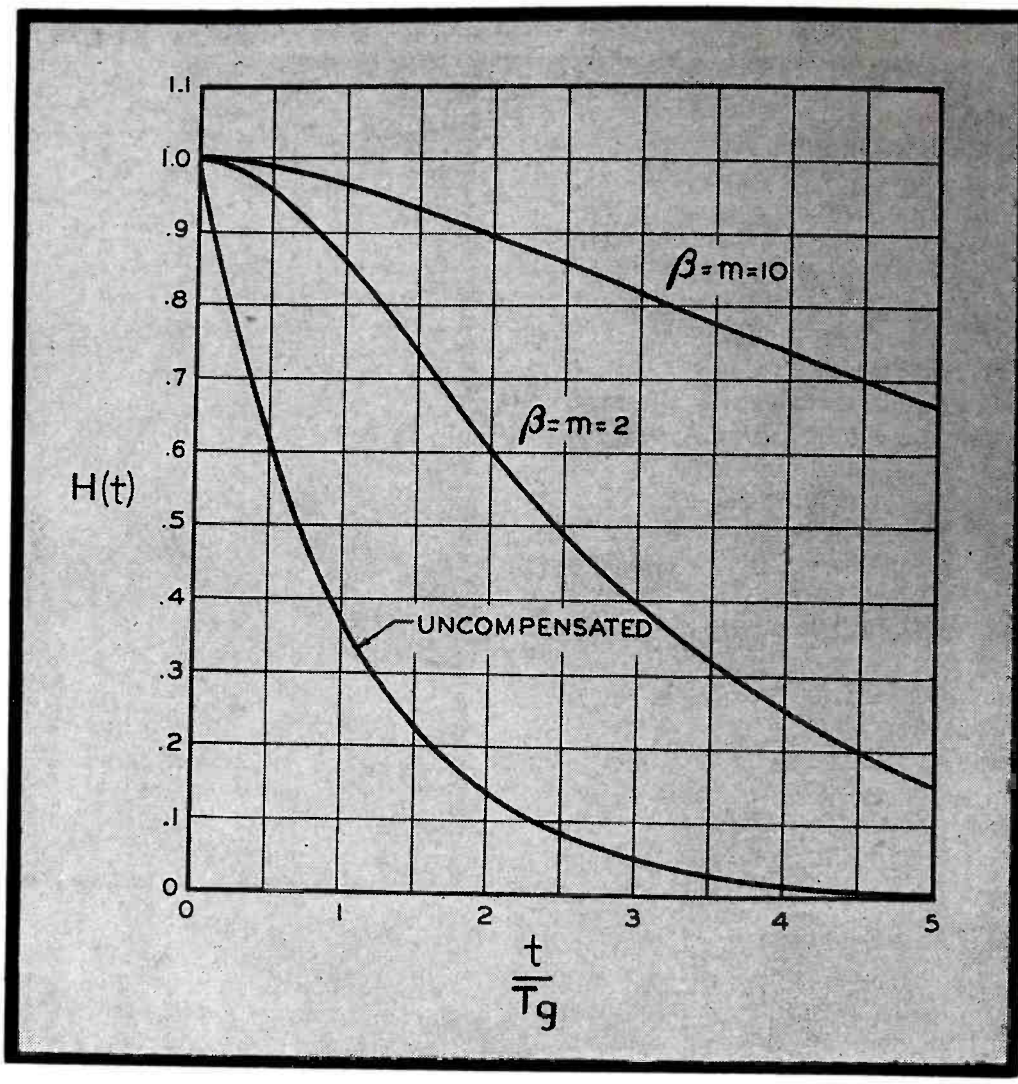
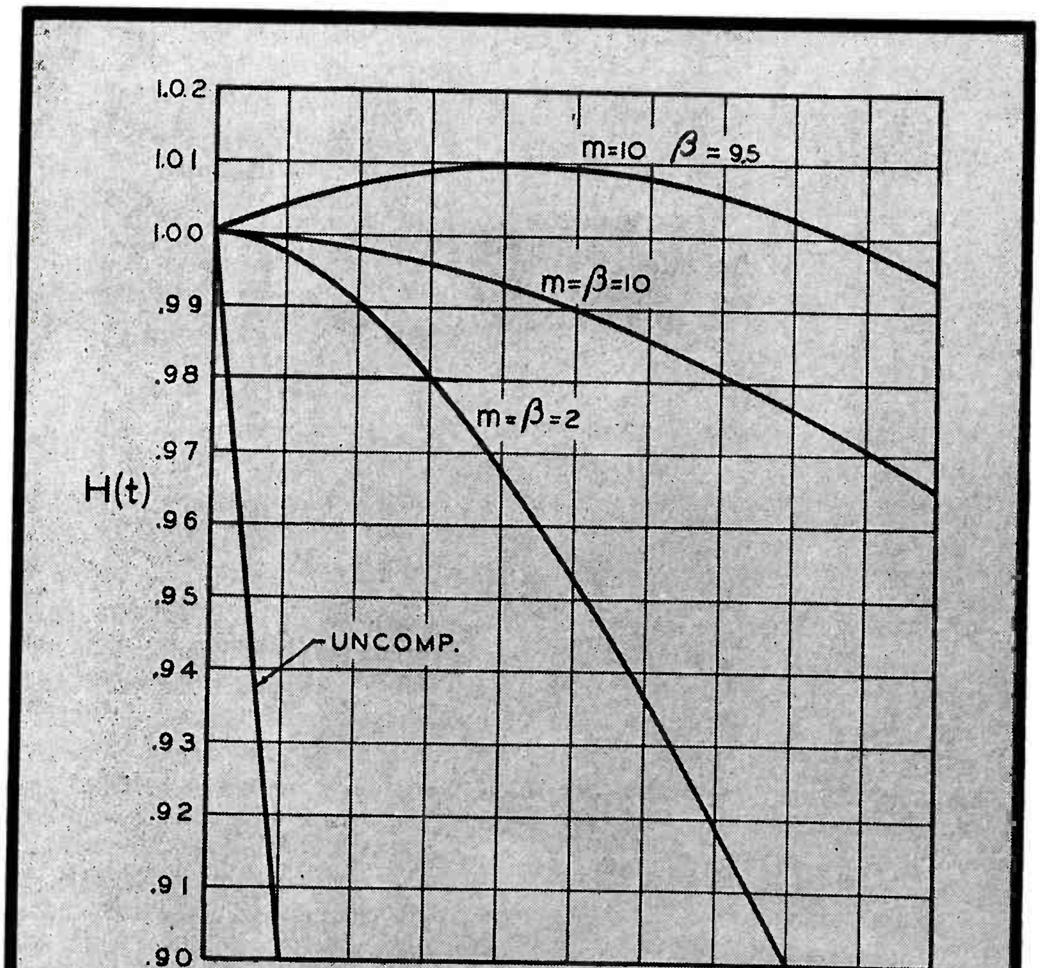




Figures 9 (top), 10 (right) and 11 (below)

Figure 9 illustrates the general appearance of the load voltage e_L across the load resistor and the plate circuit filter with a square wave input. Figure 10 shows the transient response of the compensated amplifier for the two values of *m* compared with the uncompensated response. In Figure 11, we see an enlargement of the transient response curves for the duration of one grid-time constant. The upper curve shows the response when β is made slightly smaller than β_2.



flat as T/T_g is increased, but instead is characterized by the behavior of the fourth-power term in the denominator. For fractional values of the variable, $(T/T_g)^4$ will be a very small quantity, so that its effect will be slight up to the point where $T/T_g = 1$ (the nominal cutoff). Beyond this point, as T/T_g takes on integer values, the denominator will increase in magnitude very rapidly, so that ultimately, $|H|$ will approach zero.

Before evaluating the constants, some thought should be given to practical limitations. The ratio, *m*, cntrols the size of the resistance R_F, which is limited because of the resulting d-c voltage loss across it. This, in turn, depends upon the size of *R* and upon the available plate-supply voltage, if a sufficient operating voltage is to be maintained at the plate of the tube. We have determined that the low-frequency response is improved if β is made large, and knowing that if we equate coefficients the constants will be interdependent, we conclude that *m* should be set at its maximum practical value. This is desirable also from the point of view of providing adequate decoupling.

It is convenient, therefore, to consider *m* as fixed by practical considerations, and to determine the proper value of β to achieve a specific

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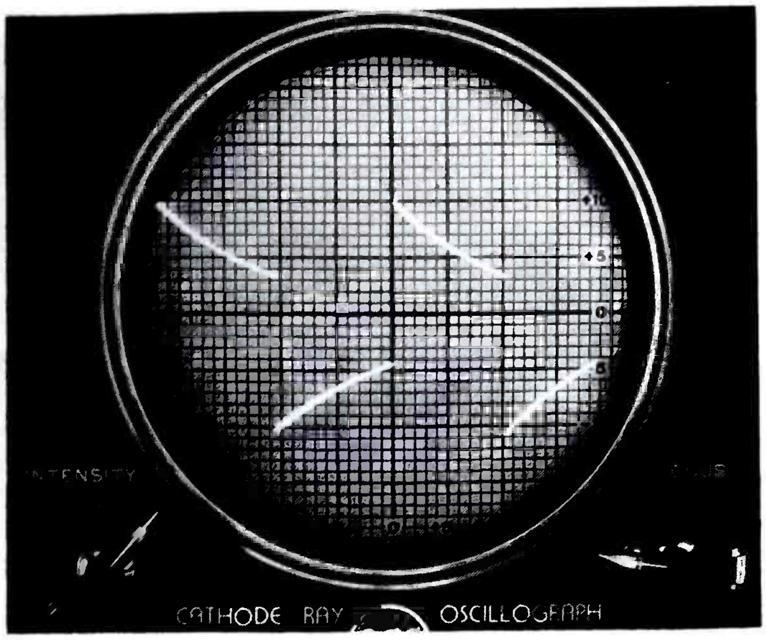


Figure 12
Square wave response of the uncompensated amplifier
The period is the same for all photos, ie.,

$$\frac{t_0}{2} = T_g \text{ or } \frac{T}{T_g} = \frac{2}{\pi} = 0.637$$

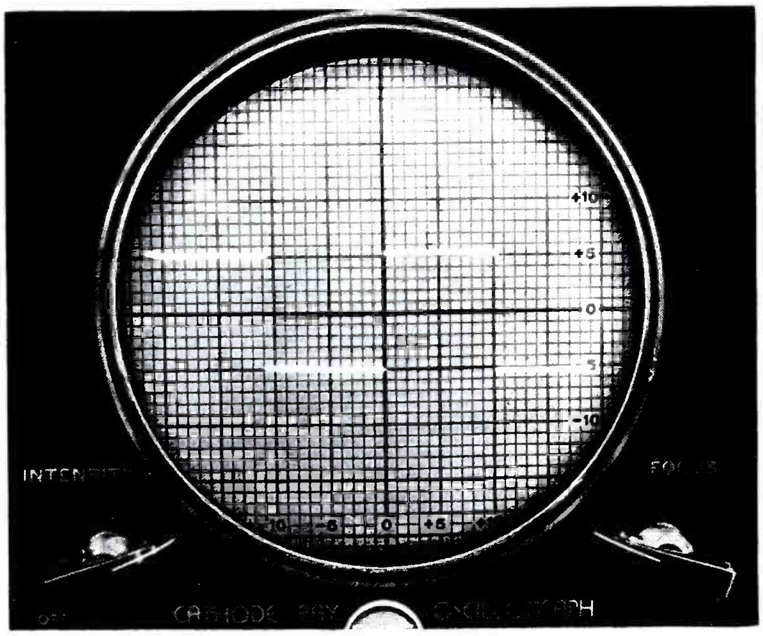


Figure 13
Optimum correction o fthe square wave response for the value $m = 10$. $\frac{\beta}{m} = 0.9$.

response characteristic. Returning to Equation 18, and equating the coefficients of $(T/T_g)^2$, it is found that

$$\beta = \sqrt{m\,(m+2)} = \beta_1 \qquad (20)$$

This is a *critical* value of β yielding the flattest amplitude response which does not exhibit a peak beyond the origin. If β is made larger than β_1, the response will be poorer, while if β is smaller, the response curve will peak at some value of T/T_g greater than zero, and instead of being flat near the origin, it will start out with a rising characteristic. While β_1 is a unique value, it is not necessarily the optimum value. It is reasonable to suppose that the response curve may be made to fall within a given plus

(*Continued on page* 50)

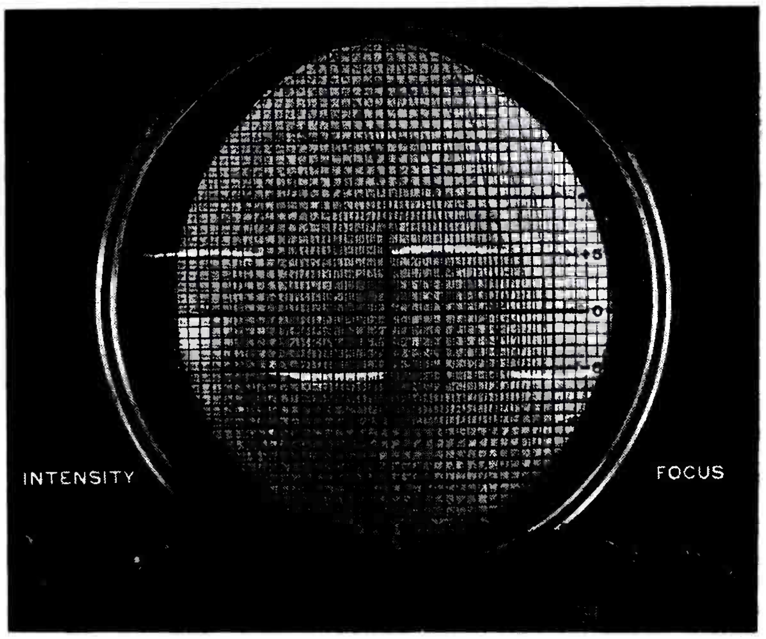


Figure 14
With m reduced to 2, it is no longer possible to achieve a flat wave top, since the frequency is too low for so small a value of m. Here $\frac{\beta}{m} = 0.82$.

Figure 15
The square wave response when $m = 2$ and $\beta = (m + 1)$. Here C_F is too large.

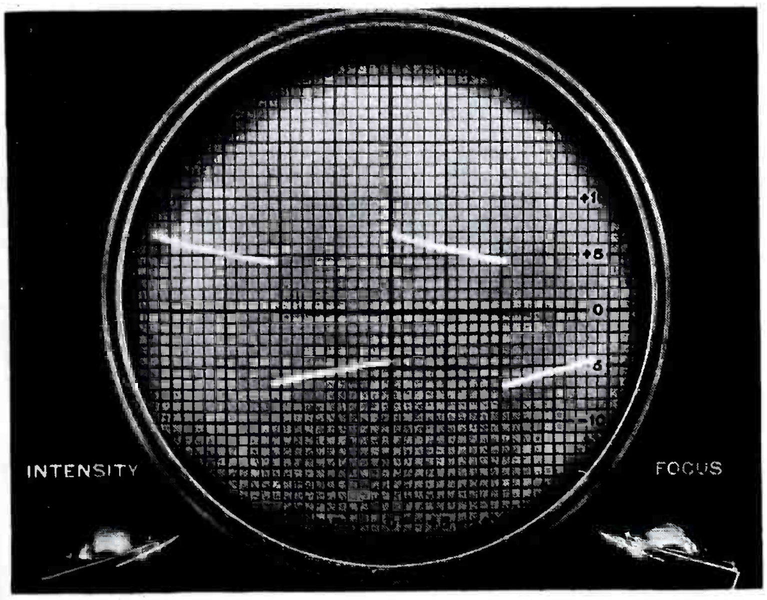
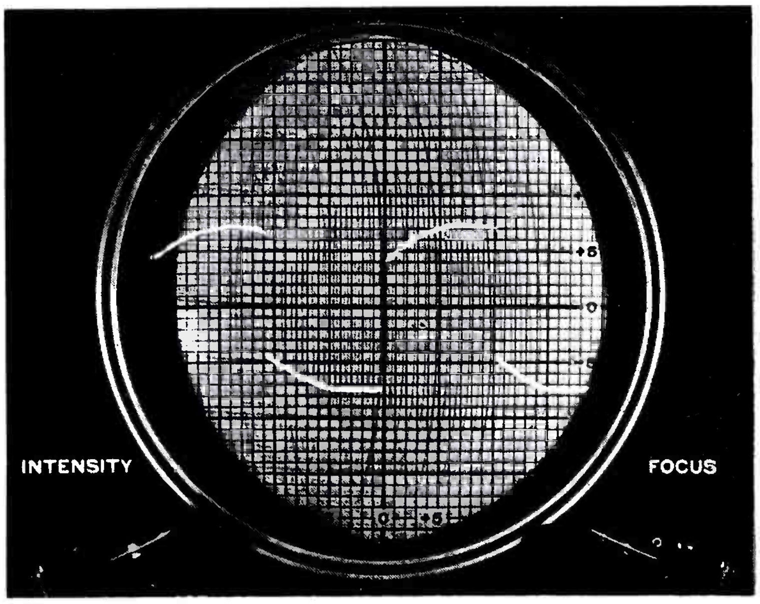


Figure 16
The effect of overcompensation caused by C_F being too small. Here $m = 2$, $\frac{\beta}{m} = 0.47$.



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A HIGH-VOLTAGE

Using Mercury Vapor Tubes

by SCOTT HELT

Chief Engineer, WIS

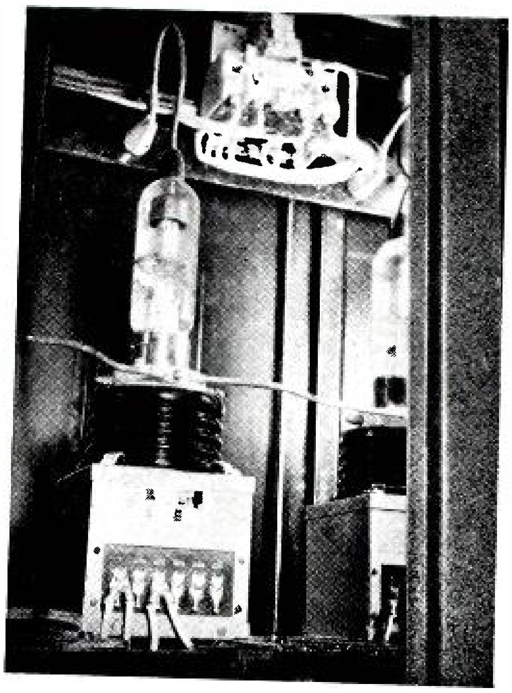

Figure 1

Closeup of tube section of the high voltage supply. Note the short point-to-point connections.

IT has been the trend for some time to employ electronic rectifying devices for the production of high voltage plate power to the vacuum tubes of transmitting equipment. Motor-generator sets are considered outmoded for the production of high-voltage direct current plate power. And the elimination of rotating equipment affords not only more economical and efficient operation, but also eliminates the maintenance problem which is of great consequence when motor-generator sets are employed. When rotating equipment fails, there is usually too much time lost off the air if auxiliary power supply equipment is not maintained, and the servicing of high-voltage armatures is an expensive and time consuming undertaking.

On the contrary, electronic rectifiers provide quiet, clean operation, and are quickly started and stopped. The replacement of exhausted tubes is ordinarily the only maintenance required. Too, it is a relatively simple matter to filter out the noise component present in the output of the rectifier through the use of a carefully designed multi-section, low-pass, "L" type filter.

Since the noise component at the output of a direct current generator is not always at the same amplitude, it is difficult to maintain the ripple at such a low level that it does not appear as an appreciable modulation component in the carrier of the transmitter, with which the power supply is associated. High-voltage rotating equipment must be continuously serviced if any predetermined noise level to be tolerated at the output of the d-c generator is to be maintained over an appreciable length of time.

A high-voltage plate supply rectifier unit of the electronic type was designed and constructed at this station. This power supply is giving continued and satisfactory performance under the rigorous conditions imposed in the broadcast service.

Preliminary Considerations

There are four preliminary design considerations. First, the load voltage requirement is 1,600 volts, d-c and the total load current to be supplied is .768 ampere. Thus the rectifier will be designed to supply .8 ampere. This includes 32 ma bleeder I. Second, a three-phase, 220 volt, 60 cycle, a-c supply voltage is available, and it is desired to use mercury vapor tubes. The low voltage drop of mercury vapor tubes will result in higher operating efficiency than would result if high-vacuum tubes were employed. Since the voltage drop is essentially constant throughout the working range of current, the voltage regulation will be superior to that obtained with high vacuum tubes. In addition, the use of three RCA type 872-A mercury vapor tubes has been found to meet the load requirement with generous safety factor. This tube will actually deliver a maximum load current of 2.1 amperes at 3,500 volts, with a maximum inverse peak voltage of 10,000 volts, and without exceeding the rated tube limits.

In the third step, we consider the three phase interconnected *WYE* (zig-zag) rectifier circuit, selected instead of a single phase circuit. Polyphase rectifiers develop an output wave that is much closer to a steady d-c potential. Also a higher output voltage in proportion to peak inverse voltage is possible. The circuit chosen utilizes the possibilities of the transformers more effectively, and by interconnecting the secondaries, d-c saturation of the cores is prevented. When the current is allowed to flow in only one direction through the secondaries, the resultant d-c component of flux saturates the cores. This reduces the efficiency by increasing the magnetizing current, the hysteresis loss, and introducing objectionable harmonics in the secondary emf. Interconnecting the secondaries in the manner shown requires that each half of the plate transformer secondary voltage rating be increased approximately 15% over that calculated for a straight delta-wye three phase half wave connection. This is to take care of the 120-degree difference between phases.

The fourth consideration involves the permissible percentage of ripple. This was found to be .089%, and is based on results obtained in another rectifier constructed by the writer for use with high fidelity broadcast equipment. This will permit a noise level at the rectifier filter output considerably lower than the present average found in high quality broadcast equipment. The present percentage of ripple tolerated in commercial equipment is found to be .25 to 1.0% in some cases,[1] but the designer is looking toward the future, when mandatory higher quality performance by the licensing authority will necessitate reduced noise level in the transmitted signal.

The ripple frequency will be 180 cps for the three phase half wave circuit, as compared with 120 cycles for the usually employed single phase full wave circuit in rectifiers, designed to

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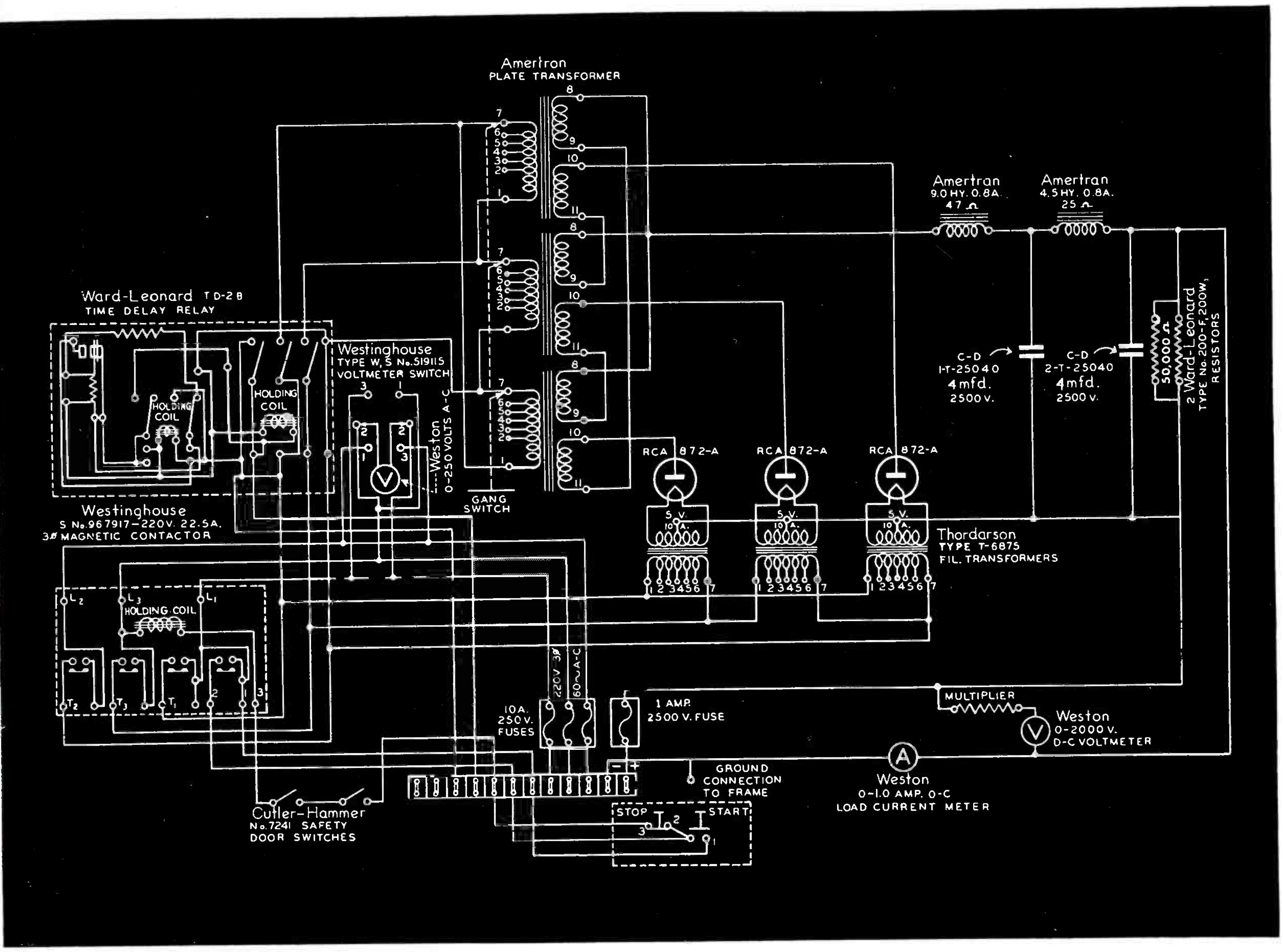
PLATE POWER SUPPLY



meet the present power requirement. The ripple frequency increases and the ripple amplitude decreases as the number of phases employed is increased. Thus, proper filtering in this case will resolve into a simple problem, and the actual filter components will not prove expensive, nor of great physical size.

A schematic diagram of the basic circuit is shown in Figure 4. A table of circuit characteristics is shown in Figure 3.

The Design

A two section "L" type, low pass, inductance-capacity filter is used, since multi-section filters of this type result in low peak anode tube current, high tube and transformer efficiency, and good voltage regulation. Assuming that the inductive reactance of the choke in each section is large in comparison with the reactance of the shunt capacitance, the required *LC*

•

Figure 2

The three-phase, half-wave, interconnected *wye* (zig-zag) high voltage plate supply rectifier, rated at 1600 volts d-c, .768 ampere, at the load.

•

product for each section of a filter having *n* sections of the *L* type is approximately . . .

$$LC = 0.0253 \frac{\sqrt[n]{\alpha_1} + 1}{f_r^2} \qquad (1)$$

When the d-c resistance of the choke is small as compared with the inductive reactance of the choke, the smoothing factor is approximately

$$\alpha = \frac{X_L - X_c}{X_c} = \omega^2 L\,C\text{-}1 \qquad (2)$$

Assuming the filter reactors L_1 and L_2 are to be so designed as to have a d-c resistance of approximately 50 ohms for the first reactor, and 25 ohms for the second reactor, E_{ave} input to the filter will be 1,600 volts plus (75 ohms × .8 A) or 1,660 volts. This allows for the *IR* drops through the filter reactors, and will require that E_T (per ½ secondary at the plate transformers) will be 1,660/2.03 or 817 volts. Actually, the transformer manufacturer can be allowed to supply transformers rated 835 volts per one-half secondary, to allow for the ten-volt drop through the 872-A mercury vapor tube, as indicated by the tube manufacturer, and for transformer losses which cannot be well anticipated in advance design. The plate transformer primaries can be tapped plus or minus the average rms line voltage to provide for d-c output voltage adjustment.

From the above calculations and the table (Figure 3) it can be seen that

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the inverse peak voltage per tube will be 1,660 x 2.09 E_{d-c} or 3,469.4 volts. This is well within the maximum safe operating limits of the 872-A tube, which has a maximum inverse peak voltage rating of 10,000 volts. Long life can therefore be anticipated from the tubes, and no tube should "flash-back" when the plates go negative on the negative alternation of the sine wave—even though the rectifier were operated at twice the rated capacity. This is generous safety factor.

Therefore, the plate transformers, three in number, can be specified as capable of developing 835 volts, rms per one-half secondary, with primaries designed for 220 volts, 60 cycles, a-c, and tapped at 210, 215, 220, 225, 230 and 235 volts.

After assembly, the primaries should be tested at 2,500 volts, rms, at 60 cycles, for breakdown, and the secondaries tested at 7,500 volts, rms, 60 cycles. The primary and secondary connections and taps must be brought out to ceramic insulated terminals for ease of connection to the external circuit. The plate transformers must be air insulated, self-cooled, fully encased, and constructed for 55-degrees centigrade maximum temperature rise. The windings must be electrostatically shielded, coils vacuum varnish impregnated, all mountings compound filled, and all the assemblies moisture proofed.

The Filter Design

The filter is designed as to draw continuous tube current and to so attenuate the a-c ripple component that it appears as only .089% at the output. If the inductance of the first reactor is made sufficiently large, the load current will become essentially constant, and the current in each phase of the circuit will resemble a square topped pulse. It is necessary to calculate the minimum amount of *L* which will permit this condition.

The a-c component for this type of rectifier at the filter input is .180 E_{d-c}. From the table of characteristics for this type of rectifier we see that E_{d-c} is equal to 2.03 E_Y; E_Y having been specified as 835 volts. Thus, E_{d-c} equals 1,695 v. and the a-c component at the filter input equals .180 × 1,695 v or 305.1 volts. The a-c ripple component at the d-c load will be .089% or 1,600 × .00089 or 1.43 volts at 180 cycles, the fundamental ripple frequency.

The filter attenuation factor will be . . .

$$\frac{\text{a-c voltage across load}}{\text{a-c voltage across filter input}} \text{ or } \frac{1.43 \text{ v}}{305.1 \text{ v}} \text{ or } .0047 \quad (3)$$

The minimum amount of inductance it is permissible to use for this attenuation factor at L_1 is such that $\frac{2\pi F L_1}{R_{load}}$ must not be less than .25, where . . .

$$R_L = \frac{E_L}{I_L} + R_{d-c} \text{ of } L_1, L_2.$$

$X_L = R_{load}/4$, and $R_{load} = [(1600/.8) + (.8 \times 75)] = 2{,}060$ and min. $X_L = \frac{2060}{4} = 515$ ohms. $L_1 = 515/2\pi F = 515/6.28 \times 180 = .49$ H.

In practice a considerably larger inductance will be specified. This is so

(*Continued on page 26*)

E_{d-c}	2.03 E_Y 1.17 E_{rms} 0.827 E_{max}
I_{d-c}	1.73 I_{rms}
E_{rms}	0.854 E_{d-c} E_{rms} equals
E_{max}	1.21 E_{d-c}
$E_{inverse}$	2.09 E_{d-c}
Secondary I_{rms}	0.577 E_{d-c}
Secondary kva in terms of d-c load	1.71 Kva
Primary kva in terms of d-c load	1.21 Kva
Trans. av. kva in terms of d-c load	1.46 Kva
Max. rms input	41% MPIV
Max. d-c output	82.7% MPPC
Ripple frequency. F equals frequency of power supply	3 F
RMS ripple volts	0.180 E_{d-c}

Figure 3

Above data assumes sine wave supply, balanced phase voltage, zero tube drop, pure resistance load, and no filter used.

