RADIO

JUNE, 1946

MANUFACTURING AND BROADCASTING

The Journal for Radio & Electronic Engineers

Design • Production • Operation

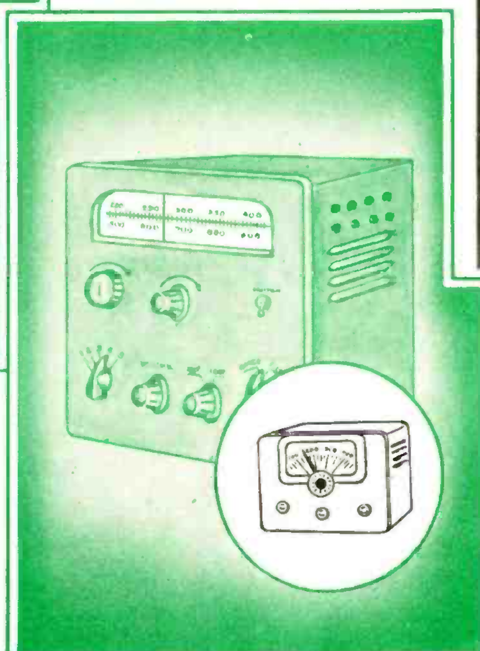
SYLVANIA NEWS

CIRCUIT ENGINEERING EDITION

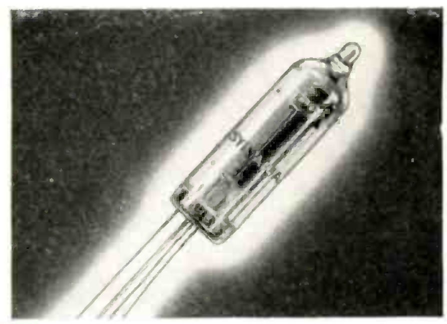
JUNE *Prepared by* SYLVANIA ELECTRIC PRODUCTS INC., Emporium, Pa. 1946

RUGGED ELECTRONIC TUBE TINY ENOUGH TO REVOLUTIONIZE DESIGN OF RADIO RECEIVERS AND OTHER EQUIPMENT

More compact television receivers will be made possible by the T-3.



Weight-saving features of the T-3 will be of special value in air-borne equipment.



The T-3 tube is shown here in its actual size.

Equipment for motor boats and yachts can be made smaller and lighter.

Much Smaller Sets Possible

Radical reductions in the size and weight of many types of electronic equipment are seen as a distinct possibility arising from Sylvania Electric's development of the extremely small T-3 tube. The T-3 is the commercial version of the peanut-sized electronic tube of proximity fuze fame.

Tiny as it is, the T-3 tube is characterized by exceptional ruggedness. It has a life of hundreds of hours, and is ideally suited for operation at high frequencies.

Savings in Space and Weight

The small size of the T-3 contributed directly to compactness and lightness in the design of radio and television receivers and other types of electronic equipment. Other features of the tube make possible still further reductions in space and weight.

Range of Applications

The design possibilities opened by the T-3 are naturally of greatest interest in the case of portable and air-borne equipment. However, its potentialities are not limited to these fields. *Write Sylvania Electric Products Inc., Emporium, Pa.*

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Emporium, Pa.

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RADIO

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John H. Potts . . . . Editor
Sanford R. Cowan . . . . Publisher

JUNE, 1946 Vol. 30, No. 6

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COVER

Testing r-f plumbing at the Microwave Research Laboratory of De Mornay-Budd, New York. Here, frequency measurements are made within 2.5 mc at 10.000 mc.

ARTICLES

MISCELLANEOUS

MID-WEST OFFICE
Harold J. Sugarman, Manager
82 West Washington St., Chicago 2, Ill.
ANDover 1395

PACIFIC COAST REPRESENTATIVE
H. W. Dickow
1387 40th Ave., San Francisco 22, Calif.

GREAT BRITAIN REPRESENTATIVE
Radio Society of Great Britain,
New Ruskin House, Little Russell St.,
London, W.C. 1, England

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