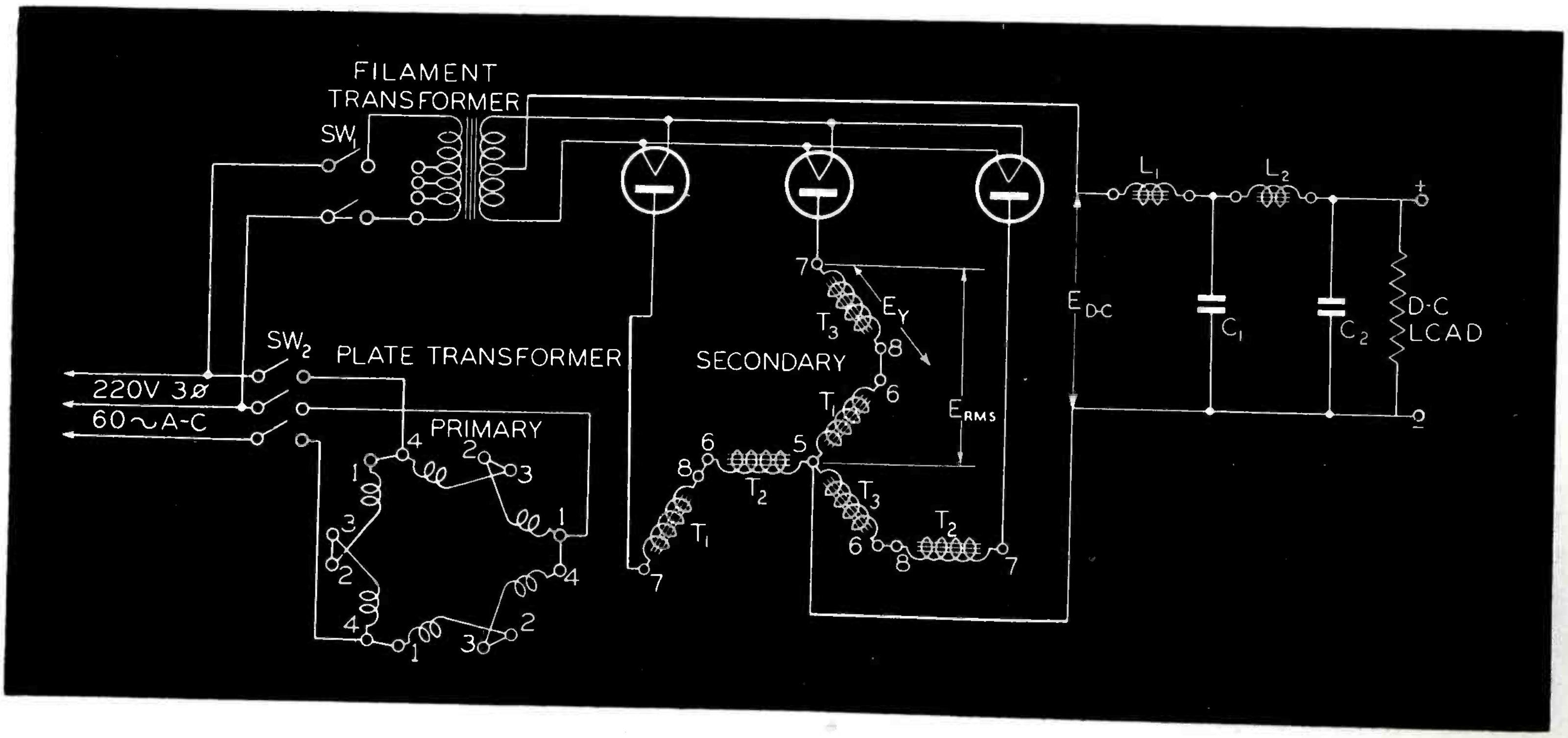

Figure 4

The basic diagram for the three-phase, half-wave, interconnected *wye* (zig-zag) rectifier circuit.

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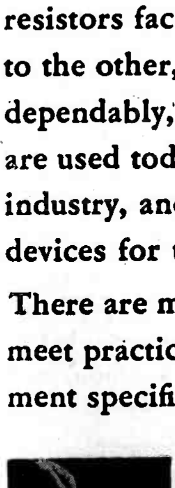

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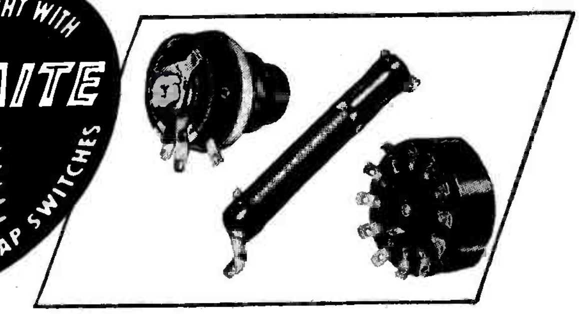

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the reactance of the first filter choke will be large as compared with the reactance of the first filter condenser. ($X_{L1} > X_{C1}$). This is so the rectifier tube current will be rectangular, in order that current will flow during the fullest possible operating period of each tube. This will materially increase tube life.

The actual value of *L* and *C* for each section of the filter, to keep the ripple within the required limit of .089% at the output, are calculated as follows . . .

$$.0047 = \frac{1}{(2\pi F)^4 \cdot L_1 L_2 C_1 C_2} \qquad (4)$$

Since the equation can be solved only by the product $L_1 L_2 C_1 C_2$ it is necessary to select arbitrarily either the desired value of *L* or *C* and solve for the other. As filter condensers of the high voltage type may be secured in standard values, it is always expedient to select arbitrary values of *C* for the two sections of the filter. The values of *L* will be calculated.

To insure good regulation, even though the load current should fluctuate over wide limits, the second filter condenser should be fairly large—to insure sufficient reservoir capacity. Let us choose 8 mfds for this section, and 4 mfds for the first section.

Rearranging the equation to determine the value of *L* required in each filter section:

$$.0047 = \frac{1}{(2\pi F)^4 \times L_1 L_2 C_1 C_2}$$

$$\text{(and)} \quad L_1 L_2 = \frac{1}{.0047 \times 163 \times 10^9 \times 32 \times 10^{-12}}$$

$$= \frac{1}{47 \times 10^{-4} \times 163 \times 10^9 \times 32 \times 10^{-12}}$$

$$= \frac{1}{47 \times 163 \times 32 \times 10^{-7}}$$

$$= \frac{10^7}{245152}$$

$$= 40.79 \text{ h}$$

In order that the two sections of the filter may have the same cut-off frequency $F_{C_1} = F_{C_2}$ and $L_1 C_1$ must equal $L_2 C_2$. This is because the cut-off frequency is inversely proportional to the square root of the product of *L* and *C* in each section. Since 4 mfds of capacity have been chosen for the first section, and 8 mfds for the second section, it follows that *L* in the first section must double that of the second

Figure 5

The completed power supply in case designed to match the transmitting equipment.

section. We also see by ratio and proportion that if we use 9 h in the first filter section, and 4.5 h in the second filter section, that this product of 40.5 h is sufficiently close to the calculated product of 40.79 h for $L_1 L_2$ to fulfill the design requirement for all practical purpose. Thus, *L* in the first filter section will be made 9 h and the second reactor will have an inductance of 4.5 h at .8 ampere.

Therefore . . .

$$L_1 C_1 = 9 \times 4 \times 10^{-6} = 36.0 \times 10^{-6}$$

$$L_2 C_2 = 4.5 \times 8 \times 10^{-6} = 36.0 \times 10^{-6}$$

It is seen that the cut-off frequency in both sections will be identical. This will fulfill the attenuation factor of .0047, closely.

The cut-off frequency in the low pass filter will be:

$$F_C = \frac{1}{\pi\sqrt{LC}} \qquad (5)$$

$$= \frac{1}{3.14\sqrt{36 \times 10^{-6}}} = \frac{10^3}{113.04}$$

$$= 8.8 \text{ cps}$$

It is seen that the cut-off frequency of 8.8 cycles is well below the ripple frequency of 180 cycles, and excellent attenuation of the hum component is to be realized. In fact the a-c component in the output will be 1.43 volts as compared to 1,600 volts d-c.

The attenuation of the a-c component through the filter in db will be

$$\text{db} = 20 \text{ Log}_{10} \frac{E_1}{E_2}$$

$$= 20 \text{ Log}_{10} \frac{305.1}{1.43}$$

$$= 20 \times 2.32838$$

$$= 46.5 \text{ db}$$

Construction of Unit

In the construction of the rectifier, the components were housed in a suitable metal cabinet, provided with interlocks at the doors so that the operating personnel is protected against coming into accidental contact with the high voltage. These interlocks cause the main power relay to drop out when either front door is opened.

A time-delay relay of conventional design is provided so that the filaments of the rectifier tubes may be heated fully before the plate voltage is applied. An auxiliary set of contacts has been provided on the relay, so that in event of a momentary power failure at the transmitter, not of sufficient duration for the filaments of the mercury vapor tubes to cool off, this circuit may be closed by means of a single circuit momentary contact push-button to eliminate the time delay. The starting and stopping of the rectifier, too, is controlled by means of a momentary contact push-button station mounted on the transmitter proper.

A three-gang, rotary six-pole per section switch is mounted on the control panel of the rectifier, to which the taps from plate transformer primaries are brought out. This allows convenient control of the d-c output voltage by means of regulation of the primaries. A rotary voltmeter switch is also mounted on the rectifier control panel to provide a convenient means of measuring the line voltage at each phase. It will be noted that the filament transformers are so connected that each is operating across a separate phase, so as to balance the load. Indicating instruments are provided to meter line voltage, load voltage, and load current.

In Figures 1 and 5, appear views of the supply, showing the general type of construction used. This rectifier is associated with a Western Electric five-kilowatt transmitter, to provide plate supply to the speech amplifier, low-level modulator (both operating

(*Continued on page 72*)

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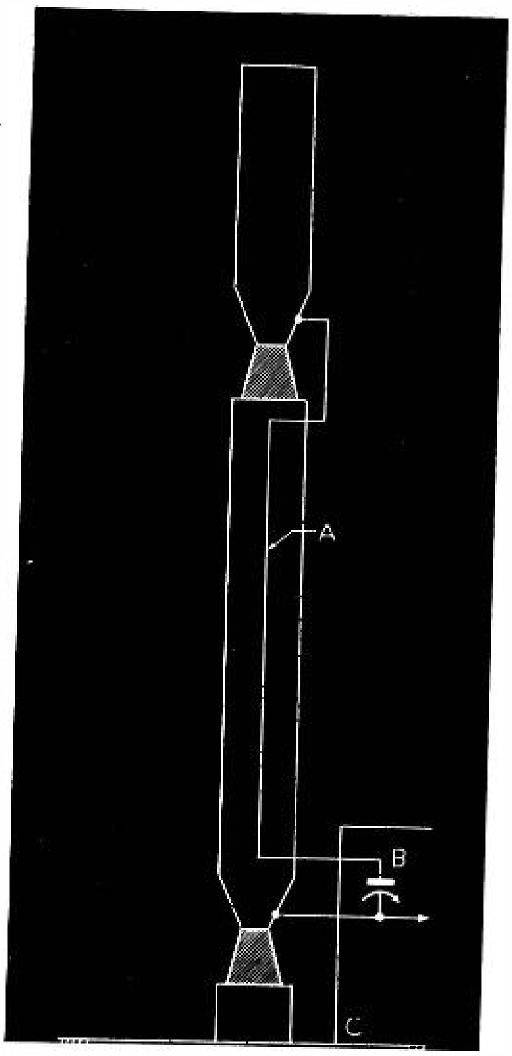

TRANSMISSION LINES AS REACTORS

In Antenna Construction

by DR. VICTOR J. ANDREW
Victor J. Andrew Company

Figure 1
A conventional split tower, originally intended for inductive loading at the upper insulator. Use has been made here of the lower portion of the tower as the outside conductor of a concentric feed system, by installing the antenna feed line A. The addition of the tuning capacitance B creates the desired inductive reactance at the point of coupling. C is the top of the antenna house.

THE theoretical gain in radiation efficiency obtained by inductive loading of a small antenna at a point half-way or more than half-way from the ground to the top, has been recognized for several years. A number of such installations have been made. However, there are two very serious drawbacks to such constructions.

In the first instance, the power losses in this inductance offset all the gain, unless the inductance is of extremely high Q. The second complication involves the mechanical difficulties of constructing a small, light weight, weatherproof inductance of high Q, and capable of handling large voltages and currents. In addition there are the difficulties of tuning it after installation.

When the Columbia Broadcasting System proposed such inductive loading at a point 330 feet above ground in a 500-foot vertical radiator at WBBM some time ago, it became apparent that a better mechanical way of accomplishing the purpose was badly needed. The following method was accordingly developed and installed, and was found so successful that a similar installation was made at WABC.

The tower was broken by a sectionalizing insulator at the 330-foot level, with the expectation of installing an inductance connected across this insulator. In the novel solution developed, instead of connecting an inductance across this insulator, the two conductors of a transmission line were connected across this insulator. The other end of the transmission line was brought to ground. Across the end of the transmission line on the ground, a reactance (inductance or condenser) can be connected. With any length of transmission line and any desired reactance at the upper end of the line, a required value of reactance across the bottom of the line may be found (either by calculation or by experiment).

In this installation, the length of the transmission line was close to a quarter-wavelength. It was therefore apparent that a capacitive reactance was needed across the bottom of the line to produce an inductive reactance across the top of the line. The line also gives a much higher Q than can be obtained with a coil.

The construction of this line from the sectionalizing insulator to ground is much simpler than it at first appears. Since one side of the line is connected to the tower below the insulator, it is possible to use the tower itself as one of the conductors. The other conductor was suspended inside the tower, and insulated from the tower, Figure 1. By making the insulated conductor of 1⅜" copper tube, it was possible to carry inside the tube the wires supplying power to the lights on the top of the tower. Other possible circuits which may be carried in this tube are u-h-f transmission lines to auxiliary antennas on top of the tower, sampling lines for measuring magnitude or phase of current at points near the top of the tower, or power coaxial cables if it is desired to couple part of the transmitter power directly into the top section of the antenna.

There are other interesting theoretical possibilites of this method, which it is hoped can be given more thorough investigation when engineering development can again be devoted to broadcast station antennas. Let us, however, now take one of these theoretical cases, and assume that an engineer wishes to make some adjustments of the antenna tuning equipment. We find that inside the antenna tuning house, he is looking at a three terminal load, the two insulated leads and the ground return. One condenser has been provided, which will when tuned, change both the ratio of currents in the two leads and the phase difference between these currents. (Proper indicating instruments may readily be provided for indicating currents and phase difference.) It is most unlikely that the *accidental* combination of phase and current ratio obtained with this one condenser will be the one which will produce optimum results. It is possible, however, to use a network of two or more tuning elements to feed these two branches of the circuit, to obtain any desired combination of phase and current ratio in these loads. There are interesting possibilities of either a mathematical or experimental solution of this problem to obtain *optimum antenna performance.* The broad term *optimum performance* is used, since in broadcast antennas this may mean either maximum ground wave field intensity, or maximum fading-free area. In a high powered station, the former is wanted in daytime, and the latter at night. In the construction discussed, where all tuning controls are in the antenna tuning house, the antenna can readily be switched by relays to give different performance day and night.

(*Continued on page* 50)

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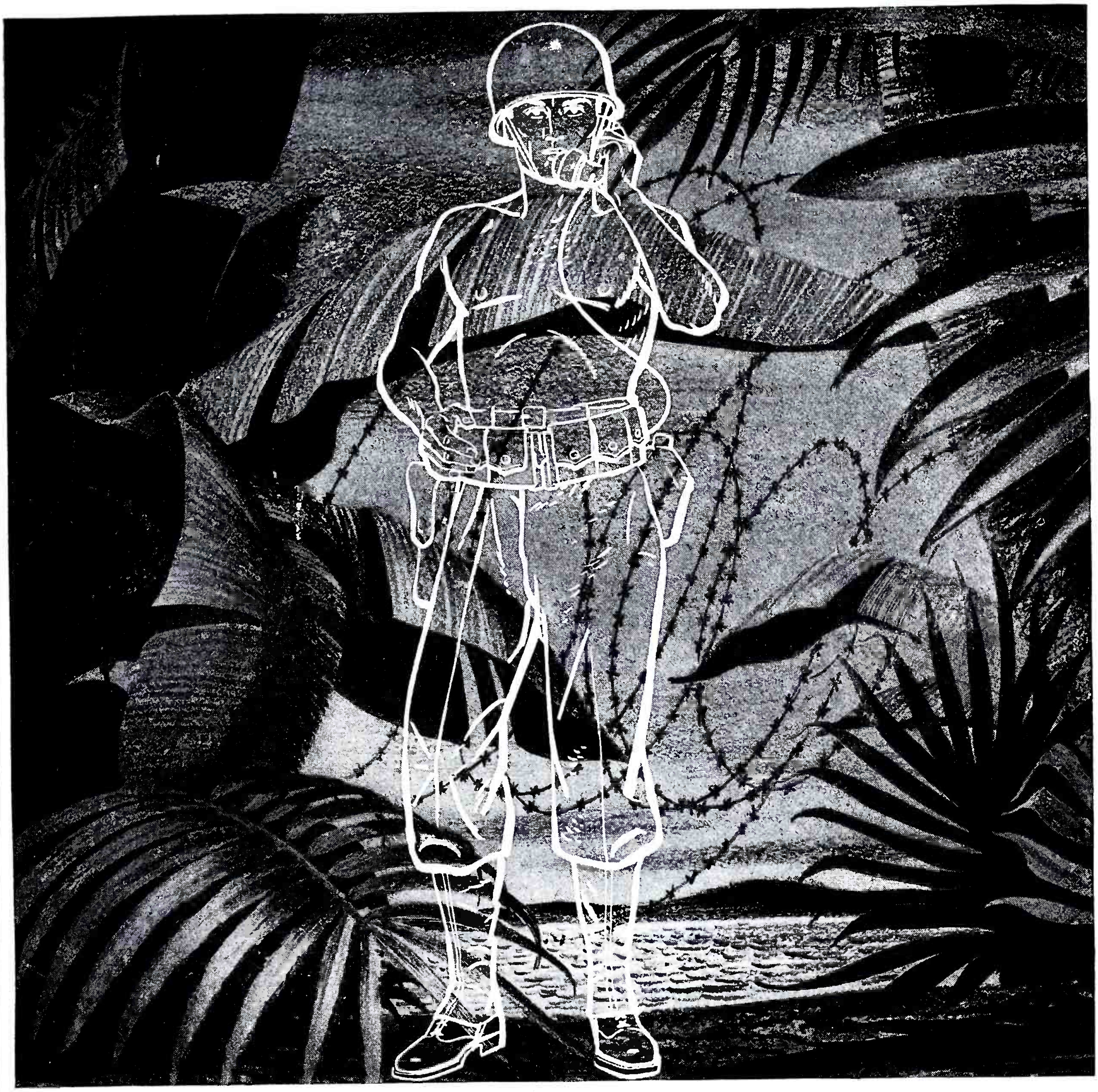

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CLOSED-FORM STEADY-STATE

For Periodic Applied Voltages

by SIDNEY FRANKEL

Radio Engineer, Federal Telephone & Radio Corp.

THIS paper covers a discussion of the application of a simple but very useful theorem representing an extension of Laplace-transform methods in which the solution for the response of an electrical system is written as a Bromwich integral by virtue of the Mellin inversion theorem.[1] To the best of the writer's knowledge, the extension used here was first presented by Professor A. Hazeltine of the Stevens Institute of Technology. He described the method in a lecture course on operational methods, given at the Institute during 1939-1940.

To cite a simple example of the type of problem involved, suppose it is desired to find, in closed form, the periodic current in an inductor of inductance L, resistance R, when a rectangular wave-form voltage (similar to pair 6, Table 1, Figure 1) defined by

$$e = E = constant,\ n\tau < t < n\tau + \tau/2$$
$$= o \qquad ,\ n\tau + \tau/2 < t < (n+1)\,\tau \qquad (1)$$

is applied to the inductor, where τ is the period of the wave and n is any integer. The classical method of solution of such a problem is well understood[2,3]; but in view of the modern methods available, the classical method seems somewhat lacking in compactness of form and thought.

The Laplace Transform

The method of procedure is much improved without loss of rigor by the use of the Laplace-transform and associated tables of integrals[1,4,6]. Briefly outlined the method is as follows: Given a function of time $f(t)$, such that $f(t) = o$ for $t<o$, and let $p = x + jy$, where $j = \sqrt{-1}$, y is real, and x is real and greater than zero. If x can be chosen so that the integral

$$\bar{f}(p) = \int_0^\infty f(t)\,\epsilon^{-pt}\,dt \qquad (2)$$

exists, then $\bar{f}(p)$ is called the Laplacian transform of $f(t)$.

For lumped-constant network analysis we need two other fundamental transforms:

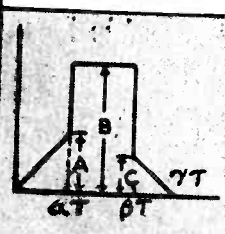
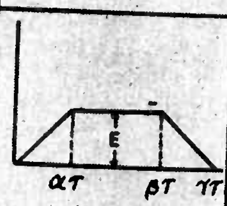
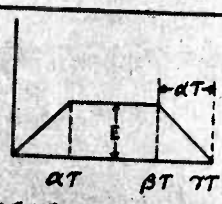
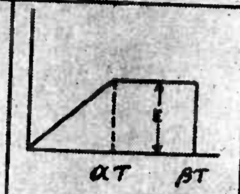
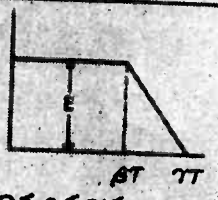
$$\overline{\left(\frac{df}{dt}\right)} = \int_0^\infty \frac{df}{dt}\,\epsilon^{-pt}\,dt = -f(o) + p\bar{f}(p) \qquad (3a)$$

$$\overline{\left(\int f dt\right)} = \int_0^\infty \left[\int f dt\right]\epsilon^{-pt}\,dt = -\frac{1}{p}\left[\int f dt\right]_{t=o} + \frac{\bar{f}(p)}{p} \qquad (3b)$$

If now, we identify $f(t)$ with some current, $i(t)$, in a network, then

Figure 1 (right)

TABLE 1 PERIOD TRANSFORMS

PAIR NO.	SINGLE PERIOD WAVE SHAPE	TIME FUNCTION $E_\tau(t)$	TRANSFORM $\bar{E}_\tau(p)$
1	 $0 \leq \alpha \leq \beta \leq \gamma \leq 1$	$A\frac{t}{\alpha\tau}$, $0 \leq t < \alpha\tau$ B, $\alpha\tau < t < \beta\tau$ $\frac{C}{\gamma-\beta}\left(\gamma-\frac{t}{\tau}\right)$, $\beta\tau < t \leq \gamma\tau \leq 1$	$A\left\{\frac{1-(1+\alpha\tau p)\epsilon^{-\alpha\tau p}}{\alpha\tau p^2}\right\}$ $+\frac{B}{p}\left\{\epsilon^{-\alpha\tau p}-\epsilon^{-\beta\tau p}\right\}$ $+\frac{C}{(\gamma-\beta)\tau p^2}\left\{\left[(\gamma-\beta)p\tau-1\right]\epsilon^{-\beta\tau p}+\epsilon^{-\gamma\tau p}\right\}$
2	 $0 \leq \alpha \leq \beta \leq \gamma \leq 1$	PAIR (1) WITH A=B=C=E	$\frac{E}{\alpha(\gamma-\beta)\tau p^2}\left\{(\gamma-\beta)-(\gamma-\beta)\epsilon^{-\alpha\tau p}\right.$ $\left.-\alpha\epsilon^{-\beta\tau p}+\alpha\epsilon^{-\gamma\tau p}\right\}$
3	 $0 \leq \alpha \leq \beta < \gamma-\beta+\alpha \leq 1$	PAIR (2) WITH $\gamma-\beta=\alpha$ (EQUAL SLOPES)	$\frac{E}{\alpha\tau p^2}\left(1-\epsilon^{-\alpha\tau p}\right)\left(1-\epsilon^{-\beta\tau p}\right)$
4	 $0 \leq \alpha \leq \beta \leq 1$	PAIR (1) WITH A=B=E, C=O	$\frac{E}{\alpha\tau p^2}\left(1-\epsilon^{-\alpha\tau p}-\alpha\tau p\,\epsilon^{\beta\tau p}\right)$
5	 $0 \leq \beta \leq \gamma \leq 1$	PAIR (2) WITH $\alpha \longrightarrow 0$	$\frac{E}{p}\left[1-\frac{\epsilon^{-\beta\tau p}-\epsilon^{-\gamma\tau p}}{(\gamma-\beta)\tau p}\right]$
6	 $0 < \beta \leq 1$	PAIR (5) WITH $\gamma \longrightarrow \beta$	$\frac{E}{p}\left(1-\epsilon^{-\beta\tau p}\right)$

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RESPONSE OF NETWORKS

$\int f(t)\,dt$ becomes the accumulated charge, $q(t)$, and $[\int f dt]_{t=0}$ the initial charge, $q(o)$ (on a capacitor, for instance). Thus equations (2), (3a), and (3b) are re-written.

$$\left.\begin{aligned}\bar{i}(p)&=\int_0^\infty i(t)\,\epsilon^{-pt}dt\\ \overline{\left(\frac{di}{dt}\right)}&=-i(o)+p\,\bar{i}\\ \overline{\left(\int i dt\right)}&=\frac{q(o)}{p}+\frac{\bar{i}}{p}\end{aligned}\right\} \quad (4a)$$

We emphasize that $i(o)$ and $q(o)$ are the values of i and q at $t = o$. In case $i(o) = q(o) = o$ the transforms equations (4a) reduce to the important forms

$$\left.\begin{aligned}\bar{i}(p)&=\int_0^\infty i(t)\,\epsilon^{-pt}dt\\ \overline{\left(\frac{di}{dt}\right)}&=p\,\bar{i}\\ \overline{\left(\int i dt\right)}&=\frac{\bar{i}}{p}\end{aligned}\right\} \quad (4b)$$

Fundamental Examples

As fundamental examples, the transform of the instantaneous voltage drop across a resistor, Ri is $R\bar{i}$; the transform of the drop across an inductor, $L\frac{di}{dt}$ is $Lp\bar{i}$; the transform of the drop across a capacitor, $\frac{\int idt}{c}$, is $\frac{\bar{i}}{pc}$. Now if we compare these three forms with the expressions for voltage across these same elements when a steady alternating current flows through them, namely, RI, $j\omega LI$, and $\frac{I}{j\omega c}$, we see that we can get the transforms of the instantaneous voltage drops merely by writing p for $j\omega$ and $\bar{i}$ for the effective value I in the usual a-c expressions. Of course, this is true when, and only when $i(o) = q(o) = o$.

The transform of a constant, E, is very useful and is easily seen to be

$$\bar{E}(p)=E\int_0^\infty \epsilon^{-pt}dt=\frac{E}{p} \quad (5)$$

It is also seen that the transform of the sum of any number of functions is the sum of the transforms of the individual functions, providing all transforms concerned exist; for

$$\overline{\left[\sum_{K=1}^{n} f_K(p)\right]}=\int_0^\infty\left[\sum_{K=1}^{n} f_K(t)\right]\epsilon^{-pt}dt$$

$$=\sum_{K=1}^{n}\int_0^\infty f_K(t)\,\epsilon^{-pt}dt=\sum_{K=1}^{n}\bar{f}_K(p) \quad (6)$$

Application of the Laplace Transform

With this basic theory the solution of network problems is illustrated with the resistive inductor previously mentioned. We start, however, with a constant voltage E suddenly applied at $t = o$, so that

$$L\frac{di}{dt}+Ri=E\ ,\quad t>o \quad (7)$$

Multiply both sides by $\epsilon^{-pt}dt$, and integrate from zero to infinity. By equations (4) and (5)

$$-Li(o)+(pL+R)\,\bar{i}=\frac{E}{p}$$

$$\bar{i}=\frac{E}{Lp(p+\alpha)}+\frac{i(o)}{p+\alpha}\ ,\quad \alpha=\frac{R}{L}$$

$$=\frac{E}{Rp}-\frac{E}{R(p+\alpha)}+\frac{i(o)}{p+\alpha}$$

TABLE 1 CONT'D

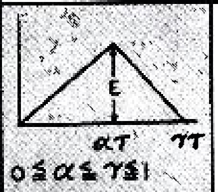
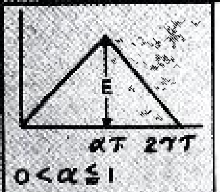
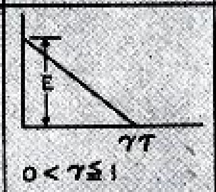
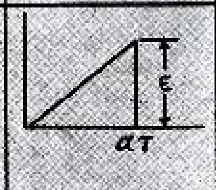
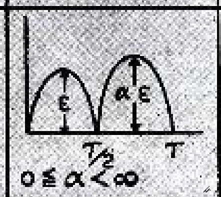
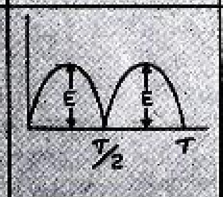
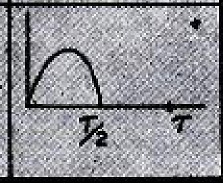
PAIR NO.	SINGLE PERIOD WAVE SHAPE	TIME FUNCTION $E_T(t)$	TRANSFORM $\bar{E}_T(p)$
7	 $0 \leqq \alpha \leqq \gamma \leqq 1$	PAIR(2) WITH $\beta=\alpha$	$\frac{E}{\alpha(\gamma-\alpha)Tp^2}\left\{\gamma-\alpha-\gamma\epsilon^{-\alpha Tp}+\alpha\epsilon^{-\gamma Tp}\right\}$
8	 $0<\alpha\leqq 1$	PAIR (7) WITH $\gamma=2\alpha$ (EQUAL SLOPES)	$\frac{E}{\alpha Tp^2}\left(1-\epsilon^{-\alpha Tp}\right)^2$
9	 $0<\gamma\leqq 1$	PAIR (7) WITH $\alpha\rightarrow 0$	$\frac{E}{p}\left(1+\frac{\epsilon^{-\gamma Tp}}{\gamma Tp}\right)$
10		PAIR (7) WITH $\gamma\rightarrow\alpha$	$\frac{E}{\alpha Tp^2}\left\{1-\left[1+\alpha Tp\right]\epsilon^{-\alpha Tp}\right\}$
11	 $0\leqq\alpha<\infty$	$E\sin\frac{2\pi}{T}t,\ 0\leqq t\leqq T/2$ $-\alpha E\sin\frac{2\pi}{T}t,\ T/2\leqq t\leqq T$	$\frac{2\pi E}{T}\frac{\left(1+\epsilon^{-p\frac{T}{2}}\right)\left(1+\alpha\epsilon^{-p\frac{T}{2}}\right)}{p^2+\left(\frac{2\pi}{T}\right)^2}$
12		PAIR (11) WITH $\alpha=1$	$\frac{2\pi E}{T}\frac{\left(1+\epsilon^{-p\frac{T}{2}}\right)^2}{p^2+\left(\frac{2\pi}{T}\right)^2}$
13		PAIR (11) WITH $\alpha=0$	$\frac{2\pi E}{T}\frac{1+\epsilon^{-pT/2}}{p^2+\left(\frac{2\pi}{T}\right)^2}$

Figure 2 (right)

From tables of integrals[5] it is readily ascertained that

$$\frac{1}{p+\alpha}=\int_0^{\infty}\epsilon^{-(p+\alpha)t}dt$$

i.e.

$$\left(\epsilon^{-\alpha t}\right)=\frac{1}{p+\alpha}$$

The well-known solution for the current is, therefore

$$i=\frac{E}{R}-\frac{E}{R}\epsilon^{-\alpha t}+i(o)\epsilon^{-\alpha t} \quad (8)$$

Now suppose that the rectangular wave of equation (1) is applied at $t=o$, and suppose further that $i(o)=o$. The transform for the voltage is

$$\bar{e}(p)=E\int_0^{T/2}\epsilon^{-pt}dt+E\int_T^{3T/2}\epsilon^{-pt}dt+\dots+E\int_{KT}^{(K+\frac{1}{2})T}\epsilon^{-pt}dt$$

$$=-\frac{E}{p}\left[-1+\epsilon^{-\frac{pT}{2}}+\epsilon^{-pT}-\epsilon^{-\frac{3pT}{2}}+\dots\dots\right]$$

$$=\frac{E}{p\left(1+\epsilon^{-\frac{pT}{2}}\right)} \quad (9)$$

The transformed differential equation is, therefore,

$$(pL+R)\bar{i}=\frac{E}{p\left(1+\epsilon^{-\frac{pT}{2}}\right)}$$

or

$$\bar{i}=\frac{E}{p(pL+R)\left(1+\epsilon^{-\frac{pT}{2}}\right)}=\int_0^{\infty}i(t)\epsilon^{-pt}dt \quad (10)$$

Equation *(10)* is not readily solved for $i(t)$ by recourse to a table of integrals. The best approach is the direct solution of equation *(2)* by application of the Mellin inversion theorem, which involves contour integration in the complex plane[1,6,7,8]. By this theorem the solution of equation (10) is

$$i(t)=\frac{E}{2\pi jL}\int_{Br}\frac{\epsilon^{pt}dp}{p(p+\alpha)\left(1+\epsilon^{-\frac{pT}{2}}\right)}, \quad \alpha=\frac{R}{L} \quad (11)$$

the integration being carried out along a contour in the complex p-plane represented by a line to the right of, and parallel to, the y-axis, closed by an infinite semi-circle to the left. The contour thus encloses all singularities[6,8] of the integrand of equation *(11)*, which are simple poles corresponding to

$$p(p+\alpha)\left(1+\epsilon^{-\frac{pT}{2}}\right)=o \quad (12a)$$

i.e.,

$$p=o,-\alpha$$

and

$$1+\epsilon^{-\frac{pT}{2}}=o$$

From the latter

$$\epsilon^{-\frac{pT}{2}}=-1$$

$$-\frac{pT}{2}=\log_\epsilon(-1)=\pm j(2n+1)\pi, \quad n=0,1,2,3$$

or,

$$p=\pm j\frac{2(2n+1)\pi}{T}, \quad n=0,1,2,3\dots\dots \quad (12b)$$

The solution of equation (11) is

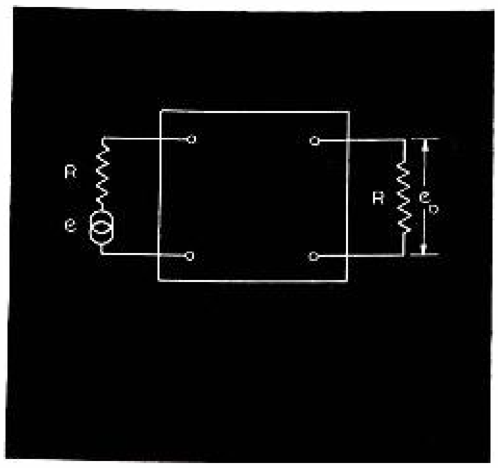


Figure 3
Four pole linear passive network terminated in resistances R.

simply the sum of the residues at the poles[1,6,8]. We have

$$Res(o)=\frac{1}{2\alpha}=\frac{L}{2R}$$

$$Res(-\alpha)=\frac{\epsilon^{-\alpha t}}{-\alpha\left(1+\epsilon^{\frac{\alpha T}{2}}\right)}=-\frac{L}{R}\frac{\epsilon^{-\alpha t}}{1+\epsilon^{\frac{\alpha T}{2}}}$$

$$Res\left(\pm j\frac{2(2n+1)\pi}{T}\right)$$

$$=\frac{\exp\left(\pm j\frac{2(2n+1)\pi}{T}\right)}{\pm j\frac{2(2n+1)\pi}{T}\left(\pm j\frac{2(2n+1)\pi}{T}+\alpha\right)\left[-\frac{T}{2}\exp\left(\mp j\frac{2(2n+1)\pi}{T}\cdot\frac{T}{2}\right)\right]}$$

Since these latter roots occur in conjugate pairs, we can write the sum of any conjugate pair as twice the real part of either of the pair: thus

$$Res\left(j\frac{2(2n+1)\pi}{T}\right)+Res\left(-j\frac{2(2n+1)\pi}{T}\right)$$

$$=\frac{2L}{(2n+1)\pi}\cdot\frac{R\sin(2n+1)\omega t-(2n+1)\omega L\cos(2n+1)\omega t}{R^2+(2n+1)^2\omega^2L^2}, \quad n=0,1,2,$$

where we have written $\frac{2\pi}{\tau}=\omega=$ fundamental angular frequency of the applied voltage. Putting these results in equation (11) we get, for the current

$$i=E\left\{\frac{1}{2R}-\frac{1}{R}\frac{\epsilon^{-\alpha t}}{1+\epsilon^{\alpha T/2}}+\frac{2}{\pi}\sum_{n=0}^{\infty}\frac{1}{2n+1}\left[\frac{R\sin\omega_n t-\omega_n L\cos\omega_n t}{R^2+\omega_n^2L^2}\right]\right\}$$

where

$$\omega_n=(2n+1)\omega, \quad n=0,1,2,\dots\dots\dots \quad (13)$$

Except for the second term in the right member, equation (13) represents the Fourier analysis of the steady-state response current. The second term is obviously the starting transient, since it tends to zero as t increases indefinitely, whereas the remaining terms repeat in magnitude with frequency ω.

The important point to notice is that only poles which lie to the left of the imaginary axis can give rise to terms in the solution which decay steadily with time; if a pole p_n lies on the imaginary axis, the corresponding factor $\epsilon^{p_n t}$ can only oscillate, **but never** decays.

Thus it is clear that the transient part of the solution is due to poles to the left of the imaginary axis; and by subtracting this transient part from the actual solution, we obtain the periodic solution.

Periodic Solution . . . Closed Form

To obtain the periodic solution in closed form, however, it is necessary to alter the method. The procedure, as stated previously, is, as nearly as the writer can ascertain, due to Professor Hazeltine. In equation *(11)* we substitute

$$\frac{1}{1+\epsilon^{-\frac{pT}{2}}}=\sum_{K=0}^{\infty}(-1)^K\epsilon^{-\frac{KpT}{2}}$$

getting

$$i(t)=\frac{E}{2\pi jL}\sum_{K=0}^{\infty}(-1)^K\int_{Br}\frac{\epsilon^{p(t-\frac{KT}{2})}dp}{p(p+\alpha)} \quad (14)$$

in which it can be shown[1] that

$$\int_{Br}\frac{\epsilon^{p(t-\frac{KT}{2})}dp}{p(p+\alpha)}=o \quad \text{for } t<\frac{KT}{2} \quad (15)$$

The physical interpretation of equations (14) and (15) is that equation *(14)* gives the solution for the current period by period. For our purpose, any one period after $t=o$ will suffice, and for greatest simplicity we choose the first period after $t=o$. Thus the actual current in the first period is

$$i_a(t)=\frac{E}{2\pi jL}\left\{\int_{Br}\frac{\epsilon^{pt}dp}{p(p+\alpha)}-\int_{Br}\frac{\epsilon^{p(t-T/2)}dp}{p(p+\alpha)}\right\} \quad (16)$$

but

$$\int_{Br}\frac{\epsilon^{pt}dp}{p(p+\alpha)}=\frac{1}{\alpha}\left[1-\epsilon^{-\alpha t}\right], \quad o<t<\infty$$

and

$$\int_{Br}\frac{\epsilon^{p(t-T/2)}dp}{p(p+\alpha)}=\frac{1}{\alpha}\left[1-\epsilon^{-\alpha(t-T/2)}\right], \quad T/2<t<\infty$$

$$=o, \quad o<t<T/2$$

Thus the actual current in the first period is

$$i_a(t)=\begin{cases}\frac{E}{R}\left(1-\epsilon^{-\alpha t}\right), & o<t<T/2\\ \frac{E}{R}\left(\epsilon^{-\alpha(t-T/2)}-\epsilon^{-\alpha t}\right), & T/2<t<T\end{cases} \quad (17)$$

The transient current is due to the pole at $p=-\alpha$. We have already found this (equation (13)) to be

$$i_t=-\frac{E}{R}\frac{\epsilon^{-\alpha t}}{1+\epsilon^{\alpha T/2}} \quad (18)$$

The periodic part of the current for the first period, and therefore for all periods, is

$$i_p=i_a-i_t$$

which, from equations *(17)* and *(18)*, is

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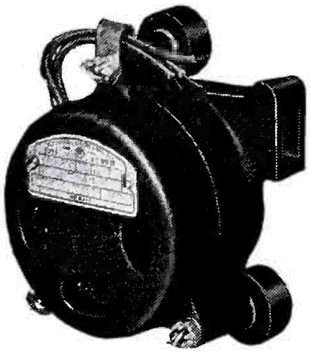

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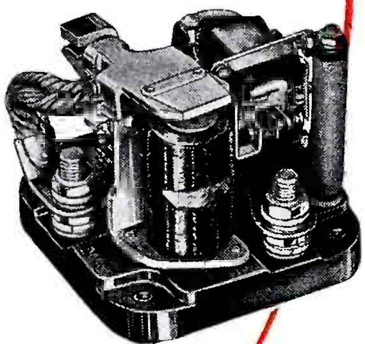

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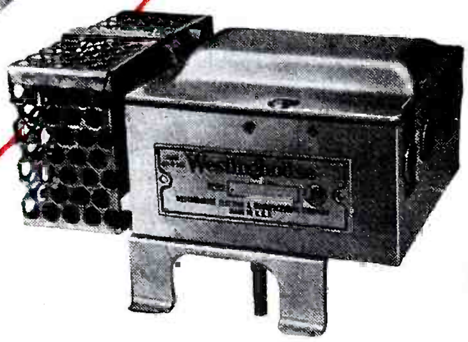

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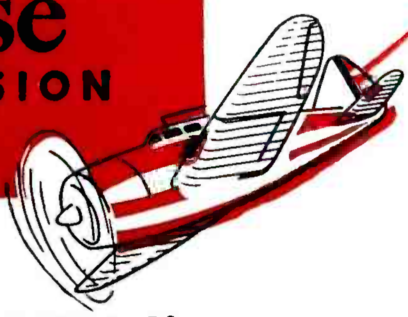

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$$i_p = \frac{E}{R}\left\{1 - \frac{\epsilon^{-\alpha(t - T/2)}}{1 + \epsilon^{\alpha T/2}}\right\}, \quad o < t < T/2$$
$$= \frac{E}{R}\,\frac{\epsilon^{-\alpha(t - T)}}{1 + \epsilon^{\alpha T/2}}, \quad T/2 < t < T \qquad (18)$$

Transforms for Periodic Voltages

We have already obtained the transform for the rectangular wave (equation (9)). The general expression for the transform of a function periodic after $t = o$ is readily obtained. Thus, let

$$\begin{aligned} E(t) &= o, \quad t < o \\ &= E(t + KT), \quad t > o,\ K = 1, 2, 3, \end{aligned} \qquad (19)$$

we have

$$\bar{E}(p) = \int_o^{\infty} E(t)\epsilon^{-pt}dt$$
$$= \sum_{K=o}^{\infty}\int_{KT}^{(K+1)T} E(t)\epsilon^{-pt}dt$$

In the general integral, let $t = \lambda + k\tau$, so that

$$\bar{E}(p) = \sum_{K=o}^{\infty}\int_o^{T} E(\lambda + KT)\epsilon^{-p(\lambda + KT)}d\lambda$$
$$= \sum_{K=o}^{\infty}\int_o^{T} E(\lambda)\epsilon^{-p(\lambda + KT)}d\lambda \qquad \text{by eq. (19)}$$
$$= \int_o^{T} E(\lambda)\epsilon^{-p\lambda}d\lambda \sum_{K=o}^{\infty}\epsilon^{-KTp}$$
$$= \frac{\int_o^{T} E(t)\epsilon^{-pt}dt}{1 - \epsilon^{-Tp}} = \frac{\bar{E}_T(p)}{1 - \epsilon^{-Tp}} \qquad (20a)$$

where

$$\bar{E}_T(p) = \int_o^{T} E(t)\epsilon^{-pt}dt \qquad (20b)$$

The values of $\bar{E}_\tau(p)$ for some useful types of wave-forms are listed in Table 1, Figures 1 and 2. The transforms are listed as "pairs" in line with accepted terminology[4]. The first column lists the pair number used for identifying each function. The second column shows the shape of the voltage wave and the third column indicates its analytical formulation. Finally the last column gives the value of $\bar{E}_\tau(p)$ which is to be divided by $(1 - \varepsilon^{-p\tau})$ to get the whole transform.

Pair (1) which appears quite impractical, is given first because from it pairs (2) to (10) inclusive, are readily derived by passage to suitable limits. In particular, if in pair (6) we set $\beta = 1/2$ we get

$$\bar{E}_T = \frac{E}{p}\left(1 - \epsilon^{-\frac{Tp}{2}}\right)$$

so that the transform is

$$\bar{E} = \frac{E}{p}\,\frac{1 - \epsilon^{-\frac{Tp}{2}}}{1 - \epsilon^{-Tp}} = \frac{E}{p\left(1 + \epsilon^{-\frac{Tp}{2}}\right)}$$

as previously obtained (equation (9)).

Response to "Square" Waves

Let a voltage given by equation (1)

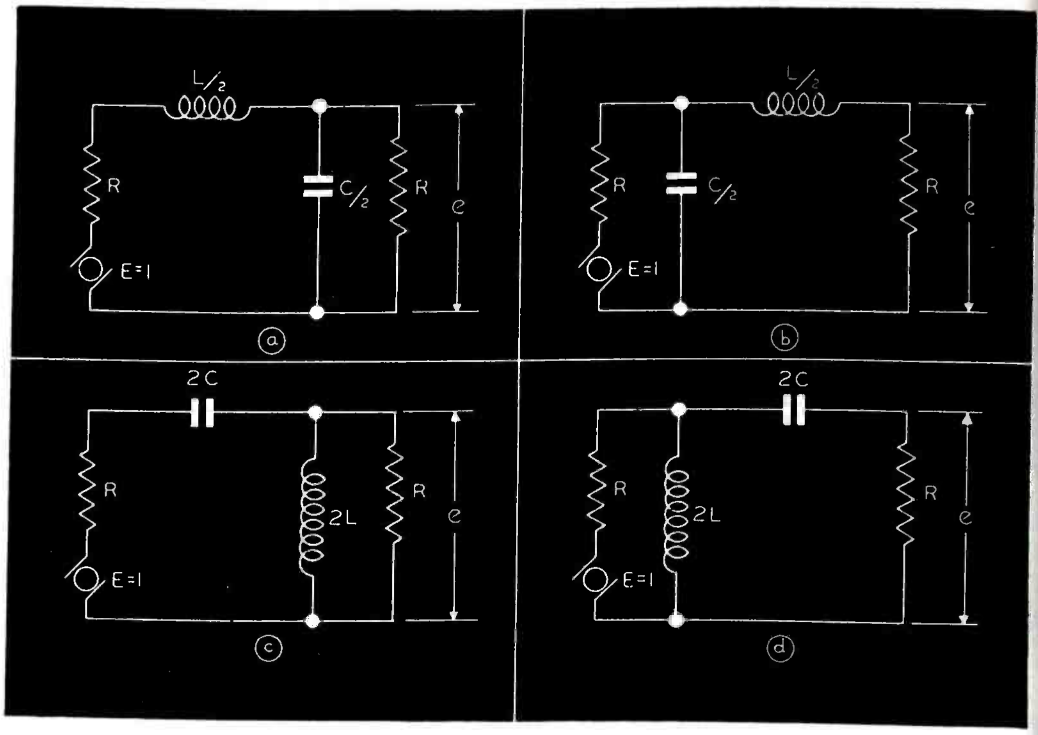

Figure 4

Wave filter half sections, wherein (a) is a low-pass mid-series input type; (b)—low-pass mid-shunt input; (c)—high-pass mid-series input; (d)—high-pass mid-shunt input. By the reciprocity theorem, the output voltage e is the same for (a) as for (b), when the applied voltage E is the same for both. The statement applies also to (c) and (d).

be applied to a four-pole linear passive network (Figure 3) terminated in resistances R. For quiescent initial conditions (all charges and currents equal to zero), the transforms of the differential equations of the network may be written in the form

$$\sum_{\beta=1}^{n} \bar{Z}_{\alpha\beta}\bar{I}_\beta = \bar{E}_\alpha, \quad (\alpha = 1, \ldots\ldots, n)$$

in which the Z's are merely the mutual- and self-impedances of the n meshes of the network at angular frequency ω, but with $j\omega$ replaced by p, and in which

$$\begin{aligned} \bar{E}_1 &= \bar{e} \\ E_\alpha &= o, \quad (\alpha = 2, 3, \ldots\ldots, n) \end{aligned}$$

$\bar{e}$ being given by equation (9). From these the output voltage transform is

$$\bar{e}_o = R\,\frac{Y_o(p)}{Z_o(p)}\,\frac{E}{p\left(1 + \epsilon^{-\frac{pT}{2}}\right)} \qquad (21)$$

where Z_o is the determinant of the network, and Y_o is the cofactor in Z_o, of the row that corresponds to the input mesh and the column that corresponds to the output current[9]. From these the output voltage is

$$e_o = \frac{RE}{2\pi j}\int_{Br}\frac{Y_o(p)\epsilon^{pt}dp}{pZ_o(p)\left(1 + \epsilon^{-\frac{pT}{2}}\right)} \qquad (22)$$

provided this integral converges when an appropriate contour for integration is selected. Assuming the network parameters may be lumped, the admittance Y_o/Z_o can be converted to a rational fractional form. Let this form be $A(p)/B(p)$ where A and B are rational integral polynominals in p.

As previously stated the solution consists of a periodic part and a transient part. For passive networks all zeros of $pB(p)\,(1 + \varepsilon^{-\tau p/2})$ lie on, or to the left of, the imaginary axis. The transient part of the solution is obtained by finding the residues of the poles to the left of the axis. Such poles can only be contained in the function $B(p)$. It is readily seen that all roots of $B(p) = o$ must lie to the left of the imaginary axis. For by fundamental circuit theory[9] the roots of $B(p) = o$ determine the natural modes of oscillation of the network. In any physical network these must be such that the oscillations die out eventually; i.e., the roots must have a negative real part. Suppose $B(p) = o$ has only simple roots $a_k (k = 1, \ldots, n)$ such that $A(a_k) \neq o$ for every k. The residue at any point a_k is

$$\mathrm{Res}(a_K) = \frac{A(a_K)\epsilon^{a_K t}}{a_K B'(a_K)\left(1 + \epsilon^{-\frac{Ta_K}{2}}\right)}, \quad K = 1, \ldots, n \qquad (23)$$

where

$$B'(a_K) = \left.\frac{dB}{dp}\right]_{p = a_K}$$

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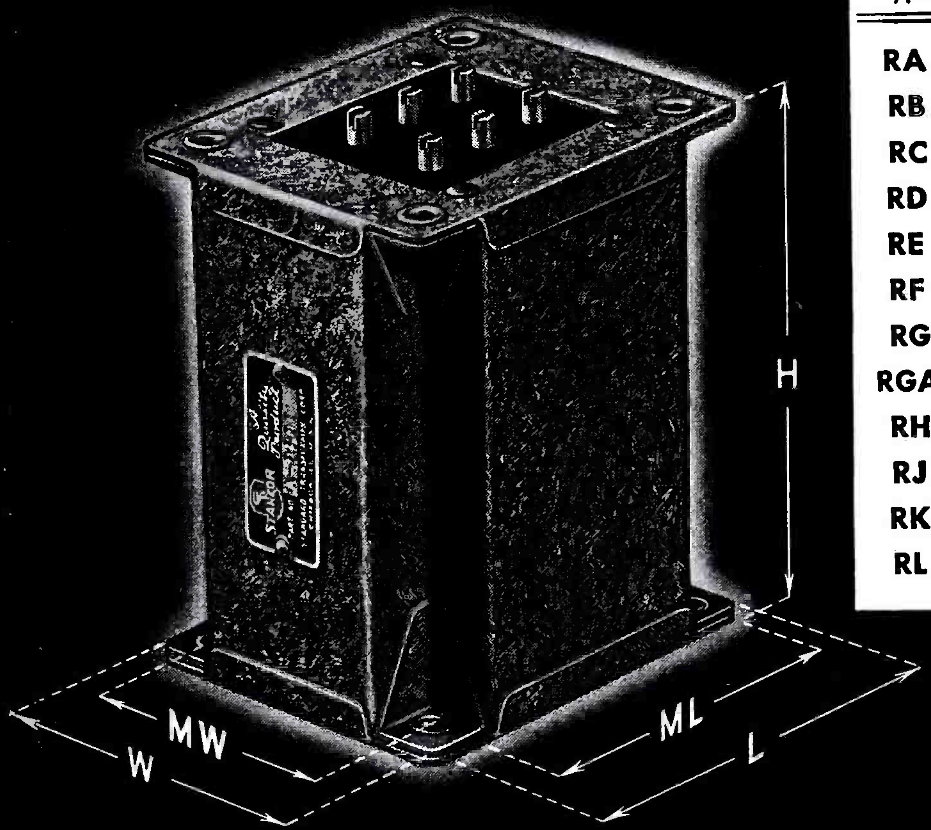

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RG	5"	5⅛"	5"	4½"	4 5/16"
RGA	7⅛"	5 11/16"	6 9/16"	4 13/16"	5¾"
RH	9"	7½"	7"	6½"	6"
RJ	9"	8¾"	7½"	7¾"	6½"
RK	9"	8¾"	9"	7¾"	8"
RL	13"	8¾"	10"	7¾"	9"

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Then the transient part of the solution is

$$e_t = RE\sum_{K=0}^{n} \frac{A(a_K)\epsilon^{a_K t}}{a_K B'(a_K)\left(1+\epsilon^{\frac{-\tau a_K}{2}}\right)} \quad ,t>0 \qquad (24)$$

The actual response in the first period is given by

$$e_a = \frac{RE}{j2\pi}\left\{\int_{Br} \frac{A(p)\epsilon^{pt}dp}{pB(p)} - \int_{Br} \frac{A(p)\epsilon^{p(t-\frac{\tau}{2})}dp}{pB(p)}\right\}$$

$$= RE\left[\frac{A(o)}{B(o)} + \sum_{K=1}^{n} \frac{A(a_K)\epsilon^{a_K t}}{a_K B'(a_K)}\right] \quad ,0<t<\frac{\tau}{2}$$

$$= RE\sum_{K=1}^{n}\left[\frac{A(a_K)}{a_K B'(a_K)}\right]\left[\epsilon^{a_K t} - \epsilon^{a_K(t-\frac{\tau}{2})}\right] \quad ,\frac{\tau}{2}<t<\tau \qquad (25)$$

The periodic response is, from eqs. (24) and (25)

$$e_p = e_a - e_t$$

$$= RE\left\{\frac{A(o)}{B(o)} + \sum_{K=1}^{n} \frac{A(a_K)\epsilon^{a_K t}}{a_K\left(1+\epsilon^{\frac{a_K \tau}{2}}\right)B'(a_K)}\right\} \quad ,0<t<\frac{\tau}{2}$$

$$= -RE\left\{\sum_{K=1}^{n} \frac{A(a_K)\epsilon^{a_K(t-\tau/2)}}{a_K\left(1+\epsilon^{a_K \tau/2}\right)B'(a_K)}\right\} \quad ,\frac{\tau}{2}<t<\tau \qquad (26)$$

The first of these forms can be considered as an extension of Heaviside's Expansion Formula; in fact, for $\tau = \infty$, it becomes the Heaviside expansion for the response to the "unit" function[10].

By means of the results obtained in the preceding section we consider the application of the rectangular wave, to constant $-k$ type filters, and take first the case of the low-pass half-sections of Figures 4a and 4b. By the reciprocity theorems the response for these configurations is the same. If f_c is the filter cut-off frequency, in cycles per second, then the relationships between the parameters are

$$L = \frac{R}{\pi f_c}, \quad C = \frac{1}{\pi f_c R} \qquad (27)$$

where L = inductance, henrys; C = capacitance, farads, and R = resistance, ohms.

From the differential equations of either network, the transform of the output voltage is

$$\bar{e} = \frac{4\pi^2 f_c^2}{p^2 + 4\pi f_c p + 8\pi^2 f_c^2}\bar{E} \qquad (28)$$

Comparing this with the results of the previous section we have

$$A(p) = \frac{4\pi^2 f_c^2}{R}$$

$$B(p) = p^2 + 4\pi f_c p + 8\pi^2 f_c^2$$

whence

$$B'(p) = 2p + 4\pi f_c$$

$$a_1, a_2 = -2\pi f_c(1 \pm j) = -\alpha \pm j\beta$$

where $\alpha = \beta = 2\pi f_c$. From these we get

$$\frac{A(o)}{B(o)} = \frac{1}{2R}$$

$$B'(-\alpha + j\beta) = j4\pi f_c$$

$$B'(-\alpha - j\beta) = -j4\pi f_c$$

$$A(-\alpha + j\beta) = A(-\alpha - j\beta) = \frac{4\pi^2 f_c^2}{R}$$

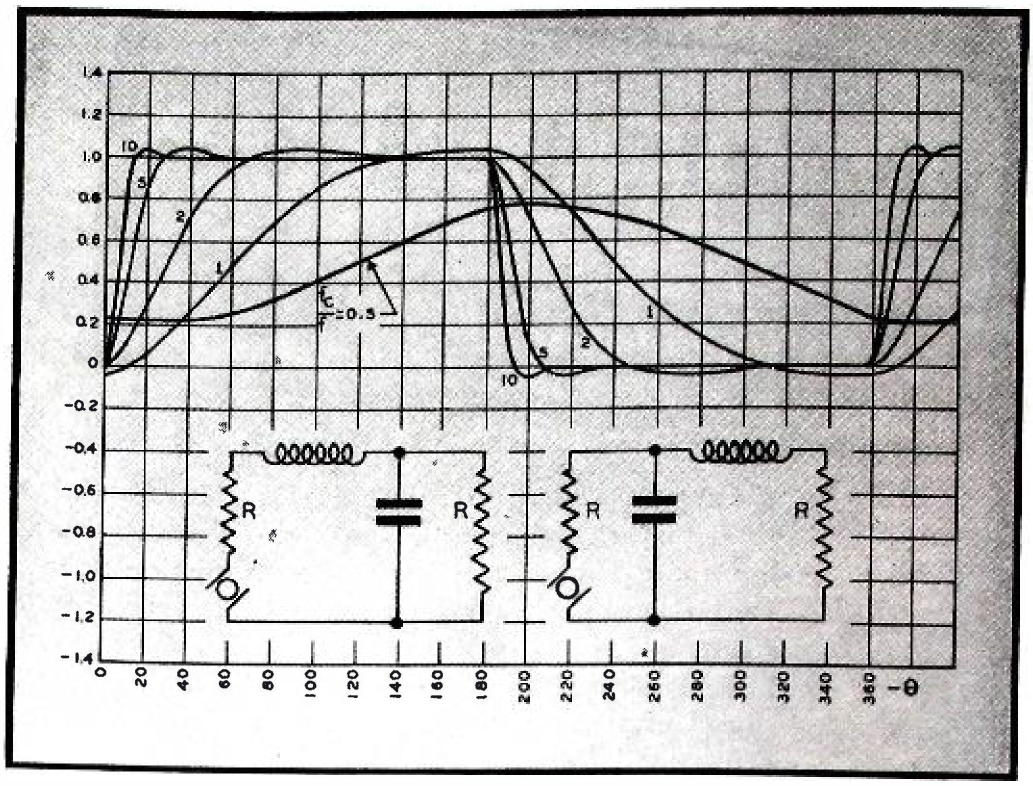

Figure 5
Response of low-pass, constant-K half-section filters to rectangular waves.

Hence the periodic solution is, by equations (26), with $E = 1$,

$$e_p(t) = R\left\{\frac{1}{2R} + \frac{4\pi^2 f_c^2 \epsilon^{(-\alpha+j\beta)t}}{R(-\alpha+j\beta)\left(1+\epsilon^{a_1 \tau/2}\right)(j4\pi f_c)} + \frac{4\pi^2 f_c^2 \epsilon^{(-\alpha-j\beta)t}}{R(-\alpha-j\beta)\left(1+\epsilon^{a_2 \tau/2}\right)(-j4\pi f_c)}\right\} \quad ,0<t<\tau/2$$

and a similar equation for the second-half period. This reduces, after some manipulation, to

$$e_p(t) = \frac{1}{2}\left\{1 - \sqrt{2}\,\frac{\sin(\omega_c t + \frac{\pi}{4}) + \epsilon^{-\frac{\omega_c \tau}{2}}\cos\omega_c\left[(t-\tau/2)+\pi/4\right]}{1+\epsilon^{-\omega_c \tau} + 2\epsilon^{-\omega_c \tau/2}\cos\frac{\omega_c \tau}{2}}\,\epsilon^{-\omega_c t}\right\}$$

$$,0<t<\tau/2 \qquad (29a)$$

where $\omega_c = 2\pi f_c$. If τ is large this reduces to

$$e_p(t) = \frac{1}{2}\left\{1 - \sqrt{2}\sin\left(\omega_c t + \frac{\pi}{4}\right)\epsilon^{-\omega_c t}\right\} \quad ,0<t<\tau/2 \qquad (29b)$$

Equation (29a) is plotted in Figure 5 for various values of f_c/f, where f is the fundamental dot frequency and is therefore equal to $\frac{1}{\tau}$.

From Figure 5 it can be seen that the essential character of the wave is not disturbed if the cut-off frequency exceeds the dotting frequency by a factor of, say, 5 or more. To retain the original steepness of the vertical sides of the wave is much more difficult; even for $\frac{f_c}{f} = 10$ the slant is quite appreciable.

Considering next the case of the high-pass constant $-k$ half-section filters of Figures 4c and 4d, we have the filter design equations

$$C = \frac{1}{4\pi f_c R}, \quad L = \frac{R}{4\pi f_c} \qquad (30)$$

Following a procedure entirely similar to that for the low-pass filters we obtain

$$\bar{e} = \frac{1}{2}\left\{\frac{p^2}{p^2 + 2\pi f_c p + 2\pi^2 f_c^2}\right\}\bar{E} \qquad (31)$$

The roots of the denominator are

$$a_1, a_2 = (-1 \pm j)\pi f_c$$

and the remaining necessary information is as follows:

$$\frac{A(o)}{B(o)} = 0$$

$$A(p) = \frac{p^2}{2R}; \quad A(a_1) = -j\frac{\pi^2 f_c^2}{R}; \quad A(a_2) = j\frac{\pi^2 f_c^2}{R}$$

$$B(p) = p^2 + 2\pi f_c p + 2\pi^2 f_c^2; \quad B'(p) = 2p + 2\pi f_c$$

$$B'(a_1) = j\cdot 2\pi f_c; \quad B'(a_2) = -j\cdot 2\pi f_c$$

Substituting in eq(26) and reducing, we get for the response

$$e_p(t) = \frac{1}{\sqrt{2}}\left\{\frac{\cos\left(\frac{\omega_c t}{2} + \frac{\pi}{4}\right) + \epsilon^{-\frac{\omega_c \tau}{4}}\cos\left[\frac{\omega_c}{2}\left(t - \frac{\tau}{2}\right) + \frac{\pi}{4}\right]}{1 + 2\epsilon^{-\frac{\omega_c \tau}{4}}\cos\frac{\omega_c \tau}{4} + \epsilon^{-\frac{\omega_c \tau}{2}}}\,\epsilon^{-\frac{\omega_c t}{2}}\right\}$$

$$,0<t<\frac{\tau}{2} \qquad (32)$$

Equation (32) is plotted in Figure 6 for various values of f_c/f. It is readily seen that even for $\frac{f_c}{f} = 1$, where the filter should be passing freely all frequency components of the wave, and practically has a small at-

(Continued on page 74)

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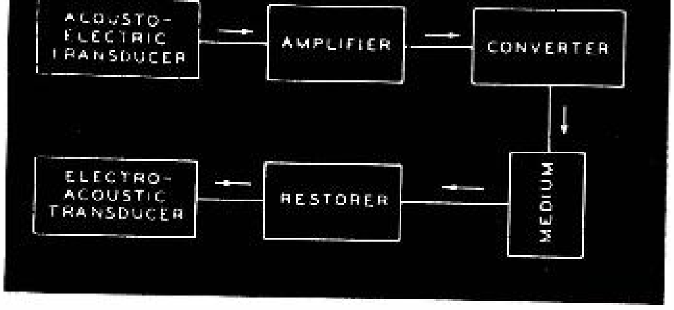

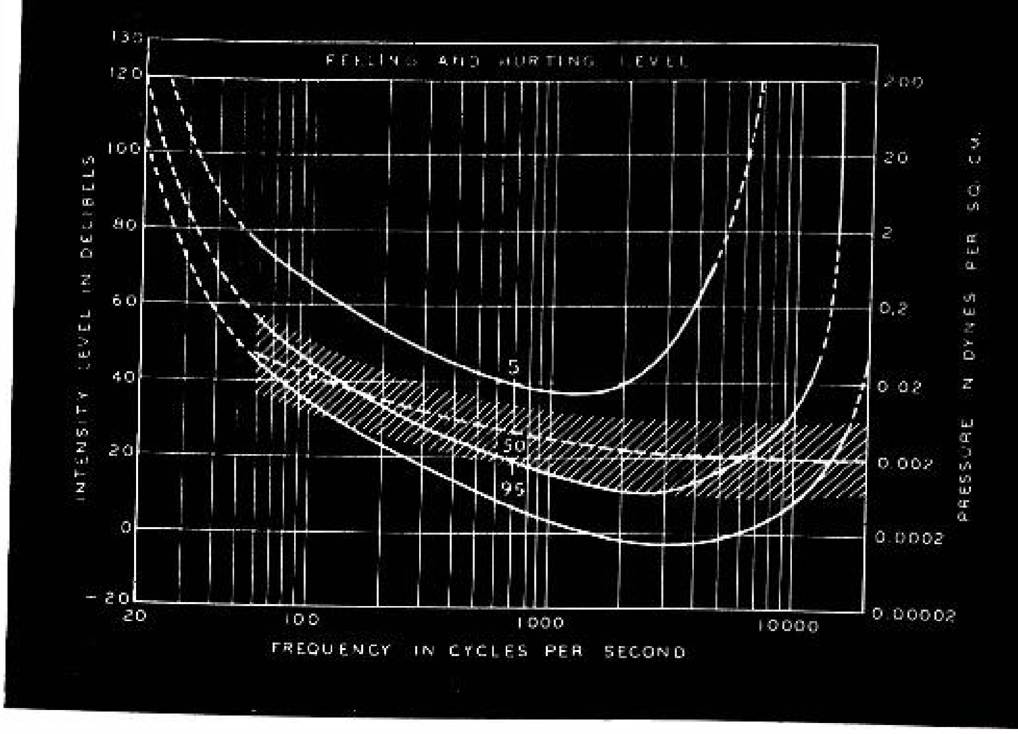

Figures 1 (top) and 2 (right)

In Figure 1, we have the six fundamental links or components of every transmission system. Although the components take different forms for the various transmission systems, their fundamental functions are identical for each system. Figure 2 illustrates the contours of hearing loss and room noise.
(*Figure 2, Courtesy Proceedings, IRE*)

CHARACTERISTICS OF

by A. JAMES EBEL
Chief Engineer, WILL

IN 1929 a new line of *high fidelity* radios was introduced to the public as the last word in faithful reproduction. A few years later ***high fidelity*** sound systems were installed in the better theatres of the nation. And back in the early twenties, orthophonic recording was the development that brought "true realism" to recorded music. By present day comparisons those electro-acoustic systems are all decidedly inferior. At the present time we have ***high fidelity*** sound reproduction in the theatre sound systems, in both a-m and f-m radio broadcasting and in radio transcriptions. But has the art developed to the point so that we have true fidelity transmission in electro-acoustic systems or is true fidelity a goal for which we strive but never quite reach?

Electro-Acoustic Systems

The term ***high fidelity*** in itself has a relative intonation, so that what was ***high fidelity*** in 1929 wasn't ***high fidelity*** five years later and what is ***high fidelity*** today, won't necessarily be ***high fidelity*** ten years from now. The term has gained wide usage in engineering circles, however, and therefore it would be well to study electro-acoustic systems from that standpoint.

At the outset it would be well to point out that excellent work has been done and is being done in this field by workers too numerous to mention here. It is the research of these men and their laboratories that makes possible the definition of standards of reproduction which we shall coinsider.

Every transmission system may be broken down into six fundamental links or components. These components are shown in the block diagram, Figure 1. The components take different forms for the different transmission systems but their fundamental functions are identical for each system. The systems we shall consider here are . . . (*1*)—radio broadcasting, and (*2*)—recording. The latter includes both disc recordings used for phonograph and transcription reproduction, and sound on film recording used for sound picture reproduction.

The first link, an acousto-electric transducer, is common to all systems, since the microphones used are interchangeable. The desired characteristics of such a transducer are that it convert variations in air pressure or air particle velocity to identical changes in electric potential. This requires that the wave form of the voltage generated by the microphone should be in all manners similar to that of the sound wave in air.

The wave form of the sound emitted by a musical instrument whose fundamental tone is determined by ω_1 may be expressed by . . .

$$P = P_0 + A_1 \sin \omega_1 t + A_2 \sin 2\omega_1 t + A_3 \sin 3\omega_1 t \; . \; . \; . \; A_n \sin n\omega t$$

Where P_0 is atmospheric pressure, $A_1, A_2, A_3 \ldots A_n$ are coefficients of the overtone amplitudes. The acousto-electric transducer will deliver voltage in the form . . .

$$E_m = K_{ae} A_1 \sin \omega_1 t + K_{ae} A_2 \sin 2\omega_1 t + K_{ae} A_3 \sin 3\omega_1 t + \ldots K_{ae} A_n \sin n\omega_1 t$$

Where K_{ae} is the conversion factor of instrument.

If it is to function properly the conversion factor, K_{ae}, will be a constant for all values of $A_1, A_2, A_3 \ldots A_n$, and for all values of ω_1 or multiples thereof within the audible range. The expression for a single tone, given in equation (1) doesn't take into consideration the transient phenomena that occurs as the tone is started and stopped.

During the transient period the relative values of $A_1, A_2, A_3 \ldots A_n$, may change rapidly through very wide ranges and new terms with values of ω, not integral multiples of the fundamental tone ω_1 may be introduced.

The more complex wave form, generated by a number of instruments sounding tones simultaneously, with different values of fundamental frequencies, may be written as the sum of the series expansions for each value of ω, with the coefficients evaluated in terms of the relative amplitudes of the various waves. The above requirement for the transducer still holds, i.e., the value of K_{ae} must be constant for all values of A and ω.

Amplifier, Second Link

The amplifier, shown as the second link of the system, is also common to the two transmission systems. To be perfect in its function the amplifier must amplify the complex wave set

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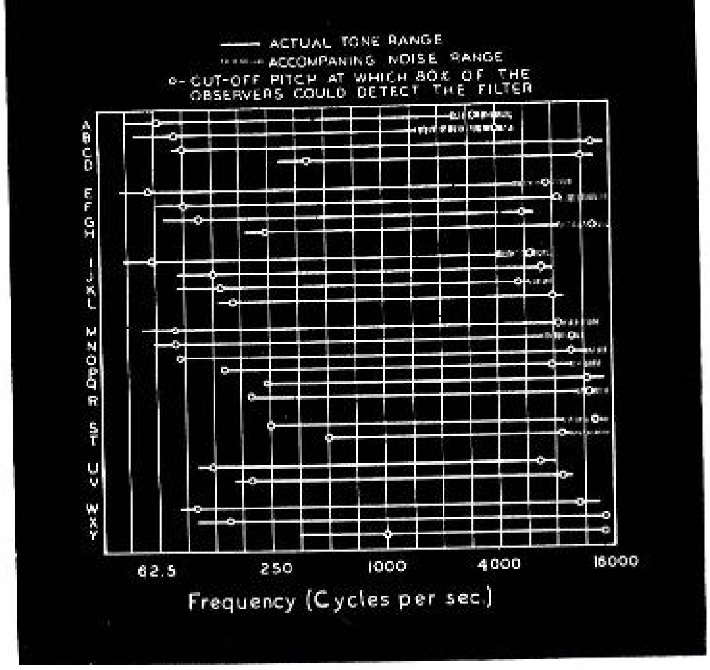

Figure 3

The audible frequency range for speech, music, and noise. Ranges for the following instruments are shown . . . A, tympani; B, bass drum; C, snare drum; D, 14" cymbals; E, bass viol; F, cello; G, piano; H, violin; I, bass tuba; J, trombone; K, french horn; L, trumpet; M, bass saxophone; N, bassoon; O, bass clarinet; P, clarinet; Q, oboe; R, soprano saxophone; S, flute; T, piccolo; U, male speech; V, female speech; W, footsteps; X, hand clapping, and Y, key jingling.

(Courtesy, Journal of Acoustic Society of America)

HIGH FIDELITY SYSTEMS

up by the microphone without changing its shape in anyway; the change should be one of magnitude only.

The wave form of the output of the amplifier will be given by . . .

$$E_a = K_a E_m = K_a K_{ae} A_1 \sin \omega_1 t + K_a K_{ae} A_2 \sin 2\omega_1 t + K_a K_{ae} A_3 \sin 3\omega_1 t \ldots K_a K_{ae} A_n \sin n\omega_1 t$$

Where K_a is the gain of the amplifier.

Here again K_a should be constant for all values of A and ω_1. The gain, K_a required, is determined by the ratio of the available input voltage to the specified output voltage necessary to properly operate the next link of the system. This varies from 10^7 for some recording systems to 10^{15} for a broadcast transmitter.

The third link shown in Figure 1, the converter, can be better discussed after the fourth link, the medium, has been treated since the form of the converter depends to a large extent on the medium. There are a number of mediums used by the transmission systems being considered. For radio broadcasting, both f-m and a-m, there is the nameless medium in which electromagnetic waves travel. In considering this medium, the characteristics of both the medium and of the electromagnetic waves used as carriers in the medium must be evaluated. In the case of disc recording the medium is wax or some soft acetate material in which the modulated groove is engraved or embossed. For sound on film the medium is a section of the picture film, on which the modulation is affected by a variation of density or by a variation of light area. If magnetic tape recording is considered, the medium is the magnetic characteristic of the tape. Any media, which will transmit variations of the form of the complex wave given in equation (*1*) instantaneously, or which will receive impressions proportional to such a wave and retain them until transported to another location, may be considered as a transmission medium. There are undoubtedly many mediums which have not been explored at all at the present time.

The required characteristic of the medium is that it transmit the wave without changing its form. In general, transmission through a medium involves a loss so that the expression for the complex wave will be further modified by the application of the constant $1/K_m$ to each term of the series expansion. As in previous constants, K_m should be constant for all values of A and ω within the audible limits.

Converters Range

Converters range from the massive structures of high power radio transmitters to the small recording head. There are two general types of converters used in radio broadcasting . . . *amplitude modulated* and *frequency modulated.* They differ in their manner of utilizing the electromagnetic wave to transmit the variations originated at the microphone. For disc recording there are two general types of recording heads, namely . . . *vertical* and *lateral.* Sound film recording is done in several ways with the *variable area* and the *variable density* methods as outstanding. All of these converters must have a conversion characteristic K_c which is constant under the frequency and amplitude limits.

Restorers, Fifth Link

Restorers, the fifth link in the system, vary with the system, just like the converters. In general they are not as large or complex as the converters. For radio broadcasting they are the f-m and a-m receivers; for recording, the pickup with its associated amplifier, whether it be a magnetic or piezoelectric pickup for disc recording, or a photoelectric pickup for sound track. This fifth link might well be broken down into two links, the restorer and its associated amplifier. However, in the case of radio receivers, the two are very hard to separate and no particular benefit is gained by the separation. The same is true of the f-m or a-m transmitter. Amplification takes place in both units before and after conversion. The restoration characteristic K_r will include both the reconversion factor and the gain of the associated amplifiers. This characteristic must also be constant within the amplitude and frequency requirements defined by the equation for the complex wave being transmitted.

The final element of the system is the electro-acoustic transducer which is common to all systems. This device must change complex waves of electric variation to a similar variation of air pressure which will constitute a sound wave. This resulting sound wave must be identical in form to the electric wave setting it up in the electro-acoustic transducer, so that the conversion factor K_{ea} should be a constant within the limits of all possible

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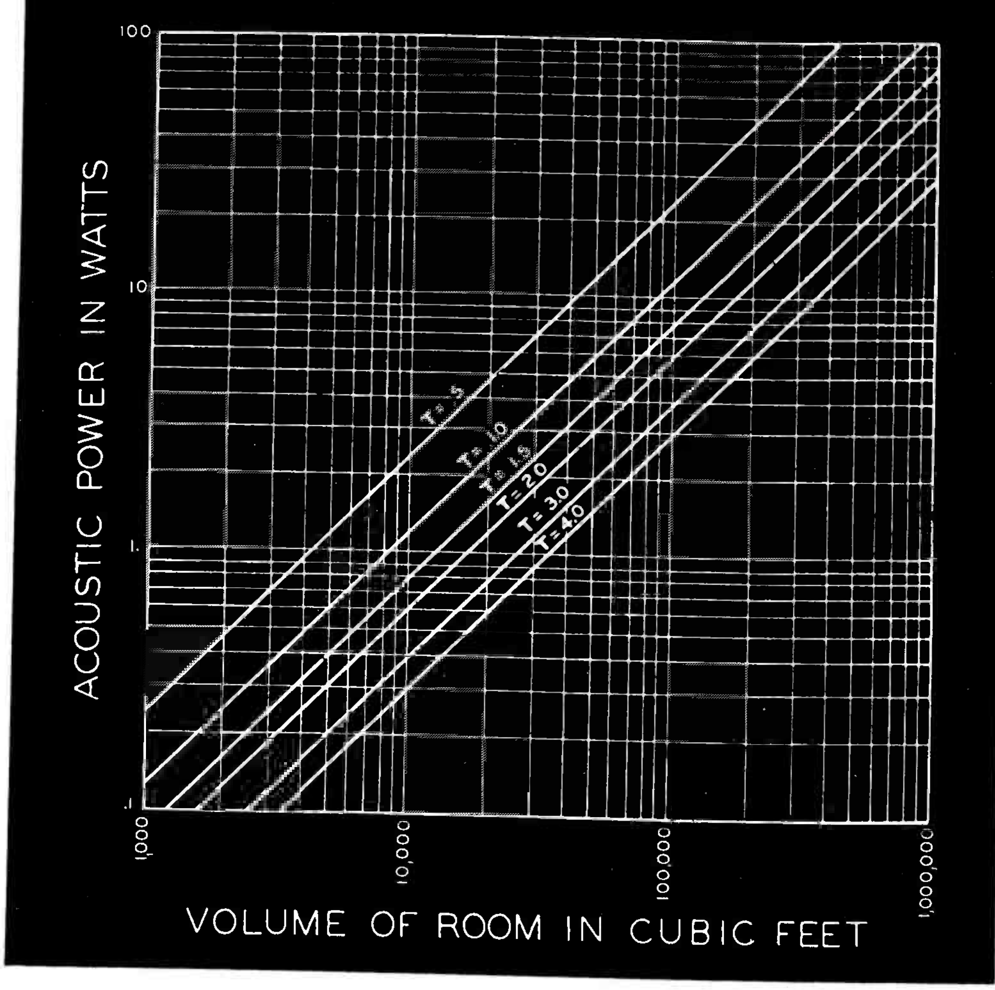

Figure 5

the first two terms of the series of a nonlinear system with two frequencies applied will give . . .

$$i = a_1E_1 \cos \omega_1 t + a_1E_2 \cos \omega_2 t + \frac{a_2E_1^2}{2} + \frac{a_2E_1^2}{2} \cos 2\omega_1 t + a_2E_1E_2 \cos (\omega_1 + \omega_2) t + a_2E_1E_2 \cos (\omega_1 - \omega_2) t + \frac{a_2E_2^2}{2} + \frac{a_2E_2^2}{2} \cos 2\omega_2 t$$

where E_1 is the amplitude of the frequency ω_1;
where E_2 is the amplitude of the frequency ω_2; and
where a_1 and A_2 are the coefficients of the first two terms of the power series of the non-linear element.

The terms $a_2E_1E_2 \cos (\omega_1 + \omega_2)t$ and $a_2E_1E_2 \cos (\omega_1 - \omega_2)t$ will appear as distortion and will be objectionable because they bear no harmonic relationship to the fundamental tones. The two second harmonic terms, $\frac{a_2E_1^2}{2} \cos 2\omega_1 t$ and $\frac{a_2E_2^2}{2} \cos 2\omega_2 t$ while giving rise to distortion will not be so objectionable since they only change the overtone distribution of the tone and at low levels will not be recognized. The objectionability of an inharmonic distortion component depends on the amplitude of the component, a function of the nonlinearity of the system, and upon the masking effect which depends not only on the relative amplitudes but also on the relationship of the frequencies. Figure 4 shows the rise in the apparent threshold level of a tone when masked by another. From this it may be seen that the masking is greatest when the tones are fairly close together and when the masking tone is of higher pitch than the masked tone. In the case of the sum and difference frequency components of equation (*5*), the amplitude of the component is given by the coefficient $a_2E_1E_2$. The two fundamental frequencies may be considered to be masking the sum and difference frequencies. From the characteristics of the masking phenomena it is apparent that the sum frequency will be masked less and therefore will be more objectionable than the difference frequency. This fact is borne out by the observation that as the response band of the system is increased toward the high frequency end the more objectionable a fixed amount of distortion becomes. If an expansion of the third, fourth, and fifth term of the power series for a nonlinear element were carried out, frequency terms shown in Figure 6 will be obtained.

From this it may be seen that if a distortion-free system is to be maintained the amplitude of all distortion components must be below the apparent threshold established by masking effect of the fundamental tones. This will not allow for very much distortion when converted to percent. Present day practice seems to dictate between 1 and 2 percent as the maximum allowable for the perfect system.

If the system is to transmit the full range it must have the necessary power capacity to reproduce the loudest passages within the limits of the distortion rating mentioned above. The required capacity will then depend upon two factors not within the system; the volume of the room and its time constant. Figure 5 shows the relationship between the required acoustic power and the volume of the room for various time constants. This power is that which will create a 120 db sound intensity level in the room which is the maximum that could ever be required.

The transient characteristics of music were referred to briefly above. Every tone has a transient characteristic in its rise and decay. In general the more rapid the rise and decay the wider will be the band of frequency distribution about the fundamental. This relationship is shown in Figure 7. This factor lends much of the characteristic tone to the sounds emitted by various musical instruments and must be preserved if the tone is to be faithfully reproduced. There are two factors to be considered in connection with maintaining this transient phenomena. First, the system must transmit in their original proportion all frequencies generated as a result of the phenomena, and secondly, none of the elements of the system must change the transient characteristic by

(*Continued on page* 44)

Expanded Term	**1st**	**2nd**	**3rd**	**4th**	**5th**
Resultant Frequencies	f_1 f_2	$2f_1$ $2f_2$ f_1-f_2 f_1-f_2	$3f_1$ $3f_2$ $2f_1$- f_2 $2f_1$- f_2 f_1-$2f_2$ f_1-$2f_2$	$4f_1$ $4f_1$ $3f_1$- f_2 $3f_1$- f_2 $2f_1$-$2f_2$ $2f_1$-$2f_2$ f_1-$3f_2$ f_1-$3f_2$	$5f_1$ $5f_2$ $4f_1$- f_2 $4f_1$- f_2 $3f_1$-$2f_2$ $3f_1$-$2f_2$ $2f_1$-$3f_2$ $2f_1$-$3f_2$ f_1-$4f_2$ f_1-$4f_2$

Figure 6

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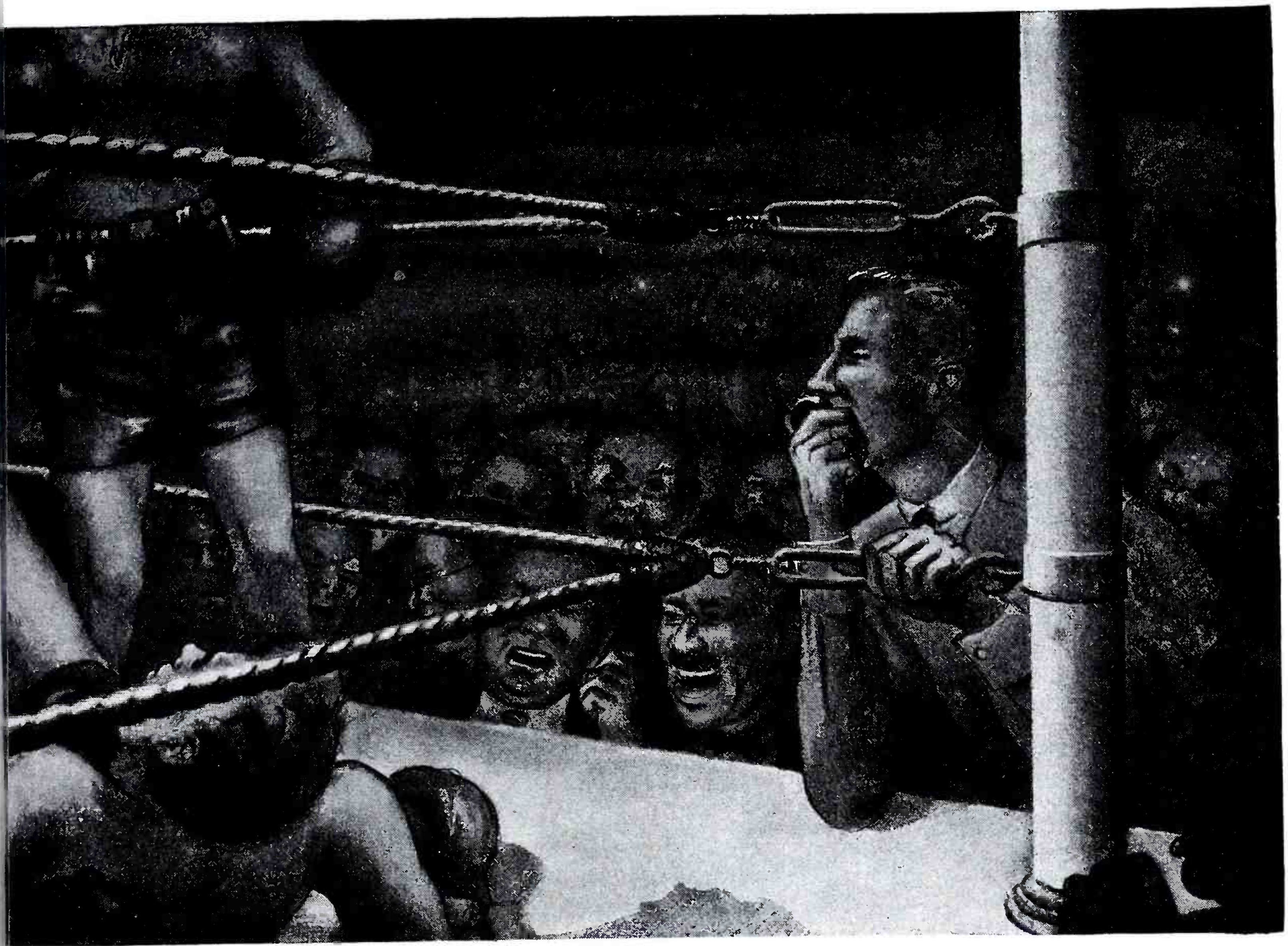

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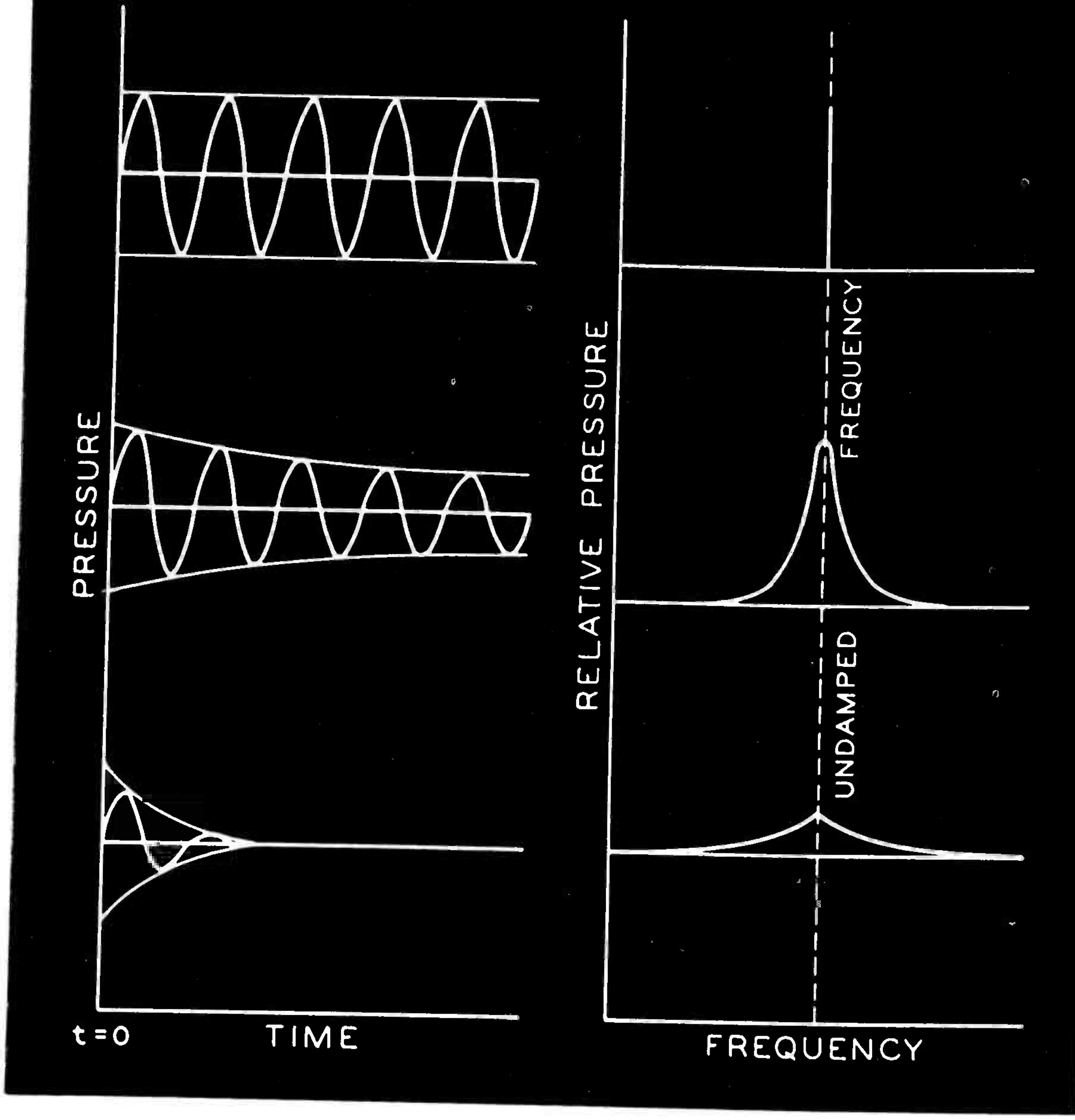

Figure 7

having a transient characteristic of its own. There is very little data on how small a transient variation is perceptible. It is known that electro-acoustic transducers are serious offenders in this respect. The possibility that resonant equalizers could have a decrement of such value as to be noticeable might account for the splattering "s" sounds on some highly equalized recordings. While the observable transient response hasn't been defined, it is quite clear that the elements of the system must have very nearly critical damping.

The problem of determining what noise level constitutes perceptible noise involves many factors. More important are the characteristics of the noise, the characteristics of the ear, and the characteristics of the room in which the desired sounds are generated and of the room in which they are reproduced. If we consider the average residence as a place of listening, then the shaded area of Figure 2 gives the spectral masking level of the noise due to surroundings only. If we consider the filled theatre, the same average noise level will hold. In a large concert hall the noise level may be as much as 10 db lower due to the larger volume of the room. The intensity level of a sound in a room is proportional to V/T, where V the is volume of the room in cubic feet and T the reverberation time of the room. This level imposes two limiting aspects on the transmission system. First, any noises below this level will not be apparent to the listener and secondly any soft passages of the music which are below this level will not be heard. It may be seen from Figure 2, that a 60 cycle hum can have a level of 57 db and still be inaudible while a 120 cycle noise could only be 41 db and a 1,000 cycle noise only 25 db. Therefore since the lower limit of noise perceptibility depends on the frequency of the noise, the character of the noise is an important factor in determining its perceptibility.

Noise in transmission systems falls into three general categories . . . impulse noise such as ignition, static, or man-made interference; random frequency noise such as thermal hiss, and constant tone noise such as a-c hum. Impulse noise has an irregular frequency distribution with much energy concentrated in the higher frequency ranges. The apparent loudness of the noise increases with an increase in the band width of the transmission system. The peak value of such noise voltages is directly proportional to the band width and may go to such high values as to introduce new frequencies into the system due to distortion. Ran dom frequency noise has a distributio of frequencies over the whole uppe range of the transmission band. Th apparent intensity varies as the squar root of the band width of the systen with peak values of about 3.4 times th rms values. While random noise du to thermal hiss has a very even dis tribution, the random noise put out b a phonograph pickup when playin shellac records with an abrasive, ma have a definite peak frequency depend ing upon the dynamics of the picku mechanism. Constant tone noise is of course, of a specified frequency with a peak value that is seldom more than twice the rms value.

For a noise to be entirely imperceptible, the peak value should be below the threshold as established by Figure 2. No over-all figure can be given for noise level because of the variations involved. If the rms value of all noise components does not exceed a level 75 db below the peak sound level to be transmitted by a system, there is no possibility of perceptibility. In electronic devices where hum is a limiting factor of noise level, weighting should be applied in determining the noise level characteristics.

To summarize, an electro-acoustic transmission system, which will pass all frequencies uniformly from 20 to 15,000 cycles, and will have a linear response over the range of 75 db within 1% distortion, critical damping or greater and an r-s-s noise level overall of less than 75 db below peak level, will faithfully reproduce all sound waves in the electro-acoustic transducer that are applied to the acousto-electric transducer.

[To Be Continued]

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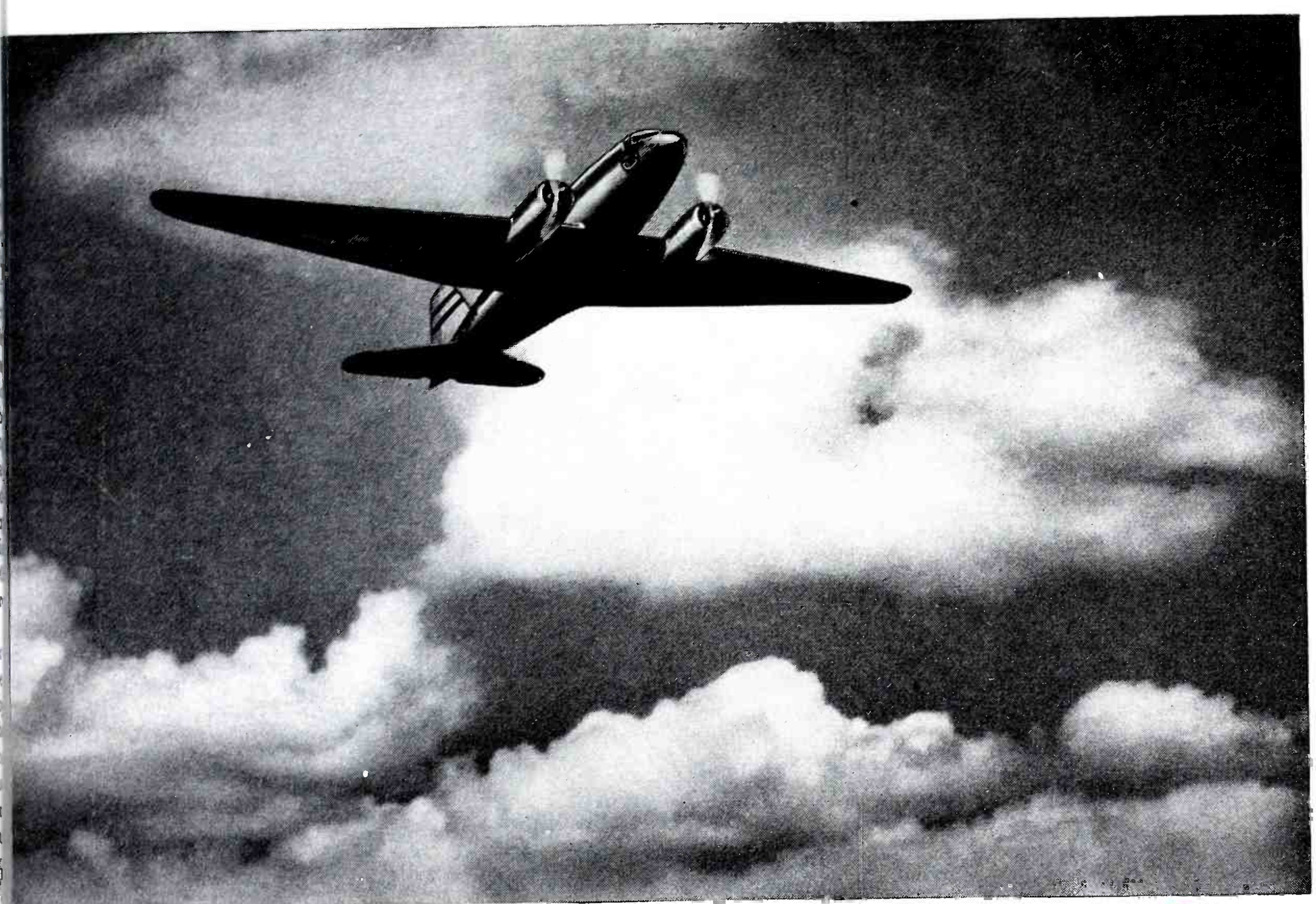

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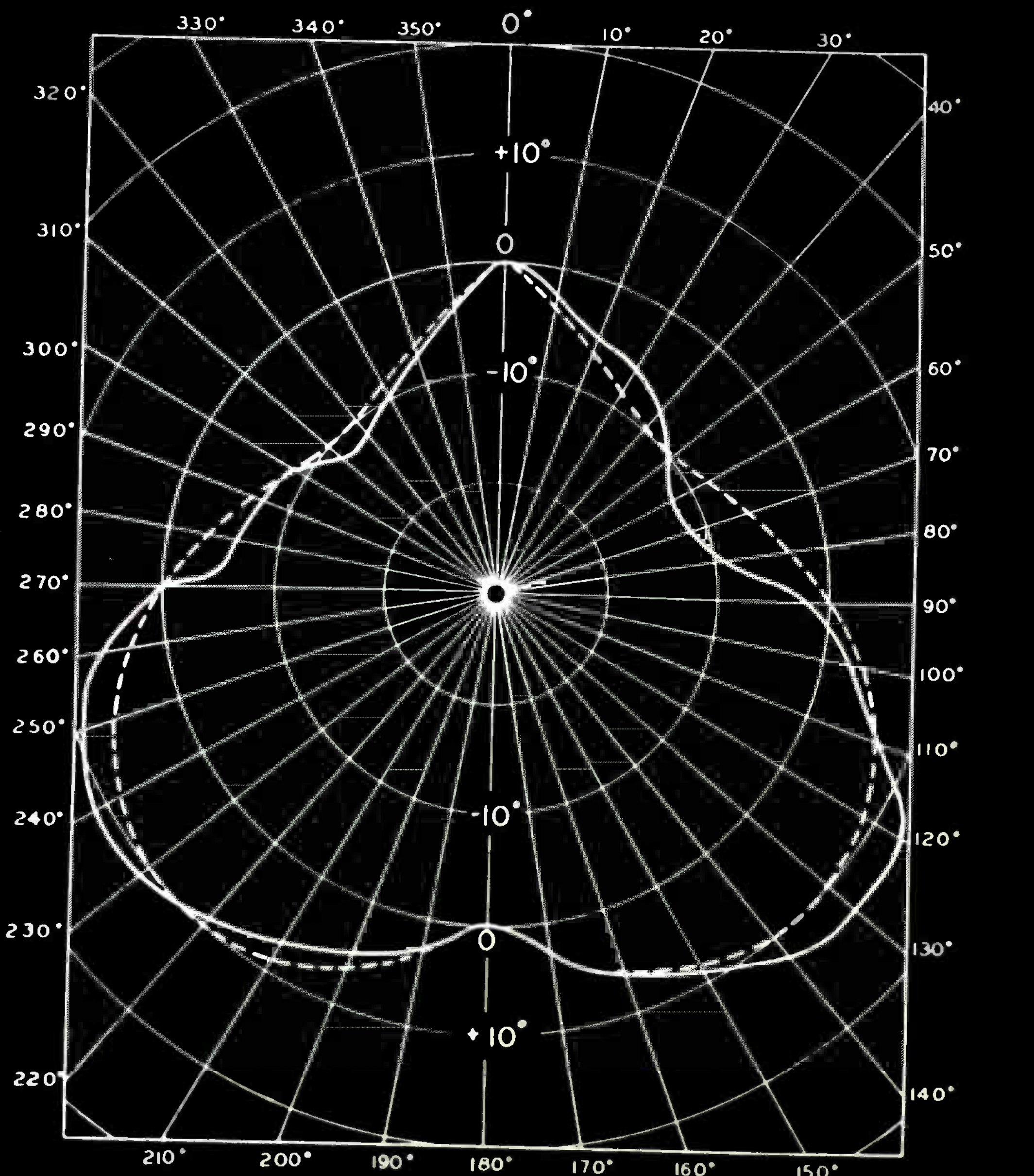

Figure 1

Polar graph showing deviation (solid line) compared with symmetrical devations (dotted line).

AIRCRAFT COMPASS CALIBRATION SYSTEMS

PART II

by CHARLES W. McKEE

Supervisor of Aircraft Radio, Eastern Air Lines, Inc.

THE fundamental principle underlining the calibration of the radio compass is to determine the deviation of the *observed radio bearing* to the *true radio bearing*. In the first section of this paper certain ground and flight procedures were described. These procedures permit the fulfilment of the essential requirement of taking radio bearings at intervals through the 360° of the azimuth. As we mentioned previously, when radio bearings are taken for the purpose of determining the magnitude of the deviation, we must establish a "radio bearing reference base." There are two logical bases namely *relative heading* and *true heading*. For the sake of convenience and since the two bases are independent and must be separately considered, it is of paramount importance that the arbitrary designation *method A* be assigned to the base that refers to the *relative heading* and *method B* to the base, *true heading*. The discussion that follows covers *Method B*.

In order to take bearings based on *true north* the radio compass dial must be set to the index point for the reading taken from the gyro compass (which was previously set to *true north*). From this it is readily seen that with all *corrected* radio bearings, when taken with the radio compass azimuth dial so set, the bearing will be the same as the true bearing from the ground reference point to the radio station. Obviously a deviation of the *observed radio bearing* from the *true radio bearing* in degrees is equal to the value of *correction*. For a summation see data in Figure 2, in which it is assumed that the *ground reference point* is directly south of the radio station on which a bearing is being taken.

From this it is evident, that during the calibration checks, the deviation is immediately known since it is equal to the difference between the *ORB* and the *TRB* to the station (*K*).

To illustrate the use of *method B*, suppose we follow a "point flight" method similar to that given in Figure 6 (part 1 of this paper). For the purposes of comparison the *TH* (true headings, column 1 of Figure 9 (part 1)), have been used with the same transmitting station location, where K = 300°. Obviously the same radio compass installation as well as the same type aircraft has been used, too. The results are given in Figure 4. Data given in column 1 (*TH*) and 3 (*ORB*) is the result of the tabulation that was taken during the radio compass calibration flight.

Data for *rel TRB* (column 2) were derived by use of the formula previously given in Figure 8 (part 1). The *rel TRB* are given in Figure 4 for reference purposes.

The *rel ORB* that appears in column 4 (when using *method B*) were derived by . . .

(*1*) When $TH \leq ORB$ then

$$ORB - TH = rel\ ORB$$

(*2*) When $TH > ORB$ then

$$TH - ORB = D$$

and

$$360 - D = rel\ ORB$$

The value for *correction* and its sign were then found by the relation and difference of *ORB* and *K* in the following way . . .

(*Continued on page* 48)

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(3) When $ORB = K$, the deviation is zero.

(4) When $ORB < K$
then
$K - ORB$ = plus correction

(5) When $ORB > K$
then
$ORB - K$ = minus correction

To obtain by interpolation the values of the *corrections* to be applied to the radio compass loop antenna compensator, the corrections (column 5) were plotted against *rel· ORB* (column 4) as shown in Figure 10 (part 1). From this curve the data, as illustrated in Figure 11 (part 1) were taken.

The characteristics of the deviation curve were described for a loop antenna symmetrically located on the aircraft, where zero deviation occurred at 0°, 90°, 180° and 270°. Loop antenna location, such as to one side of the center line, could result in a deviation at 180° and it is possible that an error could be present on the 360/0 relative heading (Figure 1). It is not advisable to tolerate abrupt deviations due to resonant interference conditions or obstruction located near the loop antenna. It is possible with certain types of aircraft construction[1] to get a reversal of the normal deviation characteristic curve where the *corrections* in the first and third quadrants are negative. This condition was referred to by C. T. Solt[2] and is equivalent to that given by an overcompensated loop antenna, where the compensation was applied by a loop purposely placed to create an induction field. It is plausible that on certain type aircraft a so-called *two-loop effect* interference is present and will exhibit this reversal.

Figure 3

Aircraft True Heading and Compass Azimuth Index Adjustment	**Radio Bearing as Read on Azimuth Dial**	**Deviation**
360/0°	360/0°	0°
10°	357°	—3°
	(. . . and correction *plus 3°* equals 360°)	
18°	355°	—5°
	(. . . and correction *plus 5°* equals 360°)	

Figure 2

An example of an adjustable steel tape type corrector unit used in compass work is shown in Figure 3. This adjustable feature is obtained by using a circular steel tape, supported at 15° intervals by adjustment screws, in such a manner, that they are used to move the tape in a radial direction thereby forming the desired contour of the cam. This contour is determined from the quadrantal correction curve. A roller is mounted on a rack gear which engages a pinion gear on the shaft, from which the angular position is transmitted to the bearing indicator. Rack and roller are mounted in a guide that allows free radial movement. This guide is attached to a circular plate that in turn is coupled to the shaft of the loop antenna. When the form of the tape is changed to an ellipse, by moving the adjustment screws in a radial direction, the radius of the roller path varies as the loop antenna is rotated. The result is a relative motion between the rack and pinion. This causes the angular position of the bearing indicator needle to be either advanced or retarded with respect to the null position of the loop antenna. The angular difference between the position of the loop antenna and the *correction* is indicated by the reading of the pointer and the scale. The null position of the loop corresponds to the position of the correction scale zero index mark on the azimuth scale. On the azimuth scale, the pointer indicates the plus or minus applied corrections to the position on the azimuth at the index. This particular type of loop antenna quadrantal error compensator is so mounted that the longitudinal line of the aircraft coincides with the 0-180° azimuth dial line.

It is usual practice and it is desirable to standardize on one type of loop antenna and bearing indicator unit. Since the corrector compensator unit has a certain relation to the assembly, it is necessary to give due consideration to whether the loop is mounted on top or the bottom of the aircraft. This condition is fully provided for by the manufacturer. For example, in the Bendix MN 31, two azimuth scales are provided. Each scale is marked and further identified by a color code as to whether it is for top or bottom installation and may be selected as needed.

[1]This is not generally encountered in present type aircraft where the wings are joined as an integral part of the fuselage and the engine nacelles are to a great extent within the wings.

[2]"Radio Direction Finding Equipment," Solt, Proceedings I.R.E., vol. 20, number 2, p. 259.

(1) TH	(2) rel TRB	(3) ORB	(4) rel ORB	(5) Corrections
300	0	300	0	0
282	18	295	13	+ 5
266	34	291	25	+ 9
245	55	288	43	+12
232	68	289.5	57.5	+10.5
217	83	295	78	+ 5
210	90	300	90	0
204	96	305	101	— 5
196	104	311	115	—11
184	116	312	128	—12
166.5	133.5	311.5	145	—11.5
147	153	309	162	— 9
120	180	300	180	0
101	199	294	193	+ 6
79	221	289	210	+11
65	235	288	223	+12
41	259	294	253	+ 6
30	270	300	270	0
22	278	307	285	— 7
10	290	310	300	—10
343	317	310	327	—10
324	336	307	343	— 7
306	354	302	356	— 2

Figure 4

In this tabulation appears the radio compass calibration data using *method B*. It will be noted that *ORB* (Column 3) is equal to a plus or minus of the 300 (*K*) by the amount of the correction; where *ORB* is that angle formed by a *true north line* and the indicated *radio bearing line* as taken from the vertex (ground reference point).

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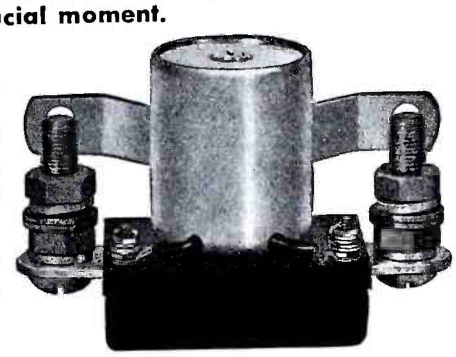

B-8 Solenoid Contactor

"B-8", six pages of Aircraft Contactors—"No. 195", on Midget and Signal Corps Relays

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TRANSMISSION LINES

(*Continued from page* 28)

Another antenna problem readily solved by use of a transmission line as a reactance involves the carrying of various circuits past the base insulator of a vertical radiator without short-circuiting the insulator for the operating frequency. The other circuits may carry either lower frequency (such as tower lighting circuits), higher frequency (such as u-h-f circuits to auxiliary antennas), or circuits at the same frequency (such as sampling circuits from pickup loops which are usually mounted at the point of maximum current on a tower).

A transmission line, a quarter-wave long, with the far end shorted offers an approximately infinite impedance at the near end. One member of the transmission line may be a copper tube in which are enclosed the other circuits. If only one other circuit, a coaxial cable itself, is needed, the outside of that cable forms the conductor for the transmission line trap, while the inside of the cable carries the auxiliary circuit.

One convenient place to install this quarter wave line is up, inside the tower. In this type of an installation, the insulated portion of the cable is just a quarter wave long, and thus there is theoretically no effect on the tower performance at the lower frequency when it is insulated or connected to ground at the lower end. In practice the length may vary greatly from a quarter-wave to an appreciable power loss. It will however have an effect when grounded, similar to connecting a reactance between the base of the tower and ground. The antenna resistance and antenna tuning adjustments will be different with the line grounded or ungrounded.

Another variation of the preceding case is obtained by running the quarter-wave isolating line horizontally, parallel with the ground, with the ground as one of the conductors.

This interesting method is used in feeding power to an f-m antenna on top of the standard broadcast antennas at WSBT. In this installation, the f-m coaxial cable is 1⅜" in diameter. In order to reduce radiation and heat losses in the quarter-wave section of transmission line, a ground return is not relied upon. Four wires are carried parallel to the coaxial cable and equally spaced, in the usual five-wire transmission line construction. These wires are grounded at both ends and at each supporting pole.

VIDEO AMPLIFIER

(*Continued from page* 21)

or minus tolerance over a greater range of frequencies, by making β slightly smaller than β_1. While this is found to be the case, it is not a simple matter to specify a definite optimum value, for the reason that the permissible size of the peak must depend upon the number of stages which are to be operated in cascade. In this connection, it should be remembered that the overall gain of a multi-stage amplifier is the product of the individual stage gains. Still another consideration is the fact that the size and shape of the peak varies as *m* is varied. As *m* is increased in size the peak becomes broader and less pronounced.

The effect of making β slightly smaller than β_1 may be observed by trying an experiment which seems to suggest itself. Suppose we arbitrarily make the response unity at the nominal cutoff point. Inserting these conditions in the gain equation, we find

$$\beta = \sqrt{m^2 + 2m - 1} \qquad (21)$$

In order to illustrate the response characteristics for large and small values of *m*, all of the accompanying curves have been plotted for two values of *m*, namely, 2 and 10. The curves for the uncompensated amplifier are shown also for purposes of comparison. Figure 6 shows the amplitude and phase response when $\beta = \beta_1$ (Equation 20). Figure 7 shows the same curves when the gain is arbitrarily adjusted to unit value at $T/T_g = 1$.

The location of a peak value in the response curve (when $\beta < \beta_1$) can be predicted by maximizing equation 18 and solving for $(T/T_g)_{max}$. This is found to be

$$\left(\frac{T}{T_g}\right)^2_{max} = \left(\frac{\beta}{1+m}\right)^2 \left\{ \sqrt{\frac{[(m+1)^2-1][(m+1)^2-\beta^2]}{\beta^2}} - 1 \right\} \qquad (22)$$

The magnitude of $|H|$ at the peak is then found by substituting the value of $(T/T_g)_{max}$ back in Equation 18. It is noted in Figure 7 that the peak is higher when $m = 2$, being 1.012 as compared with only 1.001 when $m = 10$. Comparing these curves with those of Figure 6, it can be seen that the greater improvement has resulted for the case where $m = 2$. By permitting a small peak in the response,

(*Continued on page* 72)

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NEWS BRIEFS OF THE MONTH . . .

CENTRAL SUPPLY SOURCE FOR RADIO LABORATORIES

Laboratories working on radio problems for the Army or Navy will be able to secure vital parts, not quickly available in commercial firms, from a central source of supply. The Electronic Research Supply Agencies have been formed by the Defense Supplies Corporation at the request of the armed services, the Office of Scientific Research and Development and the WPB, to supply these vital parts.

Laboratories will, of course, not be compelled to place their orders with or through these new agencies. However, approved laboratory orders which may be filled by the agency can be placed there or can be channeled to the agency through commercial distributors. Thus, distributors can place with the agencies for the account of laboratories, those portions of orders which they themselves are not in a position to fill.

The agency will operate without profit, and will be managed by an executive committee on which Army, Navy and other government agencies will be represented. Offices and stockrooms of this new expediting unit will be located in New York City.

* * *

NAVY NEEDS ENGINEERS

The rapid expansion in Navy surface vessels, submarines, and aircraft has created several hundred additional billets for officers trained in electrical engineering. They are needed to serve in engineering work in connection with ultra-high frequency electronic apparatus.

To qualify for these new posts, it is necessary to hold a degree in electrical engineering and have practiced in the field in engineering since graduation, or to have majored in physics, mathematics, or other branches of engineering and acquired a sound working knowledge of alternating-current circuits and electronics.

Men commissioned for this electronic work will be given the Navy's three-months' course in ultra-high frequencies at either Harvard University or Bowdoin College, followed by an additional three-months' laboratory course at Massachusetts Institute of Technology. Upon graduation, these officers will be assigned to responsible engineering positions involving research, design, instruction or maintenance of the Navy's ultra-high frequency equipment.

Qualified engineers are urged to apply for a commission at the nearest Office of Naval Officer Procurement. These offices are located in principal cities throughout the United States.

* * *

COAXIAL CABLE DATA RELEASED BY ANDREW

The Victor J. Andrew Company, 363 East 75th Street, Chicago, has just released a sixteen-page folder describing coaxial cables and fittings.

The technical characteristics of coaxial cables, factors affecting their choice, and data on cable terminals, are among the highlight features of the new release.

Copies of the catalog are available to engineers writing in on their letterheads.

A. E. SNYDER NOW WITH NORTH AMERICAN PHILIPS

A. E. Snyder has been appointed manager of the industrial electronics division of the North American Philips Company.

* * *

SMPE SEMI-ANNUAL MEETING IN N. Y.

The fifty-third semi-annual meeting of the Society of Motion Picture Engineers will be held at the Hotel Pennsylvania from May fourth to sixth. No program of speakers has been announced as yet.

* * *

SECOND PLANT OPENED BY CLAROSTAT

The Clarostat Manufacturing Company, Brooklyn, New York, recently opened a second plant in the same city. The new plant will be devoted exclusively to assemblies, while the original plant will be devoted to fabrication of parts and windings. The engineering and general offices will remain in the original plant.

* * *

AIRCRAFT ACCESSORIES WINS "E"

The Burbank division of the Aircraft Accessories Corporation was recently awarded the Army-Navy "E." Among those attending were (from left to right in view below): Lieutenant Colonel Maurice J. Joyce, U. S. Army Air Forces, Chief of the Production Section, Western Procurement Division; Captain W. C. Barker, U. S. Navy, Professor of Naval Science and Tactics, University of California at Los Angeles; Randolph C. Walker, president of Aircraft Accessories Corporation; Lieutenant Colonel G. M. Bates, U. S. Army Air Forces; Captain J. G. Church, U. S. Navy, Resident Inspector of Naval Materiel at Consolidated Steel Corporation, and Timothy E. Colvin, executive vice-president in charge of the Burbank division.

PRICE CEILING ON FIXED CAPACITORS

Fixed capacitors of the type or size used for military radio equipment were placed by the Office of Price Administration under a ceiling reflecting current price levels.

Until now this type of condenser was exempt from price control in order to enable increased production in the face of very unstable conditions and to allow the rapidly expanding industry an opportunity to stabilize production costs. The exemption was effective until April 1, 1943, and until that time the War Department and the Navy Department had agreed to exercise control over prices for the product.

The new ceilings reflect increases in cost caused by the rapid expansion of the industry and increased labor rates since March 31, 1942. In effect, fixed capacitors which are of the type used for military purposes, may be sold at list prices effective on April 1, 1943, less discounts, allowances, and other deductions in effect on that date.

For manufacturers who had no established list prices, the action provides a formula for determining prices based on the manufacturer's own pricing method and labor and materials costs in effect on April 1, 1943. The amendment requires all manufacturers covered by the action to file reports of their maximum prices as determined under the regulation with OPA before April 30.

* * *

J. M. ALLEN WITH ERIE

J. M. Allen has been appointed works manager of the Erie Resistor Corp., Erie, Penn. Mr. Allen was formerly with RCA, Bloomington, Indiana.

* * *

ARMY-NAVY WHITE STAR AWARDED TO HALLICRAFTERS

The Hallicrafters, Chicago, Ill., have won a white star for their "E" pennant.

* * *

WAR-WORK PROFIT PLAN AT UNITED ELECTRONICS

The United Electronics Company, 42 Spring Street, Newark, N. J., has instituted a trust fund under which its employees will share in war-work profits, to be paid at the end of the war. The plan announced by R. H. Amberg, president of the company, is believed to be the first of its kind in the East. It was designed to forestall turnover in personnel, halt absenteeism and to provide a ***nest egg*** for employees in the post-war period, according to Mr. Amberg.

To be eligible, an employee must remain with the company until after the war, unless called to military service. The amount to be paid each employee will vary in proportion to his salary. In event of his death the money will go to his heirs.

* * *

RCA ISSUES WAR EDITION TUBE GUIDE

A revised seventy-two page edition of the RCA Guide for Transmitting Tubes, designed especially for radio engineers

(*Continued on page* 54)

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IN laminated plastic materials, as in so many others, the new conditions imposed by war have led to intensified research which has developed new products and new qualities that are certain to be valuable after the war.

Formica, with the assistance of customers and suppliers, has had a share in this progress. Some of the new things: laminated plastic name plates, glass cloth base insulating material to serve some of the uses of ceramics; arc resistant insulation, "Pregwood", a light, strong, impregnated wood that serves many mechanical uses, and fluorescent instrument panels visible in the dark.

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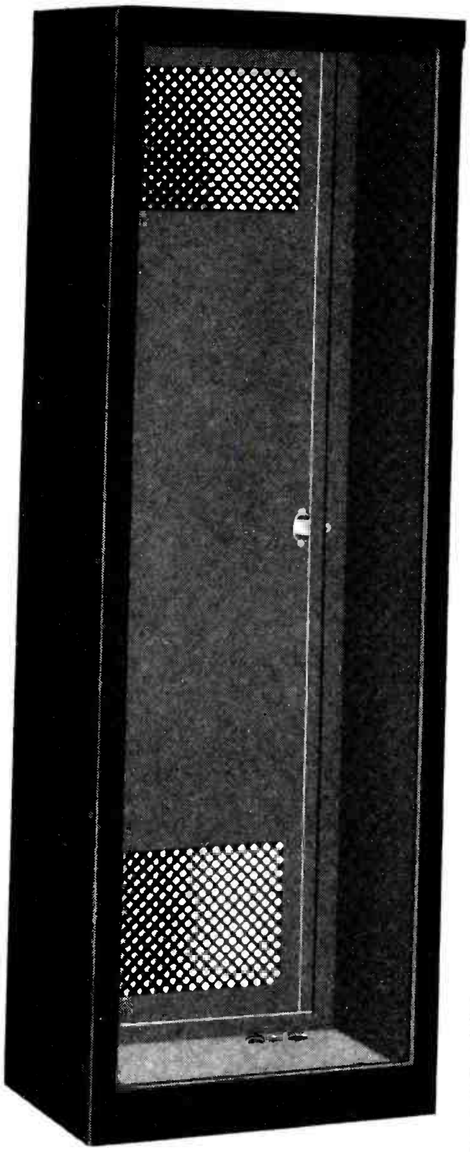

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NEWS BRIEFS

(*Continued from page 52*)

and technicians in the armed services and war industries, has just been published by RCA and is available through all RCA power tube distributors.

* * *

SOLAR WINNER OF "E" AWARD

The Solar Manufacturing Corporation, Bayonne, New Jersey, has been awarded the Army-Navy "E" for excellence in war production.

The presentation was made in the grand ballroom of the Waldorf-Astoria Hotel in New York City, before an assemblage of employees, executives and Army and Navy officials.

* * *

SCOTT MAKES UNIQUE DEMONSTRATION

The new Scott receiver used on U. S. merchant ships and employing Thordarson transformers, was recently demonstrated to the employees of the Thordarson plant by E. H. Scott, inventor of the receiver.

Thordarson employees displayed an unusual interest in this demonstration, revealing as it did, the importance of their work in the war effort.

* * *

CAPACITOR AND FILTER DATA IN AEROVOX BULLETINS

In the August issue of the Aerovox Research Worker, capacitors in control circuits are effectively analyzed. And in the September-October issues, design data for m-derived type filters are provided. Formulae derivatives and circuits for low-pass, high-pass, band-pass and band-suppression filters are also presented.

* * *

LATHE MAINTENANCE BOOKLET

A new booklet describing lathe maintenance, lathe tests and adjustments, has been issued by the South Bend Lathe Works, South Bend, Indiana.

Covered in this twenty-eight-page booklet are motor and power supply testing; shortening, splicing and belt adjusting; testing and adjusting spindle bearings, etc. The booklet known as bulletin H4 is available, free, from the manufacturer.

* * *

NATIONAL UNION PROMOTES J. J. CLUNE

J. J. Clune, who has been serving as assistant sales manager of National Union Radio Corporation, Newark, New Jersey, has been assigned to the post of director of war service. Mr. Clune has been with National Union since 1931.

* * *

ELECTRICAL TAPE BOOKLET

A twelve-page catalog describing glasflex acetate cloth, acetate film and other types

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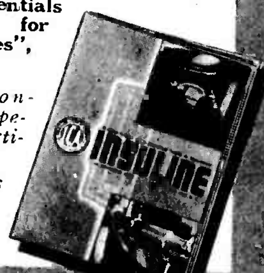

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of electrical adhesive tapes has been published by the Insulation Manufacturers Corporation, 565 West Washington Boulevard, Chicago, Illinois. The bulletin contains electrical and mechanical specifications, as well as information on weights, sizes and prices.

Folders describing varnished tubing products made by the suflex process have also been released by Insulation Manufacturers Corporation. Both bulletins are available, gratis, from the manufacturer.

* * *

COLLOIDAL GRAPHITE BULLETIN

A twelve-page bulletin, No. 430-W, describing the physical and chemical properties of "dag" colloidal graphite, has been published by the Acheson Colloids Corporation, Port Huron, Michigan.

The many uses of colloidal graphite for a variety of applications are analyzed in this bulletin. Copies are available gratis.

* * *

ZIEGLER ON WPB ADVISORY COMMITTEE

John M. Ziegler of the Crystal Products Company, Kansas City, Missouri, has been appointed to the National Quartz Crystal Industries Advisory Committee.

Mr. Ziegler, who is a former member of the RCA Research Laboratories, taught piezo electrical applications courses through the government ESMWT program sponsored by the University of Kansas.

* * *

SHAKEPROOF SCREW BULLETIN

A recent release by Shakeproof, Inc., 2501 North Keeler Avenue, Chicago, showing the microscopic qualities of the No. 1 Shakeproof thread cutting screw, has recently been released. This bulletin is available to those who write in on their business letterheads. Available, too, is a kit of assorted testing samples.

* * *

FOX OF WGAR WITH RFC

Robert A. Fox of the WGAR engineering staff is on a leave of absence, in charge of a crew of engineers working in South America under the supervision of a branch of the RFC.

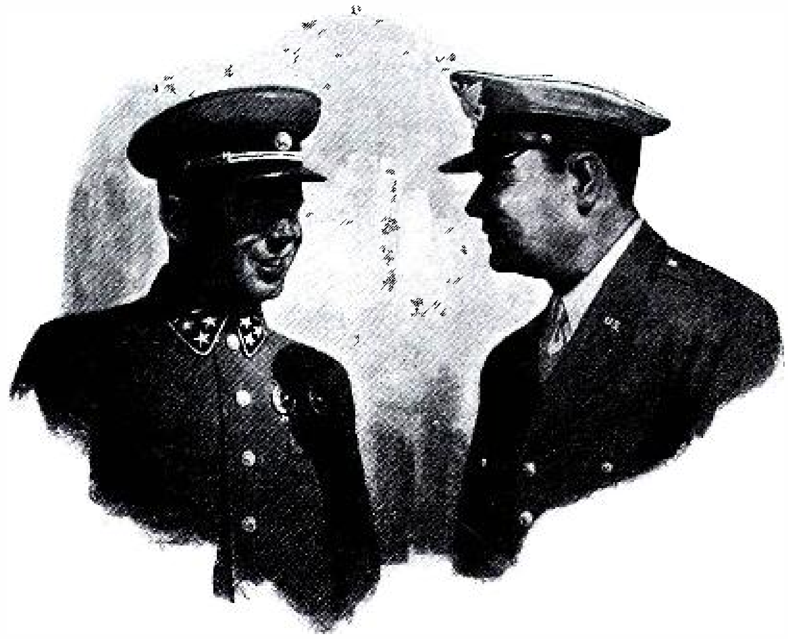
* * *

VIC MUCHER SERVING WPB

Vic Mucher of the Clarostat Manufacturing Company, Incorporated, 285-7 North Sixth Street, Brooklyn, New York,

(*Continued on page* 62)

Vic Mucher



ОНИ ХОРОШИ ОНИ ЗАМЕЧАТЕЛЬНЫ

Don't speak Russian? Then let us translate the words of a Russian General to an American War Correspondent:

"THEY'RE GOOD . . . THEY'RE EXCELLENT!"

You see, the Correspondent had just remarked upon the number of "Connecticut" field telephones in use by the famed Cossack Cavalry. , , , Like many an American industry, our reputation for know-how rests today on the performance of our products in the service of the United Nations, all around the world. , , , When we can again freely solicit your patronage, there will be no testimonial to which we shall point with greater pride than the commendation of the fighting Russians.

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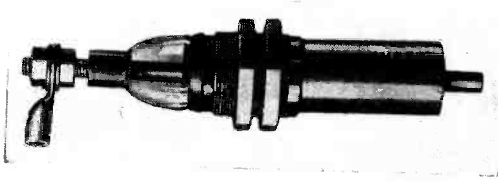
THE INDUSTRY OFFERS . . .

ANDREW GAS-TIGHT TERMINAL

A terminal that is said to be 100% gas-tight has been developed for use on radio coaxial cables by Victor J. Andrew Co., 363 East 75th Street, Chicago, Ill. The unusual seal is obtained by fusion of glass to metal. The metal alloy has a suitable coefficient of expansion.



* * *

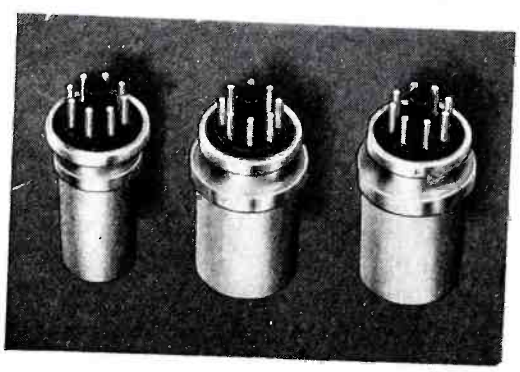
STURTEVANT SPRING TESTER

For testing compression springs in sizes to 2½" diameter and 7" in length the P. A. Sturtevant Co., Addison, Illinois, have developed a new tool. This new tool is said not only to make it possible to rapidly measure the recoil pressure of a spring when compressed to any predetermined length, but to accurately match sets of springs (as valve springs for internal combustion engines).

* * *

SPRAGUE PLUG-IN DRY ELECTROLYTICS

A new plug-in type of dry electrolytic condenser, for the elimination of low frequency ripple (2—100 cycles) has been developed by Sprague Specialties Co., North Adams, Mass. It is small in size and lightweight, and thus can be used effectively in portable and mobile units.

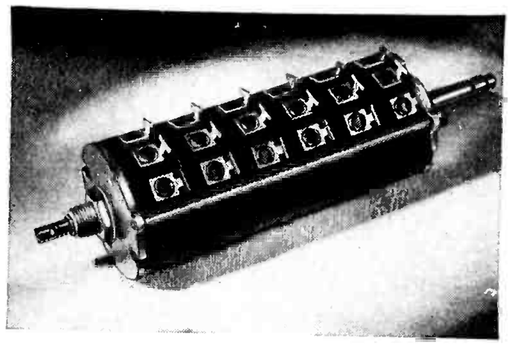


CLAROSTAT TANDEM CONTROLS

A plurality of circuits can be controlled by the single shaft of the 42 series control, developed by Clarostat Mfg. Co., Inc., 285-7 N. 6th St., Brooklyn, N. Y.

The new design of case for each unit permits the nesting and locking of all units into a compact stack. The metal end discs and tie rods hold the cases together and provide further rigidity. The single shaft passes through and locks with each rotor in the stack. All units of the control of course pass through the same degree of rotation as the single shaft is rotated. Individual units can be of any standard resistance, taper, taps and hop-offs to meet individual circuit requirements.

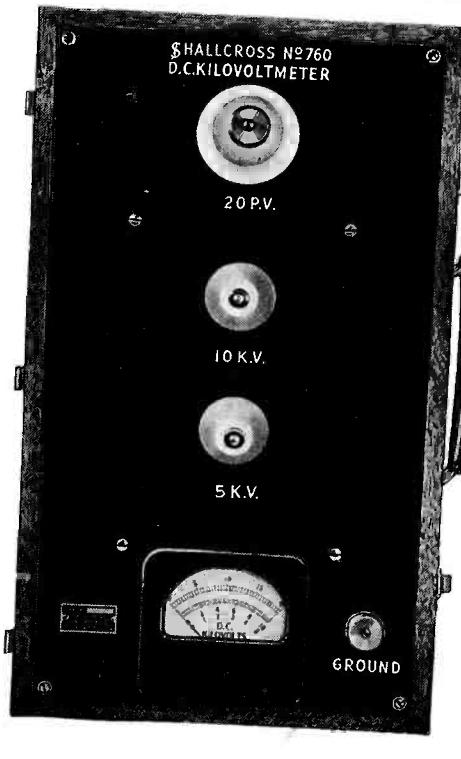
These controls are necessarily made on special order only, since the number of sections and the values vary from one application to another. Units with as many as 20 sections are being produced for critical applications.



* * *

SHALLCROSS KILOVOLTMETER

To measure 5, 10 and 20 kilovolts, with a current consumption of one milliampere, a kilovoltmeter has been designed and manufactured by The Shallcross Manufacturing Company, Collingdale, Pa. The multiplier unit contains Shallcross wire wound resistors mounted on ceramic insulators. Its accuracy is said to be plus or minus 2% at full scale.



INVISIBLE WATERPROOF CHEMICAL

An invisible "raincoat" which can be formed on cloth, paper and many other materials by exposing them to chemical vapors from a new compound known as *dri-film,* thereby making them water-repellent, has been developed in General Electric's research laboratory at Schenectady, N. Y., by Dr. Winton I. Patnode.

One of its most important uses so far is the treatment of ceramic insulators for radio equipment. It is said to be about nine times more effective than the wax used at present as a water repellent, and its results are permanent.

Dri-film is a clear liquid composed of various chemicals which vaporize at a temperature below 100 C. Articles to be treated are exposed, in a closed cabinet, to the vapors for a few minutes. Then they are taken out and, if necessary, are exposed to ammonia vapor. This is to neutralize corrosive acids which may collect during treatment. With this treatment the insulators are said to maintain their high insulating properties, even under adverse conditions.

* * *

REPEATERS' USE EXTENDED

In less than half a working day, more military telephone repeaters are being made now, than the total number used in France during World War I, according to R. E. Smith, head of the wire transmission department of Federal Telephone and Radio Corporation, Newark.

The use of repeaters in stepping up transmitted telephone messages was just beginning when World War I started. Therefore not many were available for military use.

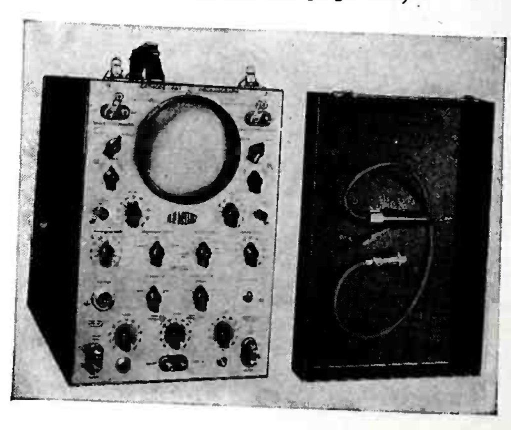
Repeaters used in the current conflict are approximately one tenth the size of the early ones, and, of course, are easily portable. The cabinet of the first repeaters did not contain the current supply source; a whole barrel of dry cell batteries was supplied for use with each of them. Today's military repeaters, even the smallest, have self-contained current supplies.

* * *

DUMONT 5-INCH OSCILLOGRAPH

A larger screen size, together with the inclusion of a Z-axis amplifier to modulate the beam with any signal applied to its input terminals or with a return trace blanking impulse produced by the linear-time-base generator, are features available with the new type 241 Du Mont

(*Continued on page* 58)



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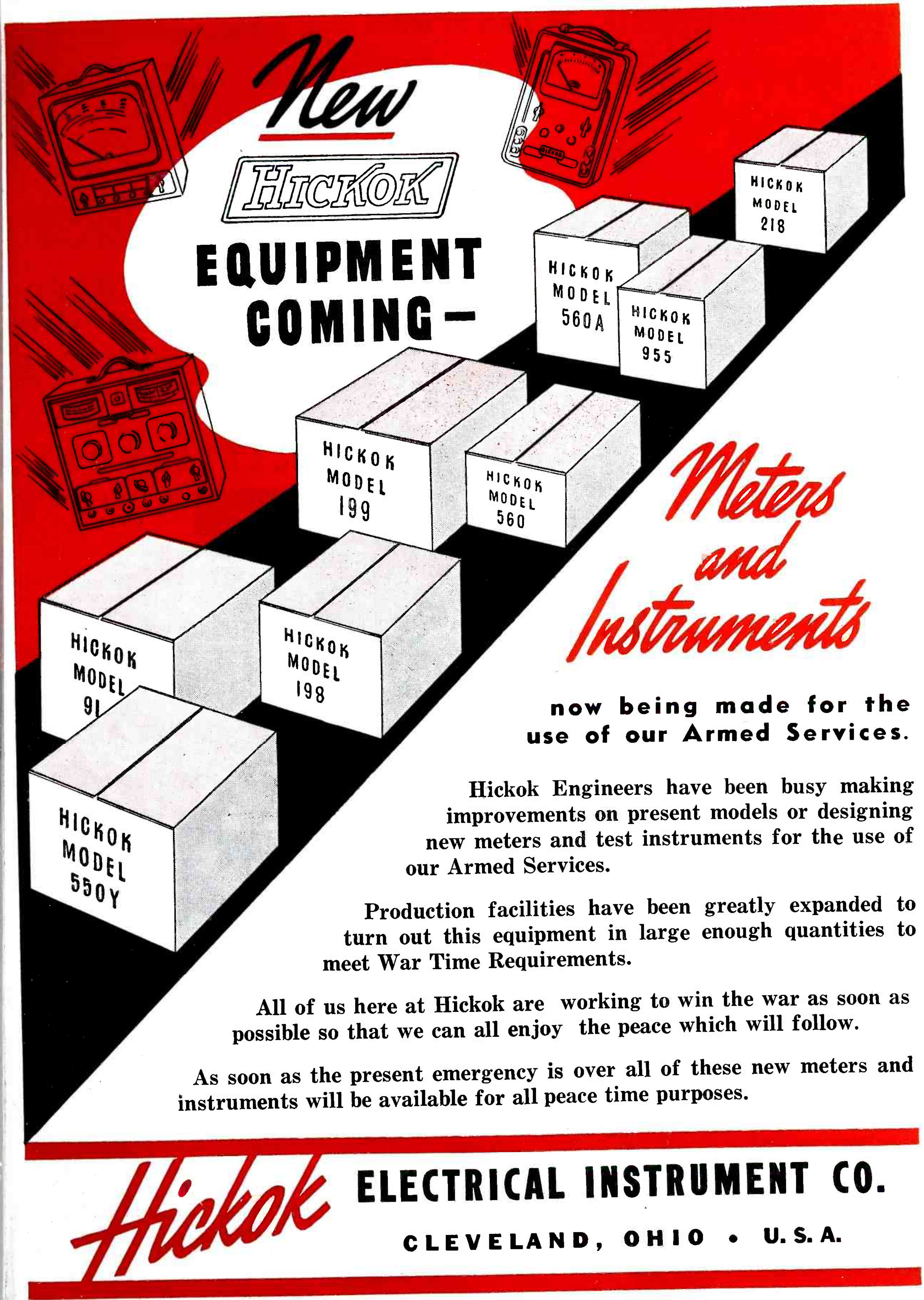

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Space Saver

OIL CAPACITOR with the NEW double terminals

- Something *new* has been added to this long-popular Aerovox "10" series Hyvol capacitor. Note the *double terminals* — spaced on the stepped bakelite threaded terminal post. This means the can of this handy inverted-screw-mounting capacitor is now insulated or "floating." No longer is an insulator washer required when non-grounded mounting on a metal chassis is desired.

  As handy to install as the usual metal-can electrolytic. And just as compact. 600 to 1500 v. D.C.W. .5 to 4 mfd.

- If you do not already have a copy of our Transmitting Capacitor Catalog in your working library, write today on your business stationery, for your copy.

AEROVOX CORPORATION
NEW BEDFORD, MASS., U. S. A.
In Canada: AEROVOX CANADA LTD., Hamilton, Ont.
EXPORT: 100 Varick St., N. Y., Cable 'ARLAB'

THE INDUSTRY OFFERS . . . .

(*Continued from page* 56)

5-inch cathode-ray oscillograph.

This new unit is said to have a uniform Y-axis or vertical deflection response from 20 cps to 2 megacycles. The X-axis or horizontal deflection amplifier is said to have a uniform characteristic from 10 cps to 100 kilocycles. Provision is made to connect signals directly with the deflection plates when frequencies to be observed are beyond the useful limits of the amplifiers.

A removable test probe, held inside the cover by clips, consists of a compensated 10:1 attenuator mounted in an insulated probe and supplied with a 3-foot length of coaxial cable and connector. This feature permits connections to relatively high impedance circuits without serious loading, while minimizing stray pickup.

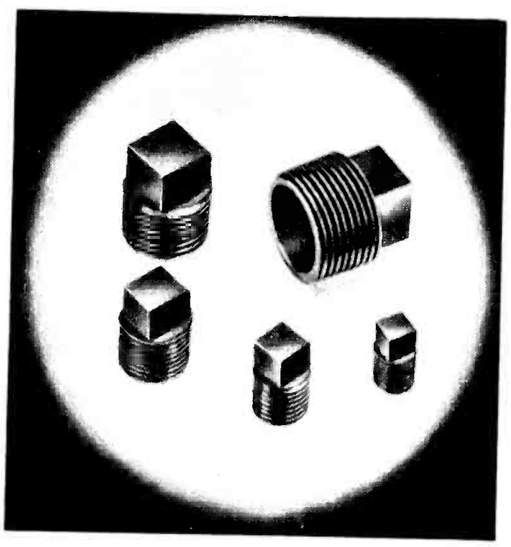
Self-contained, it operates directly off the 60-cycle 115-volt a-c line. Size is 17½" high, 10¾" wide, 21" deep.

* * *

PLASTIC PIPE SEAL PLUGS

Plastic pipe seal plugs for protection for threads of units in storage or in transit, are being made by American Molded Products Company, 1644 North Honore St., Chicago, Ill.

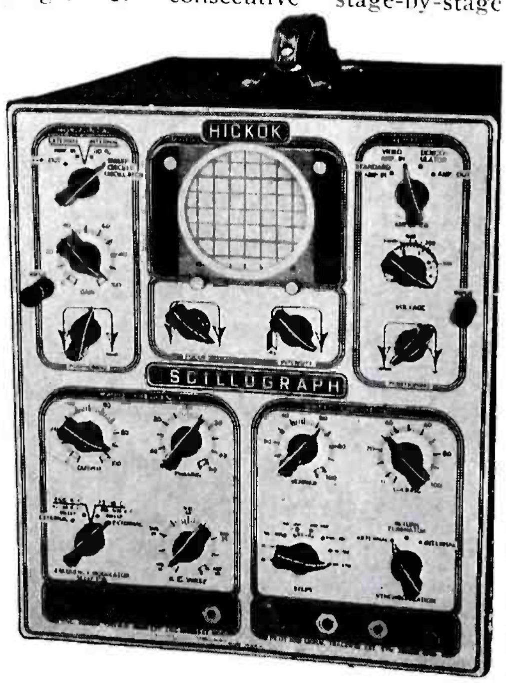
The molded plastic pipe seals are available in five sizes . . . ⅜", ½", ⅝", ¾" and 1".

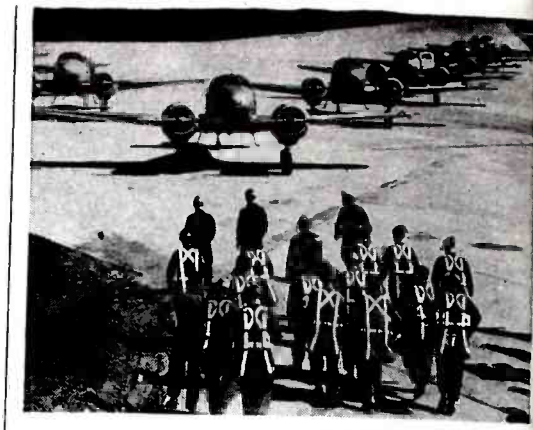


* * *

HICKOK OSCILLOGRAPH

To check both r-f and i-f stages for single or consecutive stage-by-stage





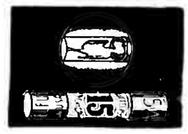
U S AIR CORPS PHOTO

Dependable Help!

FOR MEN CONCENTRATED ON AN OBJECTIVE

Elements *mechanically depolarized* against vibration from any direction are found in

ANTI-VIBRATION LITTELFUSES



Mechanically Depolarized Fuse Element

Twist of fuse elements at 90° holds against tests 10 times worst field conditions. Other exclusive Littelfuse features: "Gooseneck" taking up contraction and expansion; Littelfuse Locked Cap Assembly, sealing against "perspiration."

LITTELFUSE INC. 4751 Ravenswood Ave. CHICAGO, ILL.

MATHEMATICS FOR RADIO AND COMMUNICATION

by GEORGE F. MAEDEL, A.B., E.E.
Chief Instructor, RCA Institutes

To master the technicalities of radio — to read engineering literature intelligently — you must have the mathematical groundwork covered by these absorbing books prepared for home study. Book I (314 pp.) covers the algebra, arithmetic, and geometry; Book II (329 pp.) covers the advanced algebra, trigonometry, and complex numbers necessary to read technical books and articles on radio.

MAEDEL PUBLISHING HOUSE Room 105
593 East 38th Street, Brooklyn, New York

Send me MATHEMATICS FOR RADIO AND COMMUNICATIONS as checked below. I enclose payment therefor with the understanding that I may return the book(s) within 5 days in good condition and my money will be refunded.

Name ....

Address ....

....

☐ Book I at $3.75 plus 6c postage
☐ Book II at $4.00 plus 6c postage
☐ Books I and II at $7.75 postage prepaid

Foreign and Canadian prices 25c per volume higher.

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trouble shooting, from antenna post to receiver in frequency modulated, amplitude modulated and television receivers, the Hickok Electrical Instrument Co., 10514 Dupont Ave., Cleveland, Ohio, have developed the RFO-5 oscillograph.

It has a self-contained wide band (100 to 900 kc sweep) frequency modulated oscillator (basic frequency 23 mc) for frequency modulated and television servicing. It has a narrow band (10-30 kc sweep) frequency modulated oscillator (basic frequency 1000 kc) for visual alignment on amplitude modulated receivers, demodulators, etc. Another feature is . . . a wide band frequency modulated oscillator that can be modulated from an external frequency source such as a phonograph pickup, microphone or audio frequency oscillator. The unit also has a self-contained mixer circuit, demodulator, video amplifier, signal tracer, visual a-c vacuum-tube voltmeter 0.2 to 1000 volts, calibrated screen, phasing control.

* * *

PHOTOCOPY MACHINE

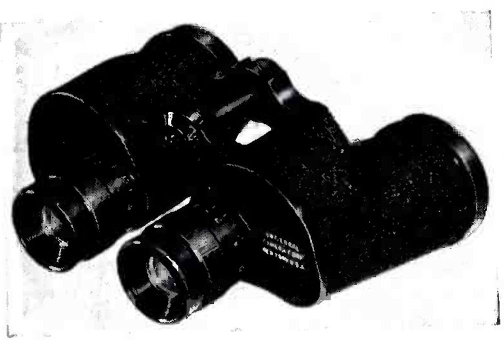
A photocopy machine that is said to have almost no working parts and presents practically no repair or maintenance problems has been developed by American Photocopy Equipment Co., 2849 N. Clark Street, Chicago, Ill. It can be set up on any desk or table and no darkroom is needed. The machine, known as the Apéco, makes same size copies of anything up to 18x22 inches.

* * *

BINOCULAR PRODUCTION INCREASED

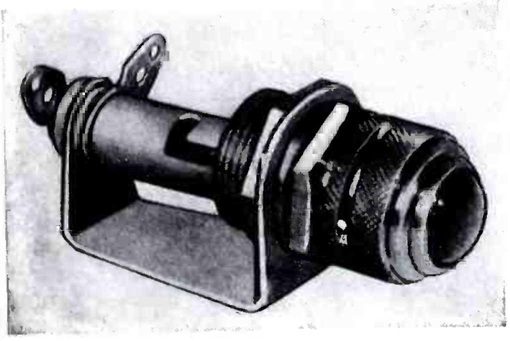
Prior to the outbreak of war, precision binoculars were made almost entirely by hand. Since then, military demands have required accelerated production of these vital devices. Thus, methods and equipment had to be developed, whereby unskilled workers could perform operations previously done only by optical workers of long experience. This unusual procedure was developed at Universal Camera Co., New York.

* * *

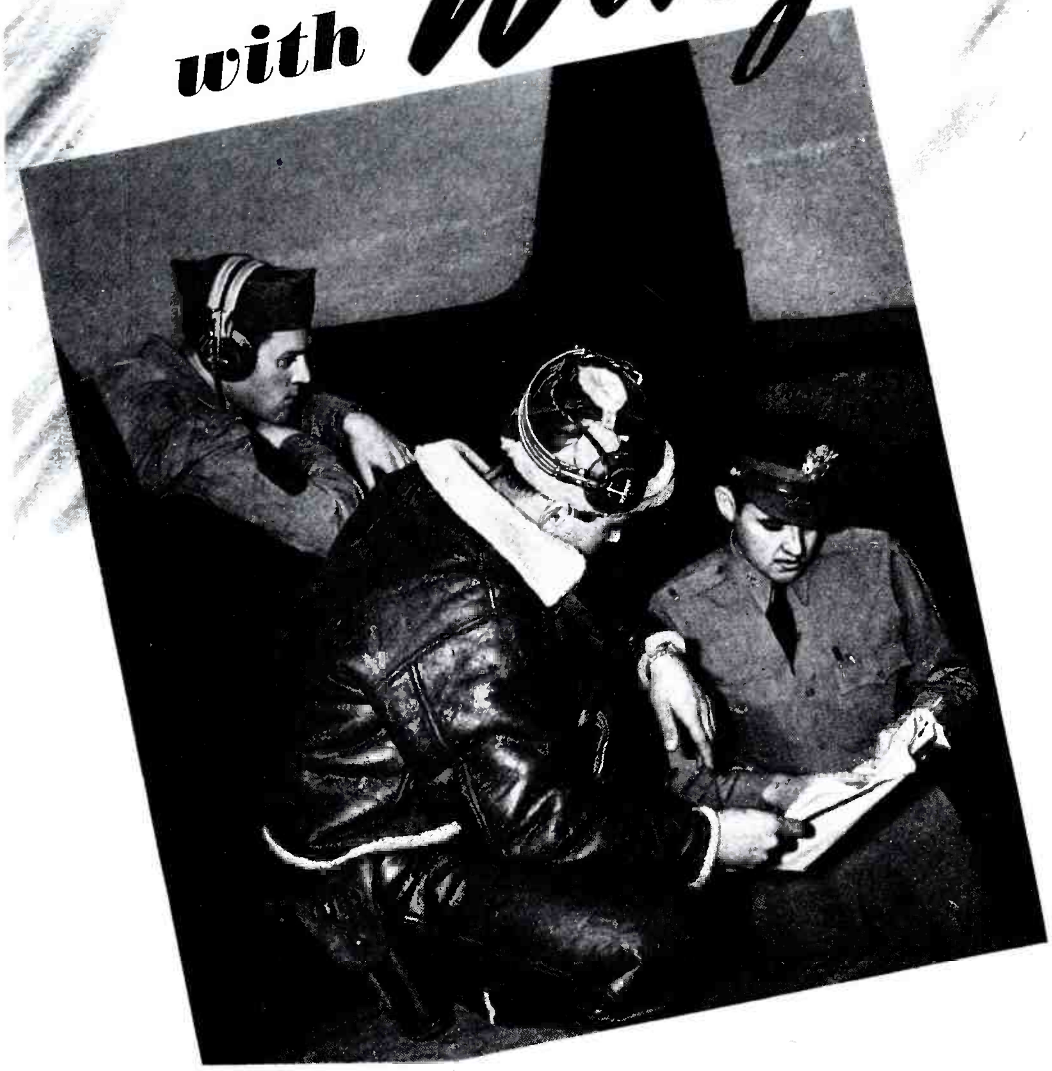


GOTHARD SHUTTER TYPE PILOT LIGHT

A shutter type pilot light, particularly suited to aircraft, marine, signal and similar applications where various intensities of light are desired under constantly changing conditions is now being made by the Gothard Manufacturing Co., 1300 N. 9th Street, Springfield, Illinois.



The Men with Wings



. . . depend on Communications

CONSOLIDATED RADIO is proud to be making headphones for the men who fly the skies of the world for the United Nations. The lives of our men—indeed Victory itself—depend upon instant, uninterrupted intercommunications, and CONSOLIDATED RADIO headphones are "delivering the goods."

Engineered for complete dependability, CONSOLIDATED RADIO headphones are withstanding the most gruelling demands of battle . . . be it in the tropics, the arctic or in the stratosphere.

Consolidated Radio's Modern Mass Production Methods Can Supply Signal Corps and Other Headphone Units in Quantities to Contractors

CONSOLIDATED
Radio Products Company
350 WEST ERIE STREET • CHICAGO, ILLINOIS

SPECIALISTS IN MAGNETIC AND ELECTRONIC DEVICES

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VETERAN WIRELESS OPERATORS ASSOCIATION NEWS

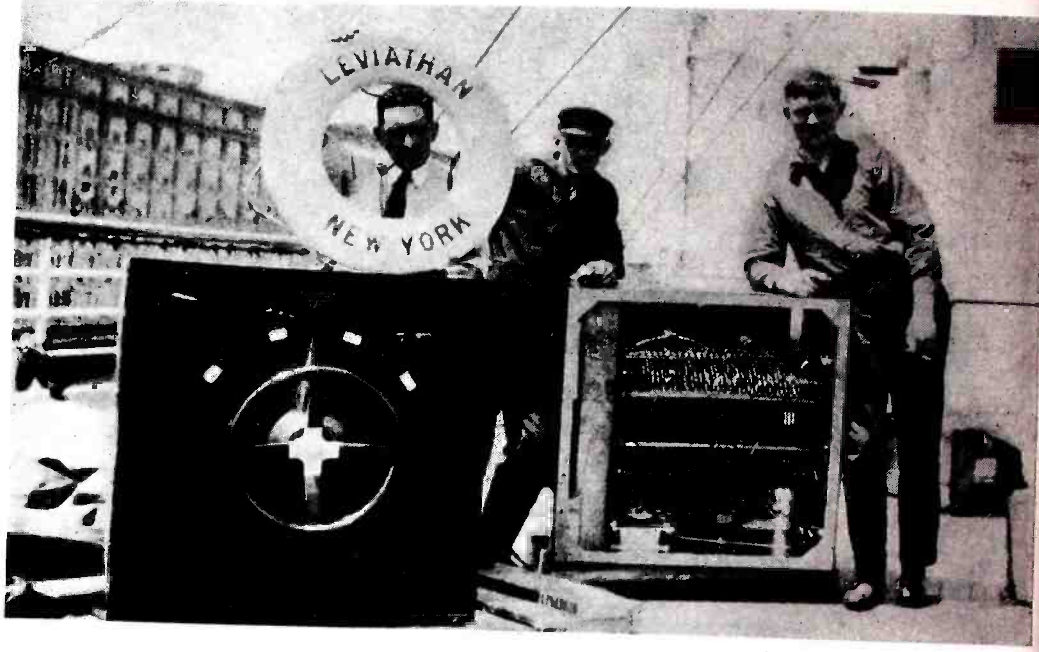
W. J. McGONIGLE, President   RCA BUILDING, 30 Rockefeller Plaza, New York, N. Y.   GEORGE H. CLARK, Secretary

RADIO lost another pioneer in the death of Henry Kasner, for thirty years associated with the Radio Corporation of America and its predecessor, the Marconi Wireless Telegraph Company of America. Kasner, who was 53 years old, unmarried, died on February 14 at the Brooklyn Jewish Hospital after a brief illness. He played a prominent part in the early days of wireless communication, and had a hand in the phenomenal growth of radio as a major industry.

Born in London, England, on May 14, 1889, Kasner joined the United Wireless Telegraph Company, an American concern, on April 11, 1911, as a wireless operator. That company was taken over by the American Marconi Company just one year later, as a result of litigation over patent rights. Kasner in this way became a member of the firm which he served faithfully for three decades.

One of his first assignments was to assist in the construction of the Marconi trans-Pacific station at Honolulu, Hawaii. Just before the entrance of the United States into World War I, he carried on similar work at New Brunswick, N. J., where a trans-Atlantic station, duplicate of its Island sister, was under construction. The engineering knowledge which he gained while working on these huge, intricate installations was of great service to him in his later work.

At the outbreak of World War I, Kasner joined the U. S. Navy, and was detailed to the Great Lakes districts. Later he served as operator on the S. S. Leviathan. At the close of the war he again entered the service of the American Marconi Company, being assigned to the maintenance, repair and installation department, better known as *MRI*. Here his work was very satisfactory. In the words of one of his superior officers of that time, he was "an efficient installer, and an impartial inspector." During this period of his service he installed several "firsts"; such as the first commercial tube transmitter on the Atlantic and Pacific coasts and on the Great Lakes, and the first commercial radiophone transmitter on the S. S. America.



Henry P. Kasner (left) and the radio equipment on the S. S. Leviathan.

On the occasion of RCA's first entrance into broadcasting, when the Dempsey-Carpentier boxing match was broadcast on July 2, 1921 over a station licensed for a short period at the Lackawanna Terminal, Hoboken, N. J., Kasner was one of the installing engineers. When the apparatus was transferred later to RCA's first permanent broadcasting station, at Roselle Park, N. J., he again took part in the installation. He was announcer and engineer of a portable broadcasting station at RCA's first display in a radio show, at the 71 Regiment Armory, New York City, in September, 1921.

By that time, radio receiving sets were being installed in many homes in America, and with their installation came many complaints of noises caused by leaky power lines. RCA developed instruments for locating the actual sources of these disturbances, as well as for minimizing their effects at the receiver. For months Kasner toured the eastern part of the United States, lecturing on the elimination of these disturbing arcs from improper power lines, and also assisting in locating them.

In the fall of 1922 a travelling radio show was set up by RCA, to tour the entire country in an effort to acquaint the public more directly with the company and its varied products. Kasner had an important part in handling this show for nine years, setting up and dismantling the displays, and acting as lecturer and also as announcer for the show's portable broadcasting transmitter. His duties took him to all states in the Union except eight, appearing to a total audience of over 23 million people.

In 1931 this show was discontinued, display activities being thenceforth handled on a local basis by the RCA Victor Company. Mr. Kasner was placed in charge of this work, and served in that capacity until shortly after the outbreak of World War II. He was then assigned to the field procurement division of the general purchasing department of the RCA Victor Division of RCA, and stationed in New York. He was chiefly engaged in procuring vital war materials for manufacturing purposes. In this work he conferred with hundreds of manufacturers of materials. His work as an expediter was outstanding. And much of RCA's "Beat the Promise" results at the Camden factories could

(Continued on page 66)

The electron tube is the dynamic force of the future. Today, National Union engineers are doing their part in tube research to harness the "dynamite" that will usher in the Age of Electronics. Their laboratories are in the thick of the battle of production. In their achievements for war, National Union engineers are creating new applications of the electronic tube, gaining new knowledge and experience for its role in the industrial life of the future. When Victory is won, they will be ready to create electronic applications for *your* production processes.

Transmitting Tubes • Cathode Ray Tubes • Receiving Tubes • Special Purpose Tubes • Condensers • Volume Controls • Photo Electric Cells • Exciter Lamps • Panel Lamps • Flashlight Bulbs

NATIONAL UNION RADIO CORPORATION
Newark, N. J. **Lansdale, Pa.**

SPOT WELDING

Infinite care and precision in delicate assemblies are a tradition of National Union manufacture. It takes rigid and expert training as well as skilled and nimble fingers to perform this Spot Welding operation . . . and to enable it to pass the "eagle-eye" test that makes it fit for use in National Union Electronic Tubes.

NATIONAL UNION ELECTRONIC TUBES

NEWS BRIEFS

(*Continued from page 55*)

has been appointed as a consultant to the Radio and Radar division of WPB.

* * *

RADIOSONDE DISCUSSED AT RADIO CLUB OF AMERICA

Leo A. Weiss, instrument design supervisor of Simmonds Aerocessories, Incorporated, recently discussed the radiosonde before members and guests of the Radio Club of America at Columbia University. Mr. Weiss covered the development of this interesting device and told of its importance today in weather forecasting. He also told of the use of this device in the study of transmission and reception.

* * *

MILITARY EQUIPMENT VIEWED BY SYNDER EMPLOYEES

So that the employees of the Snyder Manufacturing Company, Philadelphia, Pennsylvania, could judge the importance of their work in the war effort, the Industrial Relations Offices of the Army and Navy recently provided an exhibit of military equipment in their plant.

Equipment shown included field artillery pieces, jeep, motorcycle, amphibian car and a walkie talkie using a Snyder-built antenna.

* * *

INSULATING MATERIAL GUIDE BOOK

A forty-three page booklet covering a discussion of a variety of insulating materials has just been published by Mitchell-Rand Insulation Company, Incorporated, 51 Murray Street, New York City.

Tables, diagrams and other pertinent data are included in this booklet, which is free for the asking.

* * *

PHILCO'S STORAGE BATTERY DIVISION WINS WHITE STAR

A white star has been added to the Army-Navy "E" flag of the Storage Battery Division of the Philco Corporation. Ceremonies marking the addition of this star were held at the plant.

* * *

CALLITE CATALOG ISSUE

A thirty-six page catalog discussing the design, manufacture and application of electrical contacts of silver, platinum, tungsten, molybdenum and many other metals and alloys, has just been published by the Callite-Tungsten Corporation, Union City, New Jersey.

The catalog, known as No. 152, contains design and production data, with typical illustrations of use. Copies of the catalog are available direct from the manufacturer.

TEST with TELRAD

18A FREQUENCY METER

(Secondary Frequency Standard)

- ★ Transmits EXACT frequency carriers simultaneously every 10 KC and every 100 KC; also marker carriers every 1000 KC between 100 kilocycles and 60 megacycles. 1000 KC carriers usuable from 1 megacycle to 150 megacycles, accuracy .05%.
- ★ The ideal Crystal Controlled Signal Generator for I.F. or R.F. Alignment.
- ★ Checks Factory or Field Test Oscillator Accuracy.
- ★ Checks Receiver and Transmitter Calibration. Unexcelled for setting Electron-coupled Oscillator Transmitter Frequency to Close Tolerance, substituting for Spot Frequency Crystals.
- ★ Modulation "ON-OFF" Switch.
- ★ No calibration or tuning charts required.
- ★ Size 5" x 8" x 8½". Weight 12 lbs.

Operates on 115, 130, 150, 220, and 250 volt 25 to 60 cycle A.C. Supplied complete with Bliley 100 and 1000 K.C. duo-frequency crystal and one 6V6 oscillator, one 6N7 harmonic generator, one 6L7 harmonic amplifier and one 6X5 rectifier

Manufactured by

FRED E. GARNER CO.
51 E. Ohio St. Chicago, Ill.

* * *

C. M. SLACK PROMOTED BY WESTINGHOUSE

Dr. Charles M. Slack has been appointed assistant director of research of the Westinghouse Lamp division, Bloomfield, New Jersey. He is well known for his work in tube design, particularly in association with X-rays.

* * *

"E" AWARDED TO FARNSWORTH

The Farnsworth Television & Radio Corporation, Fort Wayne, Indiana, recently received the Army-Navy "E" for outstanding production.

The pennant was received by E. A. Nicholas, president of Farnsworth, from Lieutenant Colonel Kenneth D. Johnson, U. S. Signal Corps.

* * *

FORMICA PROMOTES EXECUTIVES

Five new vice-presidents were elected by the Formica Insulation Company, Cincinnati, Ohio, recently. They are: J. Roger White, in charge of sales and advertising; George H. Clark, in charge of engineering; R. W. Lytle, in charge of special engineering; Ellsworth G. Williams, in charge of manufacturing, and W. J. Gebhart, in charge of finances and accounting.

* * *

WAGENER NOW CHIEF ENGINEER OF H-K

Winfield G. Wagener has just been promoted to chief engineer of Heintz and Kaufman, Ltd., South San Francisco,

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alifornia. Mr. Wagener has been with -K for five years during which time s efforts were devoted to the development of vacuum tube applications for h-f circuits.

* * *

R. IRVING LANGMUIR HONORED

Dr. Irving Langmuir, associate director the General Electric research laboratory, has been elected to honorary membership in the Institute of Metals, London.

The Institute of Metals, which is international in character, has only two honorary members at present. One is Professor C. A. F. Benedicks, director of e Metallographic Institute, Stockholm, weden, and the other is Dr. A. M. ortevin, who was professor of metallurgy in the Central School of Arts and Ianufactures in Paris, before German ccupation.

* * *

VALL CHART ISSUED BY RAWLPLUG

ready reference wall chart covering xpansion bolt and screw anchor dimensional data, has been issued by the Rawllug Company, Inc., 98 Lafayette Street, New York City. The chart, which is 4"x20", will be sent to any user of xpansion bolts, provided the request is ent on business letterhead. A description of the chart appears in a broadside which is also available for the asking.

* * *

AEROVOX OPENS NEW PLANT

A second plant devoted exclusively to the manufacture of mica capacitors, was recently opened by Aerovox Corporation, New Bedford, Massachusetts. The new plant is located close by the Aerovox plant.

* * *

JOHN ROSEVEAR NEW WESTINGHOUSE PLANT MANAGER

The new West Virginia works of the Westinghouse Lamp Division will be managed by John Rosevear. Mr. Rosevear has been with Westinghouse for twenty-one years, having joined them in 1922 as an electrical draftsman.

* * *

NATIONAL UNION COMMENDED FOR RED CROSS EFFORT

A certificate for conspicuous achievement was recently awarded to the National Union Radio Corporation, Newark, New Jersey for their 100% participation in the war fund drive of the American Red Cross.

* * *

PHILIPS QUARTZ CRYSTAL FIXTURE BULLETIN

Accessories and fixtures extending the use of Philips X-Ray Quartz Analysis apparatus are described in a new engineering bulletin 201.

The bulletin also illustrates and describes crystal blank holders, edge correction holders, etc. Copies of the bulletin can be obtained by writing to the Philips Metalix Corporation, 419 Fourth Avenue, New York City.

* * *

BRIDGEPORT WORKS OF G. E. WINS "E"

The Army-Navy "E" award was recently made to the Bridgeport division of General Electric. Attending the ceremonies were Raymond E. Baldwin, Governor of Connecticut; Jasper McLevy, Mayor of Bridgport; Dr. Wei Taoming, Chinese Ambassador to the United States; Gerard Swope, president of Gen-

(*Continued on page 64*)

HOW TO ORDER PRESTO RECORDING DISCS, NEEDLES AND REPLACEMENT PARTS

and get prompt delivery

1. Place your order with your distributor for the discs and needles you will need during the next 90 days. The distributor will stock them and deliver at your convenience. He will need your orders to determine his stock requirements which he must estimate 90 days in advance.

2. Apply to your purchase order the AA2X preference rating which you have received under the revision of War Production Board order P-133 dated February 4, 1943, part 3037.

3. In ordering replacement parts or equipment renewals give your distributor the serial number of the equipment to be repaired or replaced and the part number as shown in your instruction book. Apply the AA2X priority to your order.

★ ★ ★

Buy Presto products through leading radio distributors or any branch office of the Graybar Electric Company.

PRESTO
RECORDING CORP.
242 WEST 55th ST. N.Y.

In Other Cities, Phone . . . ATLANTA, Jack. 4372 • BOSTON, Bel. 4510 CHICAGO, Har. 4240 • CLEVELAND, Me. 1565 • DALLAS, 37093 • DENVER, Ch. 4277 • DETROIT, Univ. 1-0180 • HOLLYWOOD, Hil. 9133 • KANSAS CITY, Vic. 4631 • MINNEAPOLIS, Atlantic 4216 • MONTREAL, Mar. 6368 TORONTO, Hud. 0333 • PHILADELPHIA, Penny. 0542 • ROCHESTER, Cul. 5548 • SAN FRANCISCO, Su. 8854 • SEATTLE, Sen. 2560 WASHINGTON, D. C., Shep. 4003—Dist. 1640

World's Largest Manufacturers of Instantaneous Sound Recording Equipment and Discs

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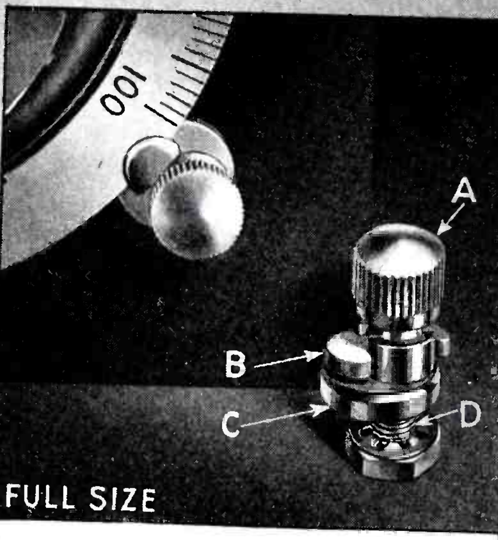
Designed for

Application



The No. 10050 Dial Lock

Designed for application! Compact, easy to mount, positive in action, does not alter dial setting in operation! Rotation of knob "A" depresses finger "B" which firmly pinches dial between "B" and "C" *without imparting any rotary motion to Dial.* Single hole mounted by means of shank "D". Made of brass —Standard finish Nickel.

JAMES MILLEN
MFG. CO., INC.

MAIN OFFICE AND FACTORY
MALDEN
MASSACHUSETTS

NEWS BRIEFS

(*Continued from page 63*)

eral Electric, and many other notables of the Army, Navy and foreign services.

In an outdoor mass meeting, during which the presentation was made, thousands of men and women workers assembled and pledged to maintain production at all costs.

* * *

AMORY HOUGHTON OF CORNING GLASS ON SHORT WAVE BROADCASTS

Amory Houghton, chairman of the board of Corning Glass Works, recently broadcast a message to our armed services in the Western and Eastern theaters of operation, over station WBOS.

This broadcast, designed as a weekly report to our fighting forces from American industry, was the twelfth in a series of weekly broadcasts presented by the National Broadcasting Company and the National Association of Manufacturers in cooperation with the War Department.

Mr. Houghton pointed out the tremendous strides that American industries have made with glass for an amazing variety of applications. He sent greetings to the many men of Corning Glass who are now in the fighting services.

* * *

ROLLER-SMITH APPOINTS R. M. SMITH AS CHIEF ENGINEER

Roy M. Smith who joined the Roller-Smith Company, Bethlehem, Pennsylvania, in August, 1942, as assistant chief engineer, has been appointed chief engineer succeeding J. D. Wood, resigned.

* * *

LINK WINS "E"

The Link Radio Corporation, New York City, recently won the coveted Army-Navy "E" production, award for excellence in war manufacture.

The "E" pennant was awarded to Fred M. Link by Major William S. Marks, Jr., Signal Corps, U. S. A. Lieutenant Clement R. Hoopes, U. S. N. R. presented the "E" pins, the initial presentation being made to William Tense, the oldest Link employee.

The entire presentation ceremonies were broadcast over WOR and the Mutual Network with Tom Slater as commentator.

The tenth anniversity of the Link Radio Corporation was also celebrated at these presentation ceremonies.

* * *

STROMBERG-CARLSON CHANGES NAME

The Stromberg-Carlson Telephone Manufacturing Company will hereafter be known as the Stromberg-Carlson Company.

Dr. Ray H. Manson, vice-president and general manager announced the change of name.

* * *

F. B. WARREN NOW RCA COMMUNICATIONS GENERAL COUNSEL

Frank B. Warren, formerly assistant counsel of the Federal Communications Commission, has been appointed general counsel of RCA Communications, Inc.

* * *

IGNITRON CONTACTORS BULLETIN

Ignitron contactors for power switching of a-c resistance welding machines are described in a new, illustrated eight-page bulletin (GEA-3058B) issued by the General Electric Company.

The publication also illustrates three standard electronic contactors, the 150-, the 300-, and the 1,200-ampere sizes. In addition, it lists the four principal ways in which these contactors may be used for the control of a-c resistance welders.

* * *

VOICE PAGING SYSTEM DATA

In a four-page bulletin, released by Bell Sound Systems, Inc., 1183 Essex Avenue, Columbus, Ohio, unit-designed voice paging systems are described.

The systems utilize remotely located amplifiers, each amplifier capable of operating up to fifteen speakers. The system can also be used for transmission of recordings.

* * *

McNAMEE AND ROEMER ELECTED TO I. T. & T. BOARD

Admiral Luke McNamee, president of Mackay Radio and Telegraph Company, and Henry C. Roemer, vice-president and comptroller of Federal Telephone and Radio Corporation, have been elected to the Boards of Directors of International Telephone and Telegraph Corporation.

* * *

L. C. ATHEY NOW INTERNATIONAL PRODUCTS VICE-PRESIDENT

Lyman C. Athey has been elected vice-president of the International Products Corporation, Baltimore, manufacturers of low-loss inorganic plastics. Mr. Athey was formerly associated with the Porcelain Enamel and Manufacturing Company, Baltimore, as director of research.

* * *

LT. GEN. KNUDSEN VISITS HALLICRAFTERS

Lt. Gen. William S. Knudsen, director of production for the United States Army, recently made a surprise visit to the Hallicrafter plant in Chicago, Ill.

He was accompanied by Lt. Colonel Walter Bain, Lt. Colonel John M. Niehaus, Major W. M. Collins, Major J. Stein, Captain John Cavanaugh, Captain Bert Roesch and Lt. R. H. Collins.

Ray Durst, Herb Hartley and Joe Frendries welcomed the General and escorted him on an inspection trip through the plant.

* * *

STANCOR TRANSFORMER CATALOG

A new catalog, No. 140, has just been issued by the Standard Transformer Company, Chicago, Illinois.

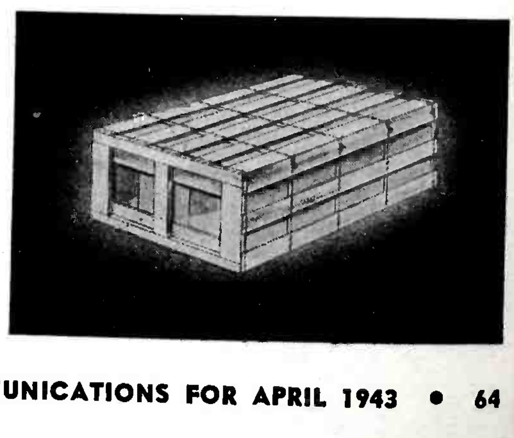
The new catalog contains detailed specifications on the Stancor line of transformers and chokes for replacement and general purposes. Featured is a handy classified and numerical index.

The catalog is available from the manufacturer, gratis.

* * *

AIRCRAFT PARTS SHIPPED IN NEW TYPE CASE

Wirebound wood crates made by the General Box Company, Chicago, are being used to ship aircraft radio parts within the bounds of the United States.



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These boxes are light weight and designed to consume a minimum amount of unpacking time.

* * *

PLASTICS HIGHLIGHT BROCHURE

An eight-page leaflet describing the mechanical and electrical properties of lucite, plastacele, pyralin, nylon and butacite, has just been released by the plastics department of E. I. Du Pont de Nemours & Co., Inc., Arlington, New Jersey.

* * *

SICKLES WINS "E"

The F. W. Sickles Company, Springfield, Massachusetts, has won the Army-Navy "E".

The pennant award was made by Lieutenant Colonel Kenneth D. Johnson, Signal Corps, U. S. A., to Roy F. Sickles, president of the F. W. Sickles Company. The proceedings were broadcast over local stations.

* * *

CERAMIC DEVELOPED FOR TUBE BASES

A new material, prestite, recently developed by Westinghouse is now being used by Heintz and Kaufman, Ltd., South San Francisco, California, for the bases of high frequency tubes.

The new material, which is in the porcelain family, is said to be made of raw materials that are found in abundance in this country and are not on restricted priority lists. The new ceramic is said to possess the electrical and mechanical strength of wet process porcelain with the moulding qualities of dry process porcelain. It is formed under heavy hydraulic pressure.

* * *

RCP INSTRUMENT CATALOG

The latest additions to the RCP line of instruments appear in a new bulletin, No. 127, just released by the Radio Products Company, Inc., 127 West 26th Street, New York City.

The bulletin describes the new model 703 signal generator, 419 master multitester in three models, 416 and 418 pocket multitesters and 446A a-c/d-c multitester.

Copies of the bulletin are available from the manufacturer.

* * *

J. L. FOUCH ELECTED UNIVERSAL MICROPHONE PRESIDENT

James L. Fouch, former vice-president of the Universal Microphone Corporation, Inglewood, California, has been elected president. He succeeds James R. Fouch, who becomes chairman of the board. Cecil L. Sly has been promoted to the position of vice-president and treasurer, while Durwood Allen becomes the new secretary.

* * *

HAYDU TUBE EQUIPMENT CATALOG

A forty-page loose-leaf catalog describing vacuum pumps, air blowers, bombarders, gas boosters, ribbon burners, hand torches, gas and air burner tips, crossfires and economizers used in the manufacture of fluorescent tubes has just been published by Haydu Brothers, Plainfield, New Jersey.

The catalog, replete with illustrations and technical data, will prove extremely useful to those engaged in the manufacture or production of the general type of tubes as well as fluorescent types. Copies of the catalog are available to engineers who send in their requests on letterhead stationery.

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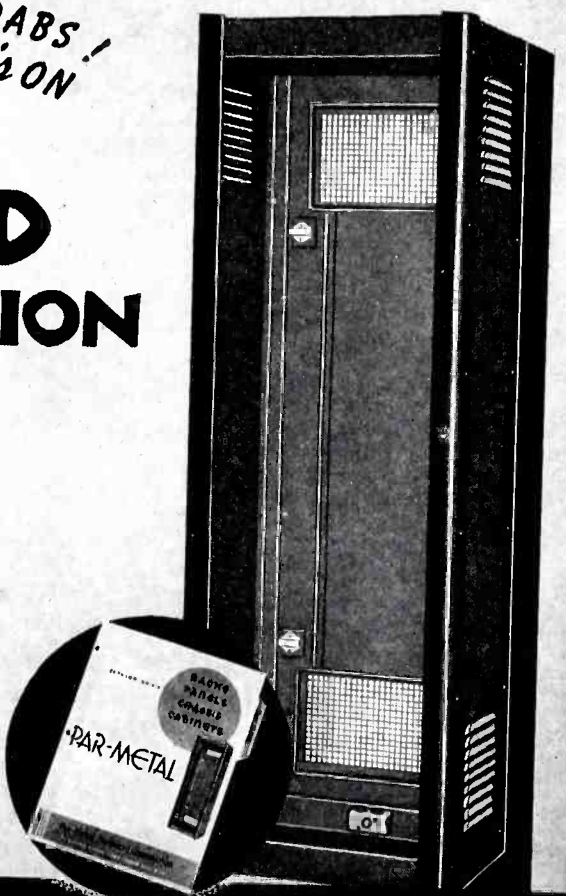

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VWOA NEWS

(*Continued from page* 60)

be directly traced back to his efforts. He was busy in this work when suddenly stricken with the illness which in a very short time proved fatal.

At his burial services, held at the Park West Memorial Chapel, New York City, many of his old time associates, from the various companies with which he had worked, came to say their final goodbye. Represented, too, were many from the Veteran Wireless Operators Association, of which he had been a member almost from its inception.

Personals

GLAD to hear that Dr. Lee de Forest, *Father of Radio* and honorary president of our Association continues to be active daily in his seventieth year. . . . Hal Styles, chairman, and Mack Schaefer, organized a fine party for the Los Angeles chapter of our Association recently. Hal and Mack have been doing a fine job of keeping alive the VWOA spirit in the West. . . . Also heard from Gilson Willets, one of our founders and a charter member and chairman of the San Francisco chapter. . . . Our sincere congratulations to Rear Admiral Joseph Redman recently returned to the position of Director of Naval Communications, after a period of sea duty in the Pacific War zone. Admiral Redman occupied the position formerly and was confirmed for advancement from Captain to Rear Admiral upon his return. . . . All good wishes of our Association go to Captain Carl F. Holden, who was relieved as Director of Naval Communications by Admiral Redman. We understand Captain Holden is to take command of a new battleship; we know he will give the Axis a run for their money. . . . We hear that Commander Pierre Boucheron is again overseas. . . . We are grateful for the following letter from C. J. Pannil, upon whom our Association conferred a Marconi Memorial Medal of Achievement, as a pioneer operator, "This is the first opportunity I have had to thank you and the Committee on Awards for your recognition of my achievements in the radio field as a pioneer. To say to you that I appreciate this honor more than words can express, is only speaking mildly of my real feelings in the matter, and I again desire to assure you of my very deep appreciation for this acknowledgment by the Veteran Wireless Operators Association. I shall always prize the medal and have fond recollections of my many very dear associates in the Association."

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U-H-F S-R RECEIVER

(*Continued from page* 15)

capacitive reactance of the grid condenser should be low enough to act as a virtual short across the grid resistor for the resonant frequency. And the value of capacitance should be at least 10 times the input capacitance of the tube, so that the a-c voltage developed across the resonant grid circuit will appear as a very small drop across the grid condenser. The limiting factor for the value of resistance will be the relationship of resistance to capacity for the audio component, as explained in the previous paper (March, 1943).

The best quench frequency, therefore, will have limits other than those previously established for best quench frequency. This is one of the drawbacks of self-quenching. Another, is the fact that the d-c bias developed across r-c is not controllable. This factor is important for the audio relationship, since a high bias will seriously reduce the gain for audio frequencies.

Percentage of Feedback

With the optimum values for resistance and capacity obtained by the above method, the next factor to be determined is the percentage of feedback. Best results from the standpoint of both sensitivity and selectivity can be obtained by using the least amount of feedback necessary to maintain stable operation. The percentage of feedback will be determined by the amount of positive resistance in the grid circuit. However, the positive resistance in the grid circuit is a function of positive resistance normally present and the resistance introduced by coupling to the previous stage. For greater selectivity, it is necessary to keep the value of coupling below the critical point. To increase sensitivity it is necessary to approach the point of critical coupling.

The condition here is somewhat complicated, since part of the energy being fed back from the plate of the detector to the grid, will be transferred to the coupling element of the previous stage. This will reduce the positive resistance in that circuit and increase its effective Q. For this reason it is unnecessary to have the plate circuit of the r-f stage tuned.

Self-Quenching

Comparatively good results may be obtained with self-quenching, but the control effected by the dual-oscillator

(*Continued on page* 68)

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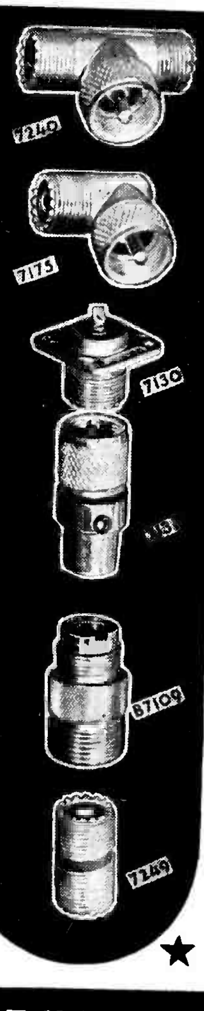

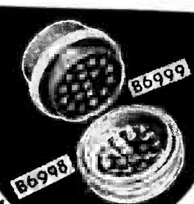

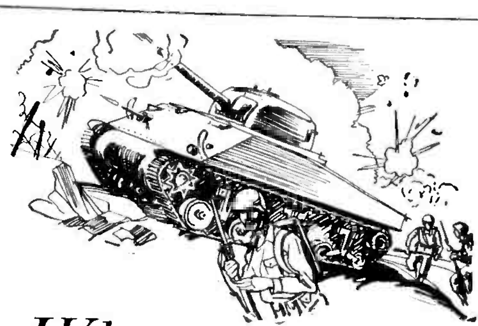

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(*Continued from page* 67)

quench detector marks it as a superior circuit.

The second type of self-quenching tried is shown in Figure 5. This type of detector is fairly easy to control insofar as d-c bias voltage and quench frequency are concerned. The bias voltage is controlled by the coupling in the quench frequency transformer and the quench frequency, by the *LC* values in the transformer's grid circuit.

In this second method, the tube oscillates at two separate frequencies. Bias voltages are developed for both frequencies, but since they are concurrent, the measured d-c across the grid resistor will be the higher of the two. The effectiveness of the detector may be determined by the increase of bias voltage due to the injection of some small value of external signal. The greater the change in bias, the more sensitive is the detector. This was found to be the most effective method of determining the best quench frequency. The best quench frequency was one which permitted the greatest change in grid bias voltage for a given small value of signal.

The fact that a small value of signal was used is stressed, since for large values of signal, the detector changes its mode of operation.

It was observed that when a very strong external signal was fed into the grid circuit of the detector the bias voltage decreased and at the same time the plate current decreased, whereas for small signals the bias voltage increased and the plate current decreased. This is an odd result, caused by a change in the mode of detector action. For strong signals the detector acted as a straight amplifier, and instead of the grid drawing power from its own plate, the driving power was supplied by the incoming signal, thereby decreasing the measured plate current. To further explain. The grid circuit of an oscillator may be considered as a load on the plate circuit. Loading the plate circuit increases the circulating current in that circuit and thereby increases the amount of d-c drawn. Now if that load is removed, the Q of the circuit increases, since the resistance introduced by the grid circuit is decreased. This action causes a reduction in plate current. When a strong signal is introduced in the grid circuit, it replaces the energy fed back from the plate circuit. The action is similar to the drop in plate current in an r-f amplifier. A value of signal input was found where the bias voltage remained at a constant value. This is an important reference value, since it is the point where the signal

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input equals the energy fed back from the plate circuit of the detector. The bias resistor should be adjusted at this point to give the greatest value of bias voltage.

Multiple Resonance Peaks

A condition frequently encountered in the design of a s-r detector is broad tuning with multiple peaks of resonance close together. This condition is generally due to either of three factors. . . . *(1)*—Too high a quench frequency. *(2)*—Too much feedback at resonant frequency. *(3)*—Insufficient positive resistance in grid tank.

If the positive resistance in the grid circuit is increased, that is, if the coupling to the previous stage is made tighter, the multiple oscillations will decrease, at a cost of selectivity. Reducing the amount of feedback will increase the sensitivity as well as the selectivity. Decreasing the quench frequency will reduce the sensitivity but increase the selectivity.

Another circuit developed from the experimental model is shown in Figure 6. The results achieved with this circuit were comparable to those obtained with Figure 5. However, the frequency acceptance was broader. The selectivity obtained was still excellent for a portable receiver, since the frequency stability of portable transmitters at ultra-high frequencies precludes the use of highly selective circuits. Absolute sensitivities of 1 microvolt may be obtained at 100 megacycles with excellent results for 10 microvolts input.

With 5 microvolts as a base, a band width of 100 kc at 10 times was obtained with the experimental model. This band width expanded rapidly as the signal was increased to 1,000 times.

By proper adjustment of the band acceptance of the detector, an excellent f-m receiver for an f-m system of communication may be constructed. The second harmonic of a nearby f-m station was picked up by the experimental receiver with excellent quality. The minimum amount of material and tubes required, and the high sensitivity of a s-r receiver admirably fulfill the requirements for portable f-m work.

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Grimes and Borden, "A Study of S-R," Electronics; 1934.

Terman and Ungan, "Some Properties of Grid Leak Power Detection," IRE Proceedings, p. 160; Dec., 1930.

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Armstrong, "Some New Developments of Regenerative Circuits," IRE Proceedings, p. 244; August, 1922.

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By Francis E. Almstead, Lieut. USNR Materiel Officer, U. S. Naval Training School; Kirke E. Davis, head of Science Department, Oceanside High School and George K. Stone, Senior Education Supervisor, State Education Department, Albany . . . 139 pp. . . . New York: McGraw-Hill Book Company, Inc. . . . $.80

An interesting contribution to the study of fundamental radio, covering twenty subjects, from the care and use of the soldering iron to studies of instruments. Presented in the experimental format of discussion, this presentation will prove quite helpful to the newcomer.

The manuscript is based on the New York State syllabus on radio and thus may be used in conjunction with standard texts covering radio fundamentals.

The subject of mathematics is covered in the manual, in a very simple, yet exhaustive manner. The application approach is used in this respect.—*O. R.*

• • •

A.S.T.M. STANDARDS ON ELECTRICAL INSULATING MATERIALS

Prepared by A.S.T.M. Committee D-9 on Electrical Insulating Materials . . . 441 pp. . . . Philadelphia, Pennsylvania: American Society for Testing Materials . . . $2.50 (heavy paper cover)

This edition, issued in February, covers the 1942 compilation of standards. Several of the interesting topics discussed in last year's issue are included, but the major part of this new volume is devoted to new material.

There are groups of specifications and tests on insulating varnishes, paints, lacquers, and molded insulating materials. In the latter section, which includes a discussion of plastics, fourteen test methods are described. The section on plates, tubes and rods, has two specifications and some ten standard test-methods.

A section is also devoted towards a discussion of mineral oils for electrical insulation. Three standards are shown here. In the discussion on ceramics, which includes glass insulators, porcelain and steatite, four standards are given.

Insulating paper, mica products, rubber and textiles are also discussed quite thoroughly and represented by several standards and tests.

An unusual feature of this new edition is the inclusion of the annual report of the committee, as well as several condensed reports on the significance of tests, and an extensive report by Flowers and Fruchtman on "Methods Used for Determination of the Oxidation Tendency of Insulating Oils."

This volume should be in every library.—*O. R.*

• • •

THE MATHEMATICS OF THE PHYSICAL PROPERTIES OF CRYSTALS

By Walter L. Bond . . . 72 pp. . . . New York: The Bell System Technical Journal, 1943, American Telephone & Telegraph Company . . . $.50.

On of the most comprehensive discussions on crystals ever presented, is offered in this excellent paper.

Discussed are such vital problems as the linear vector function, symmetry of crystals, matrix algebra, geometry of rotations, crystalline dielectrics, quadratic forms, crystal elasticity, temperature expansion, temperature variation of the isothermal elastic modulii and stress variation of the temperature, expansion coefficients, the piezo-electric effect, the converse piezo-electric effect, the converse piezo-electric effect as a non-linear function, pyro-electricity, the thermo-electric effect in crystals.

Included in the discussion also are a study of the propagation of light in crystalline media, the electro-optic effect, the piezo-optical effect, application of the electro and piezo-optical effect and transverse isotropy.

This paper should be read by not only those engaged in

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the study or production of crystals, but by everyone in any way affiliated with communications. For it presents a most thorough analysis of a product, that is unusually vital to communications today.—*O. R.*

• • •

PRE-SERVICE COURSE IN SHOP PRACTICE

By William J. Kennedy, Instructor in Machine Shop Practice, Straubenmuller Textile High School, New York City . . . 337 pp. . . . New York: John Wiley and Sons, Inc. . . . $2.00.

The fundamentals of mechanical work, as applied to the requirements of the military, are presented in this new volume.

Prepared at the request of the War Department and the U. S. Office of Education in conformance with the official pre-induction training course outline, the book is quite complete. For it covers such subjects as hand tools, machine tools, measuring tools, layout, cutting, planing and boring, jointing and assembly, finishing and preservation, filing and abrasives, drills, soldering, wiring and wire splicing, ropes, splices and knots and blocks and rigging. The section on wiring and wire splicing is effectively presented in its relationship to communications. Diagramatic illustrations serve to explain the various machine shop procedures and equipment used.

Although the book is written on a high school level, there are many basics of fact presented, that will prove valuable to everyone.—*O. R.*

• • •

PRE-SERVICE COURSE IN ELECTRICITY

By William C. Shea, Instructor in Applied Science and Coordinator of Defense Activities and High School Subjects, Straubenmuller Textile High School, New York City . . . 276 pp. . . . New York: John Wiley and Sons, Inc. . . . $2.00.

A companion volume to the shop practice book, and also prepared in conformance with pre-induction training courses. Fundamentals are also the bases here, too. Prepared in accordance with classroom-session technique, approximately ninety teaching periods are included within the scope of thirteen chapters.

Information, which is pre-requisite to work in radio and allied specialties, has been included in the chapters which cover magnetism, electrostatics, primary cells, the storage battery, voltage-current resistance, electromagnetism, meters, work-energy power, motors, rectification, etc. Symbolic mathematics are employed for the most part.

An interesting fundamental treatment.—*O. R.*

• • •

THE ELECTRON MICROSCOPE

By E. F. Burton, Head of the Department of Physics and Director of the McLennan Laboratory, University of Toronto, and W. H. Kohl, Development Engineer, Rogers Radio Tubes, Ltd., Toronto . . . 233 pp. . . . New York: Reinhold Publishing Corp. . . . $3.85.

Probably the most interesting subject of the day is the electron microscope. Its use has spread to many quarters. And with the development of compact comparatively inexpensive units, it is no longer the restricted laboratory device that it was a couple of years ago. Accordingly, a book explaining in a modified technical manner, the purpose and design of the instrument, is quite useful a tool to have today. Such a tool is this new book, by two distinguished Canadian engineers.

Covered in a graphical style are the subjects of vision, light microscopes, light, wave motion, the electromagnetic theory of light, the electron, motion of electrons, magnetic lenses, the history of the electron microscope, the electrostatic microscope and the magnetic type of electron microscope, etc.

The engineer will enjoy reading this book, even though it is fundamental in its approach. He will enjoy it, because the treatment is absorbing and instructive.—*O. R.*

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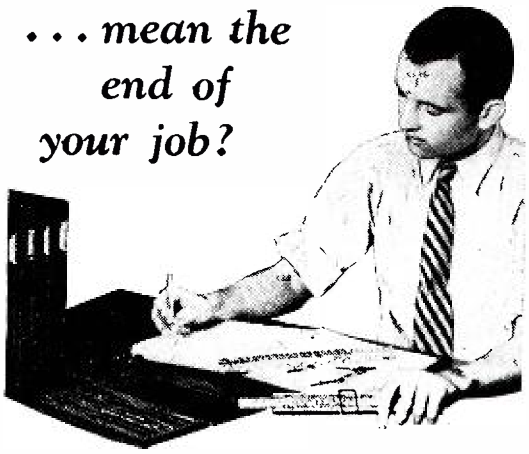

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HIGH VOLTAGE PLATE SUPPLY

(*Continued from page 26*)

class *A*), and the lower powered radio frequency stages of the transmitter. Therefore, all plate currents do not very under operating conditions, and the load on the rectifier is constant. For this reason, the input choke is not of the swinging type. The reactors in both sections of the low-pass filter are of the smoothing type.

Those who must build a power supply to furnish plate voltage to a transmitter employing class *B* modulation, where the d-c plate current to the modulator varies through wide limits under operating conditions, should use a swinging choke as the first reactor. In using the swinging choke, advantage is taken of the fact that the critical value of inductance decreases, in proportion to the decrease in load from the minimum bleeder load to maximum load on the power supply on peaks of modulation, when the modulator plates require maximum d-c plate current. The swinging choke is designed so that it automatically maintains a critical value of *L* throughout the operating range of plate current demand, the amount of *L* decreasing as the plate current demand, and consequently the load, increases.

The problem is solved by selecting a first reactor with minimum inductance equal to the critical value for the maximum load demand, while the bleeder resistance is selected so that the *L* of the swinging choke at maximum swing is the critical value for the bleeder load.

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[4]See above reference.

[5]Scott Helt, "Design of a practical Low-Pass Filter for 600 Ohm Audio Transmission Lines", COMMUNICATIONS; Vol. 20, No. 10, Oct., 1940.

VIDEO AMPLIFIER

(*Continued from page 50*)

the curve now lies within plus or minus 1% to $T/T_g = 1.1$.

By a similar process it should be possible to find a second critical value of β such as to produce a flat phase response. Referring now to Equation 19 for the phase characteristic, our aim is to establish the conditions necessary for maintaining zero phase shift over as wide a range of frequencies as possible. It can be seen that the phase angle will remain zero for all values of T/T_g if the coefficients in the numerator are zero. This condition is readily satisfied for the first-power coefficient by placing $\beta = m$, but the coefficient of $(T/T_g)^3$ can be zero only if β is infinite or if $m = -1$. While this precludes the possibility of preventing phase shift, we again note that β should be large; a requirement compatible with that for a flat amplitude response.

It can be shown that equating *m* and β causes the first derivative of θ with respect to T/T_g to vanish at the origin, hence the second critical value of β is

$$\beta = m = \beta_2 \qquad (23)$$

If β exceeds this new critical value, the phase response curve rises more sharply, while if β is made smaller than β_2 a minimum will occur, which means that the phase angle will become negative for certain values of T/T_g. Figure 8 shows the response curves for the case where $\beta = \beta_2$. Note that the amplitude response now exhibits an appreciable peak for both values of *m*.

The Transient Response of the Compensated Amplifier

Applying a unit voltage step to the compensated amplifier, the transient load voltage is found to be

$$e_L(t) = g_mR + g_mR_r(1 - \varepsilon^{-t/T_r}) \qquad (24)$$

At the instant the unit step is applied, the load voltage rises immediately to

(*Continued on page 75*)

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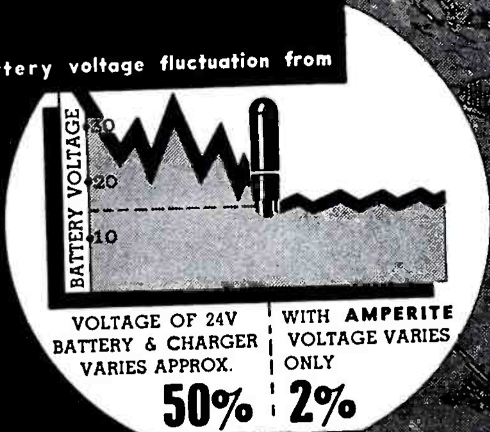

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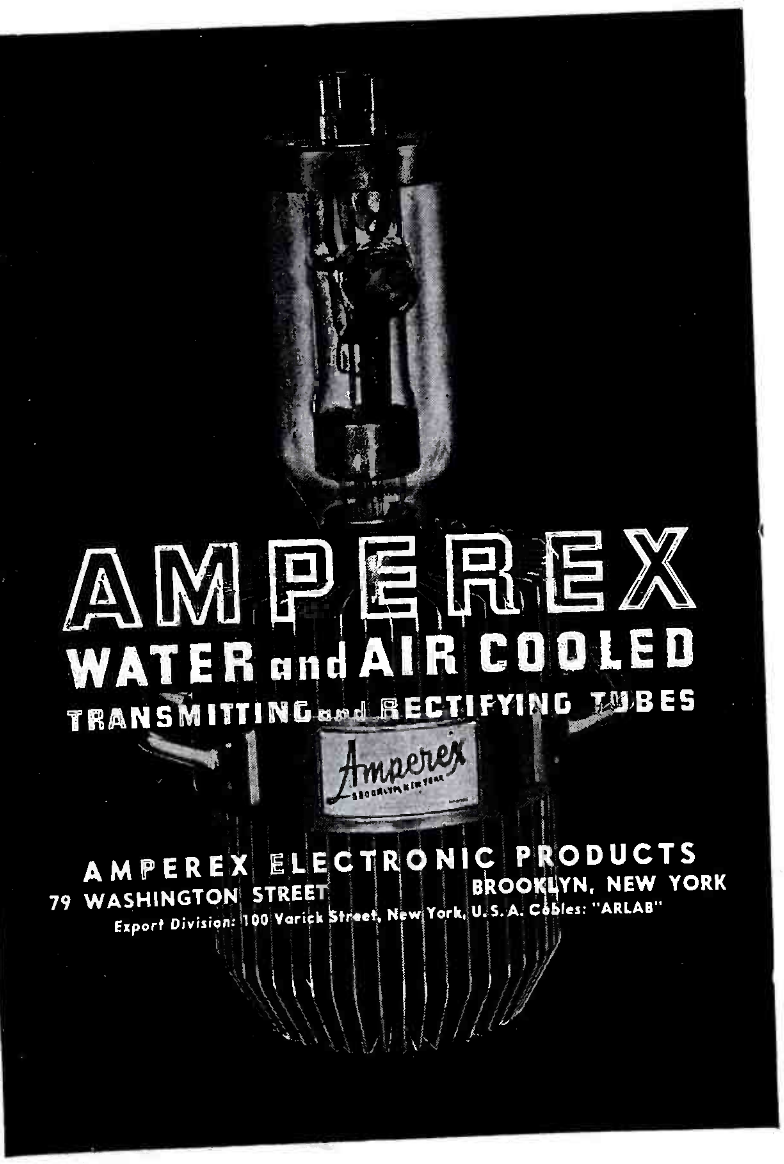

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(*Continued from page 36*)

tenuation at the fundamental frequency, the wave is badly distorted. This distortion is attributable almost completely to non-uniform time delay (now-linear phase shift) of the various frequency components. Since the total phase shift for a half-section filter is a constant independent of the cut-off frequency, the way to reduce the delay distortion is to use a smaller fraction of the filter pass-band. This is accomplished by using a smaller-value of $\frac{f_c}{f}$. Thus, if $\frac{f_c}{f} = \frac{1}{10}$, the distortion becomes quite small, as seen from the figure.

In the same manner we could compute and plot the output wave-forms for full-section filters. For example: for a constant $-k$ low-pass full-section filter of either the t or π type we have

$$e_p(t) = \frac{1}{2}\left\{1 - \frac{\epsilon^{-\omega_c t}}{1+\epsilon^{-\omega_c T/2}} - \frac{2}{\sqrt{3}}\,\frac{\sin\frac{\sqrt{3}}{2}\omega_c t + \epsilon^{-\frac{\omega_c T}{4}}\sin\frac{\sqrt{3}}{2}\omega_c\left(t-\frac{T}{2}\right)}{1+2\epsilon^{-\omega_c T/4}\cos\frac{\sqrt{3}}{4}\omega_c T + \epsilon^{-\frac{\omega_c T}{2}}}\,\epsilon^{-\frac{\omega_c}{2}t}\right\}, \quad 0 < t < \frac{T}{2} \qquad (33)$$

Application to Rectifier Filters

In designing filters for radio transmitter power supplies various factors must be taken into consideration. Some of these factors will be discussed here.

Usually the prime consideration in the design of a power supply (excepting, of course, the nature of the d-c power delivered) is the a-c ripple out-

(*Continued on page 76*)

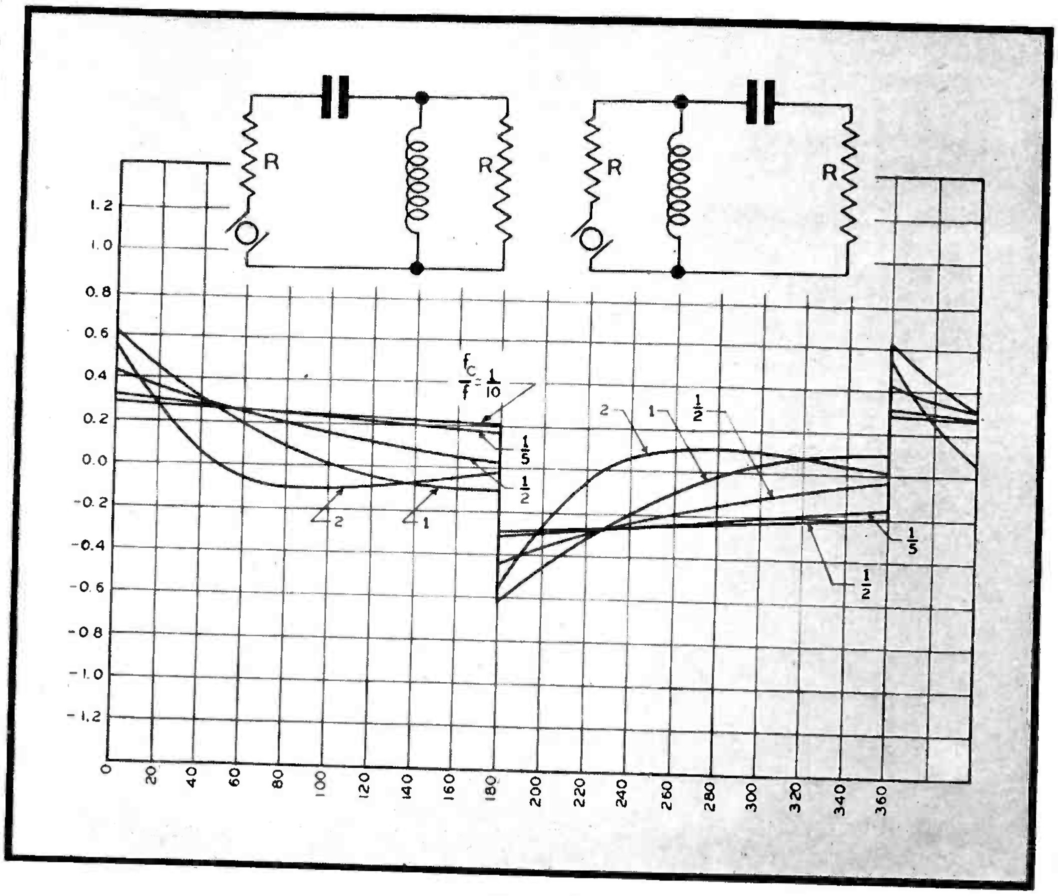

Figure 6
Response of high-pass constant-K half section filters to rectangular waves.

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(*Continued from page 72*)

a value of G_mR or G_0 volts. As the time t increases, the voltage continues to rise, only now exponentially, approaching a value $g_m (R_F + R)$ volts asymptotically, as the condenser C_F charges.

With a unit square wave applied, the load voltage has the general appearance shown in Figure 9. It is apparent that this characteristic tends to cancel the downward tilt caused by the grid time constant.

Continuing with the response to the unit step, it is a simple matter to write a differential equation for the divider action of the grid circuit, making use of the simplifying assumption that the grid circuit is a negligible shunt on the load. We may then equate the voltage drops across R_g and C_g to the transient load voltage from Equation 24. Solving this equation and writing the overall transient response as a relative quantity, we finally obtain

$$H(t) = \frac{1}{\beta - 1}\left[m\varepsilon^{-t/\beta T_g} - (m + 1 - \beta)\varepsilon^{-t/T_g}\right] \qquad (25)$$

This is a third basic equation upon which we can operate. Once again we wish to determine a value of β such that the response curve will have zero slope at the origin and not exhibit a peak beyond that point. A peak in the transient response will distort a square wave considerably, causing it to be bowed upward if the peak occurs at a time corresponding to one quarter of the square wave period, or causing it to be high on the trailing edge if the peak occurs later in the cycle.

Having discovered that phase shift distorts a square wave one feels intuitively that the critical values of β for flat phase and flat transient response will coincide. This turns out to be so, and it can be verified readily by taking the derivative of the transient response with respect to t/T_g, equating it to zero and solving for β when $t/T_g = 0$. When this is done it is found that

$$\beta = m = \beta_2$$

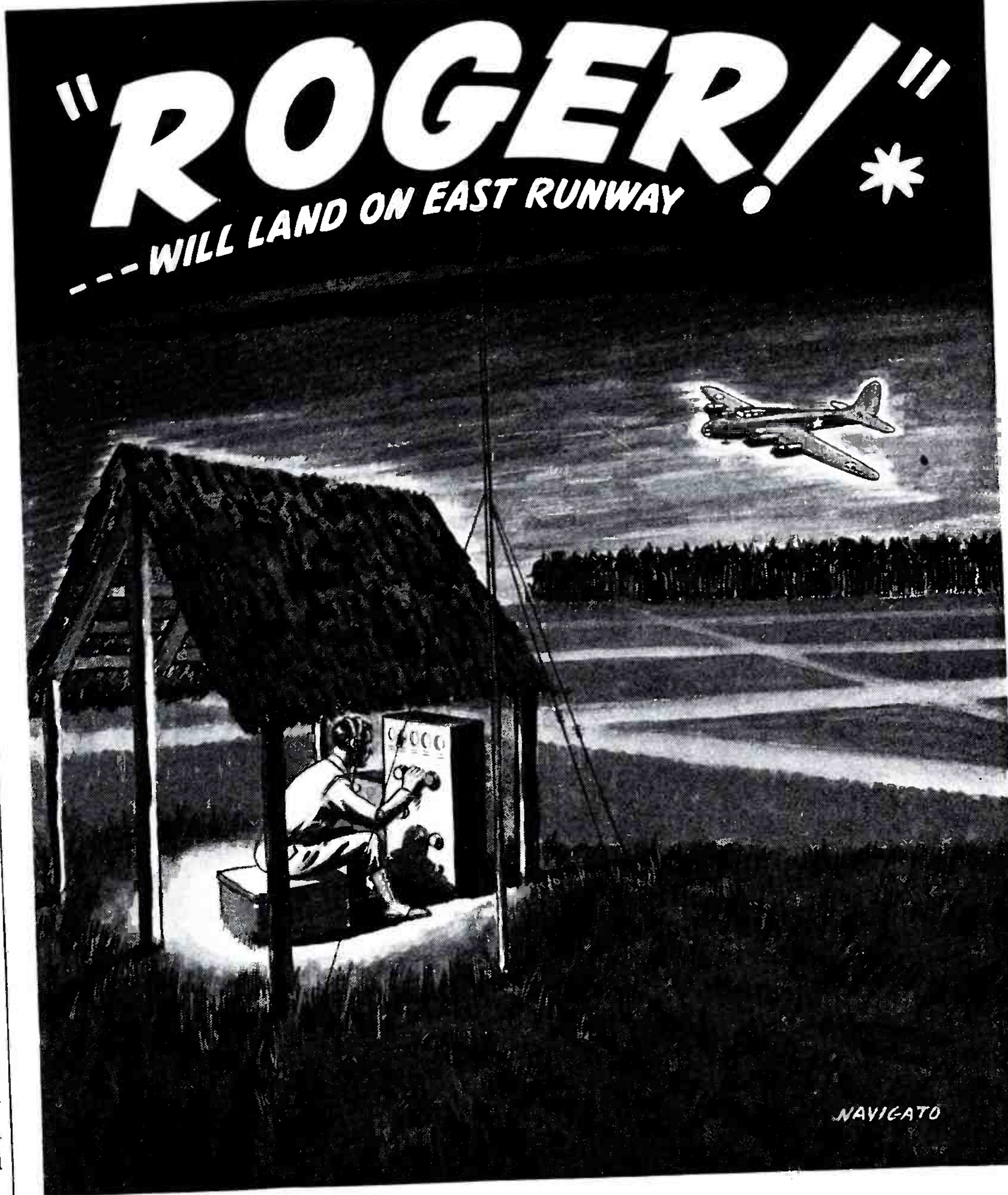
Figure 10 shows the transient response plotted for the two values of m and for the uncompensated case.

As in the case of the amplitude response, it is possible to adjust β to a value slightly smaller than m and allow the transient response to bow upward by a small amount. Thus the response can be maintained within a given tolerance for a greater period of time. Again the degree of "overcompensation" is a matter of judgment, and depends upon the number of stages involved.

Figure 11 shows an enlargement of the transient response curves, and in addition shows the response when $m = 10$ and $\beta = 9.5$. In this case the response variation lies within plus or minus 1% for the duration of one grid time constant. For this same case, the phase response is constant within plus or minus 0.5 degree to $T/T_g = 0.7$.

The square-wave response of the compensated amplifier will follow the transient response if the period of the square wave is longer than the time required for the completion of the transient. If the period of the square wave is shorter, the square wave response will be somewhat different from the transient response. In practice, however, the plate filter components can be designed with $\beta = m$. The circuit elements are then found to be

(*Continued on page 78*)



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(*Continued from page 74*)

put. The relative magnitude of this ripple is determined by the nature of the main power supply (frequency, number of phases); the attenuation of the filter, and, in some cases, where particularly low ripple requirements are specified, on the balance of the phase voltages supplied to the rectifier tubes[11].

A second well-known consideration is the input impedance to the filter for the a-c ripple voltages applied. If this impedance is not sufficiently great for all frequencies involved, the alternating component of current will be large enough to produce current cut-off during part of the a-c cycle resulting in poor regulation of the supply[11].

A third consideration is the surge current through the rectifier tubes when the supply is turned on initially. If the filter design is not inherently such as to limit this surge to a safe value, the circuit is provided with a surge resistor which is short-circuited by a relay after a suitable delay.

A fourth consideration, in telegraphic transmitters, is distortion of the marking wave owing to transients in the filter of the power supply of the final radio-frequency amplifier[12]. A similar, although less severe problem exists in amplitude-modulated amplifiers by virtue of transient power requirements produced by sharp speech transients.

In amplitude-modulated transmitters a further requirement is that the impedance of the filter measured across its *output* terminals to the lowest audio frequency used must be small compared to the plate resistance of the modulated r-f amplifier, if undesired attenuation of low audio frequencies is to be avoided.

If a class *B* modulator and the final r-f amplifier are supplied from the same rectifier, as is often the case, care must be taken with respect to the output impedance of the filter that the audio pulse currents supplied to the modulator do not develop sufficient voltage across the output of the filter to modulate the r-f amplifier additionally and thus give rise to additional audio distortion in the r-f output.

Finally, economy is almost always a factor, particularly in the comparatively high-power supplies under discussion here, and conservation of space is often of great importance.

In high-power telegraphic transmitters it has usually been the practice to employ three-phase main supplies. In view of the preceding discussion, three-phase full-wave power supplies are particularly advantageous. The total rms ripple input to the filter is only 4.2%[13], thus requiring relatively simple filtering, if indeed, any filtering is considered necessary. The input impedance of the filter is not a very important consideration from the point of view of regulation since the regulation due to insufficient input impedance cannot exceed 4.7%[13]. This is not to say that the input impedance is of no consideration whatsoever. A certain amount of impedance, readily calculable, is required to limit the peak cyclic current through the rectifier tubes to a safe value. This peak cyclic current must be computed as the sum of the d-c and a-c components of current through the tubes.

As a consequence of these facts comparatively small inductance and capacitance values may be used in the filter without violating the keying requirements. As a further consequence, however, it often becomes necessary to use a surge resistor and associated relay to limit the starting current.

[*To Be Continued*]

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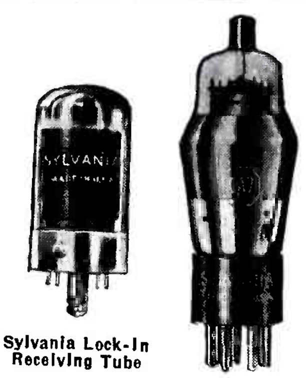

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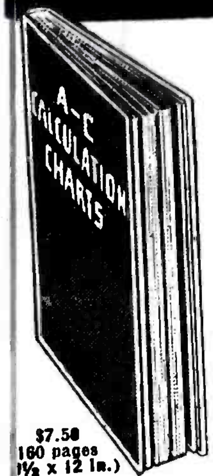

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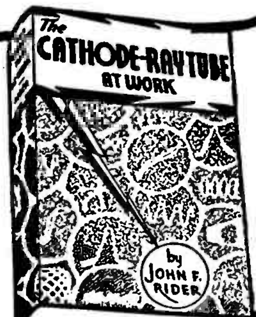

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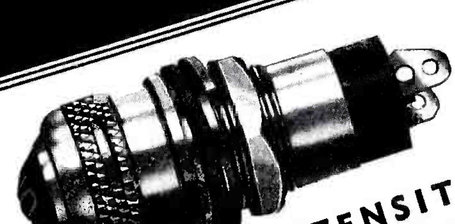

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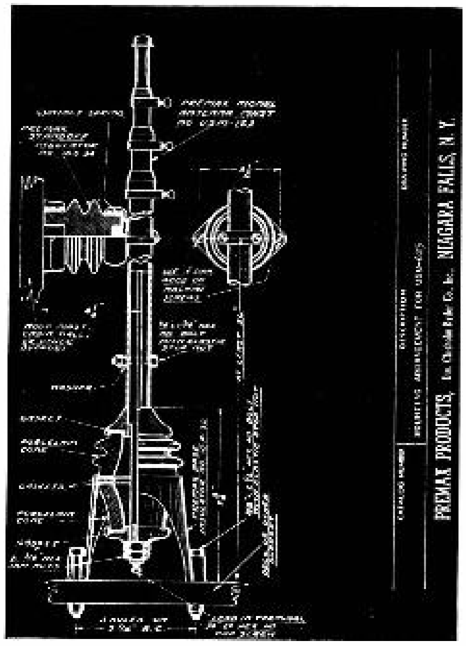

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(*Continued from page 75*)

$$R_F = mR \tag{26}$$

$$C_p = \frac{R_g C_g}{R}$$

The final adjustment of the circuit can be made by increasing R_g sufficiently to make the square-wave response as perfect as possible. The above equation for C_F is based upon β and *m* being equal. In general C_F if found to be

$$C_F = \frac{R_g C_g \beta}{R m} \tag{27}$$

To summarize the results, two unique values of β have been found, one of which produces a specific amplitude response, and the other a specific phase and transient response. If a flat amplitude response is required, β should be made equal to or slightly smaller than β_1. If a flat phase and transient response is required, β should be made equal to or slightly smaller than β_2. There is no point in making β larger than β_1.

A proportionality which is frequently used results from the equation of the following time constants

$$C_F \frac{RR_F}{R + R_F} = R_g C_g$$

This develops into an interesting special case where $\beta = (m + 1)$, a value exceeding the maximum limiting value. In this case the three basic equations reduce to the following simple forms . . .

$$|H|^2 = \frac{1}{1 + \left(\frac{T}{T_F}\right)^2}$$

$$\theta = \tan^{-1}\frac{T}{T_F} \tag{28}$$

$$H(t) = \varepsilon^{-t/T_F}$$

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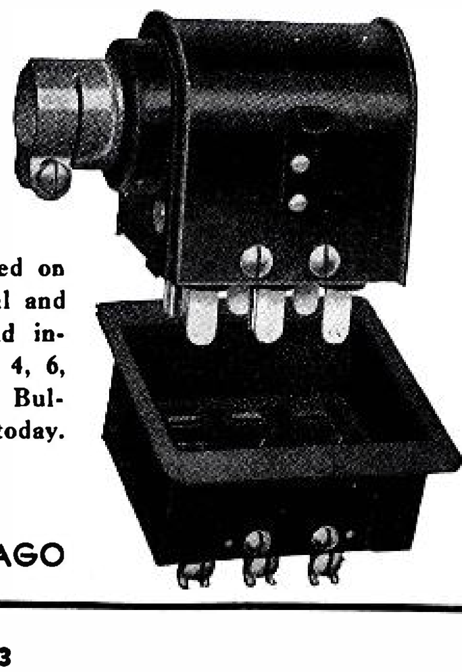

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(*Continued from page 78*)

Here, T_F has been substituted for T_g in the equations for the uncompensated amplifier. The response characteristics are precisely those of an uncompensated amplifier having a grid time constant equal to T_F. The proportionality mentioned above gives satisfactory results only when m is very large. The reason for this is that the response curves for various values of β show a smaller departure from one another when m is large. It is important therefore to use the proper value of β when m is, by necessity, small.

Overcompensation

If β is appreciably smaller than m, overcompensation results. In this case, the amplitude and transient responses peak badly, and the phase angle becomes negative by quite a number of degrees over a considerable range of values of T/T_g.

Photographic Illustrations

The accompanying photographs, Figures 12 to 16, show the square-wave response for several values of β and m and for the uncompensated amplifier. The half-period of the square-wave in each case is equal to one grid time-constant. Of course, the vertical lines of the square-wave do not show in the photographs, because the spot is moving too fast to register.

Amplifier Used

The amplifier used for the photographs is a specially constructed two-stage video amplifier used in the study of low-frequency compensation in the Graduate Electrical Engineering Laboratories of the Polytechnic Institute of Brooklyn. The circuit is arranged so that R_F and C_F can be independently and continuously varied. Furthermore, the grid time constant, T_g, is about 1/10 of the normal practical value, so that the response for large values of T/T_g can be observed on the oscilloscope, without encountering the difficulty due to distortion being introduced by the oscilloscope amplifier at very low frequencies.

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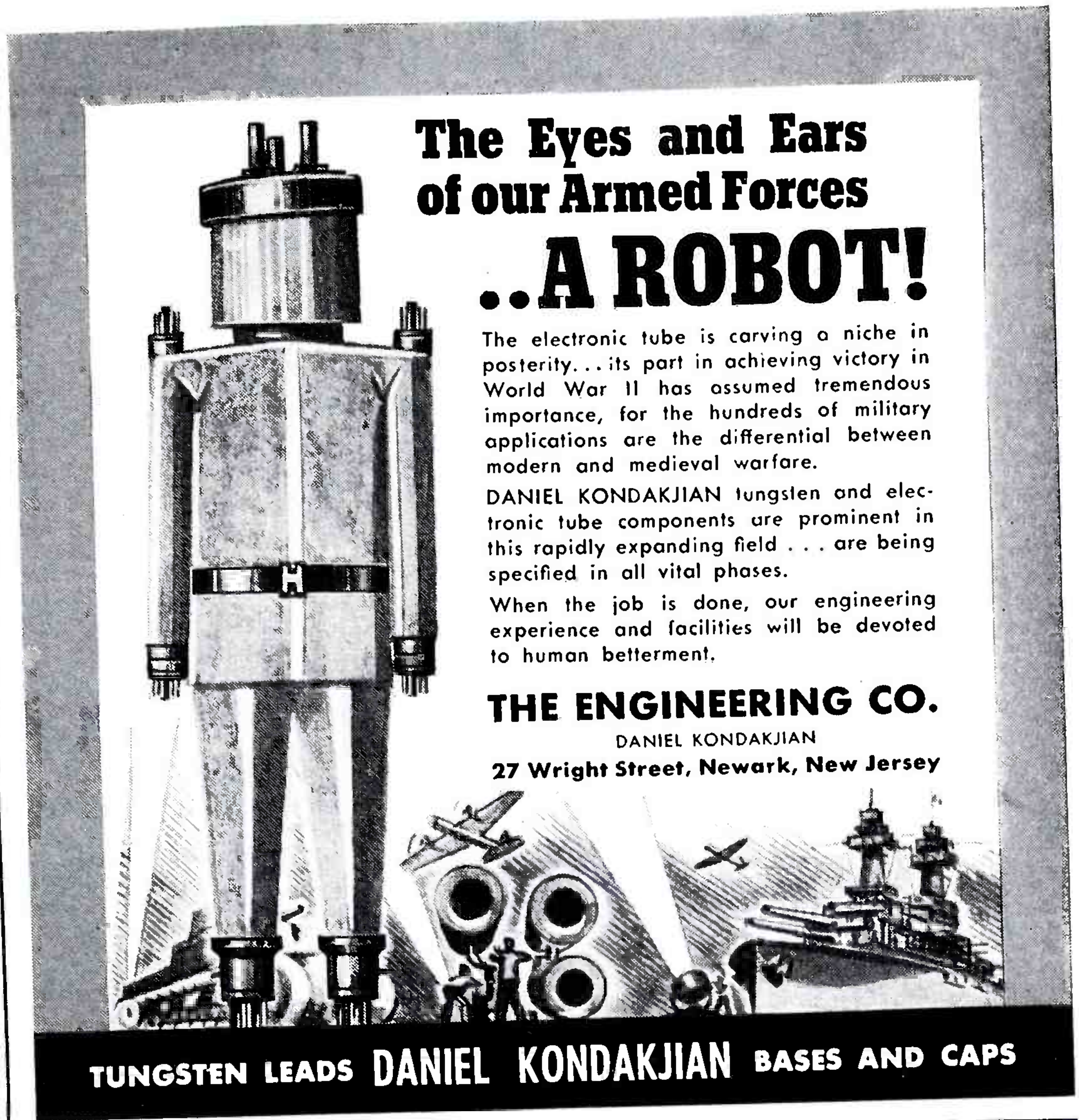

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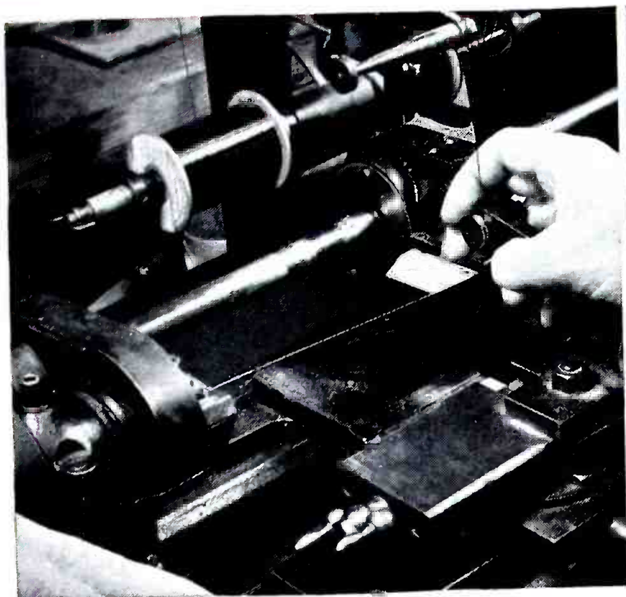

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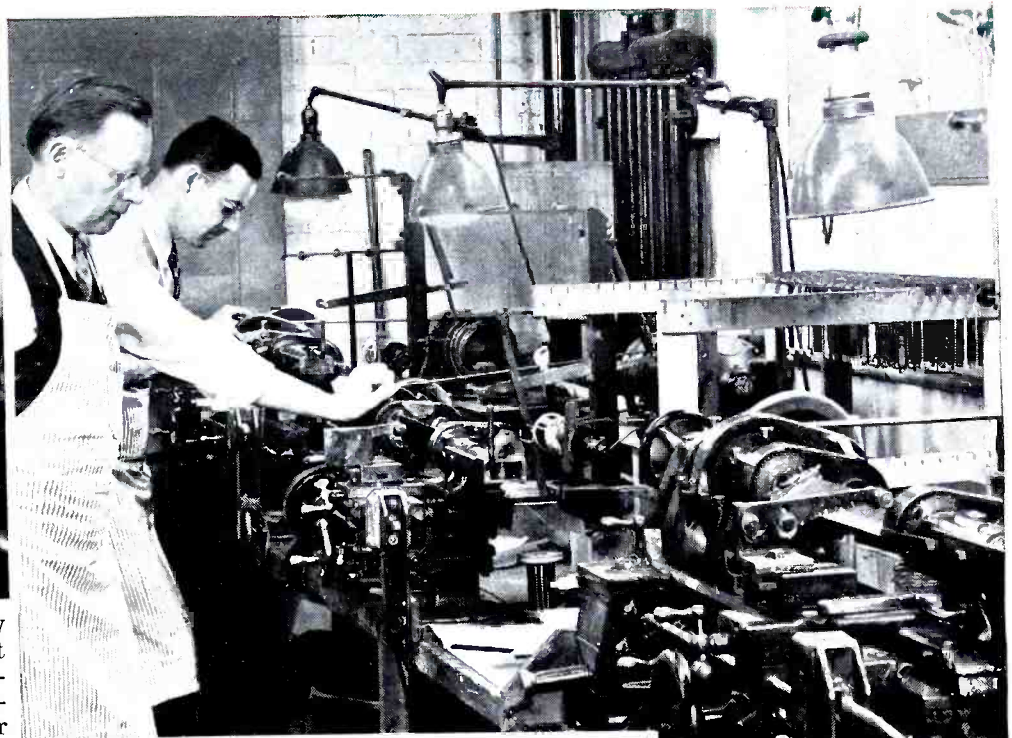
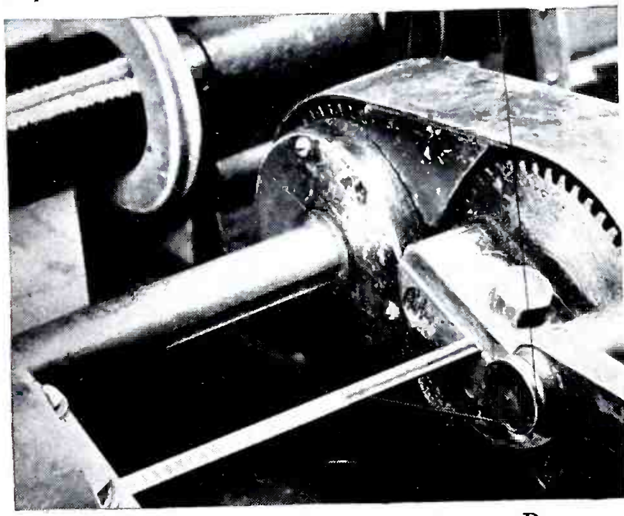
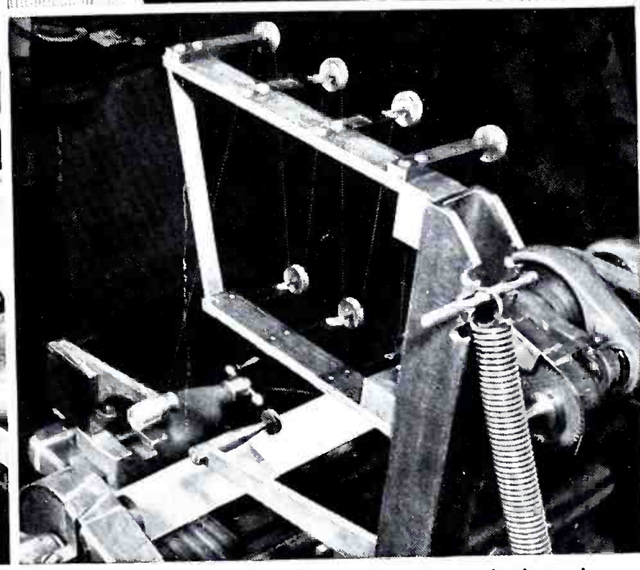




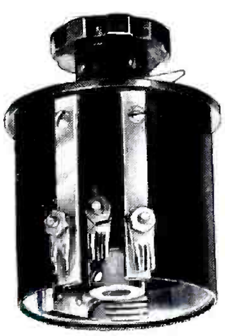
Putting the *second* winding on an Ayrton-Perry, or non-inductive, resistance unit. This unit is used as the output control in a standard-signal generator operating at frequencies up to 50 megacycles.



Spring-mounted pulleys absorb variations in wire tension as the card turns. Constant tension is provided by a spring device on the shaft carrying the spool of wire.

Although the winding of all resistance cards for General Radio potentiometers is essentially automatic, experienced workmen adapt the winding speed to each particular unit through Variac control of the motor drive.

How G. R. Rheostat-Potentiometers are Wound



Because all of our facilities are devoted to war projects, these rheostat-potentiometers are at present available only for war work.

The resistive elements are wound on flat fabric-base phenolic cards, which are then bent around molded cylindrical forms. To achieve definite resistance characteristics — linear, parabolic, or logarithmic — many sizes and shapes of cards are used. More than one size of wire on a single card necessitates abrupt changes in card width. For non-inductive units, two similar windings in opposite directions on a single card are necessary.

General Radio has developed winding methods and adapted standard lathes to produce all these various windings. Constant-tension devices and automatic feed insure precise control of winding. The finished resistance element has turns accurately spaced and presents a smooth uniform surface to the sliding contact. This results in long life and trouble-free operation.

General Radio instruments use a wide variety of these variable wire-wound resistors as calibrated controls in bridge and other measuring circuits. Originally designed for our own use, these rheostat-potentiometers are essential elements in many electronic instruments and are now widely used by other manufacturers of precise electrical equipment.

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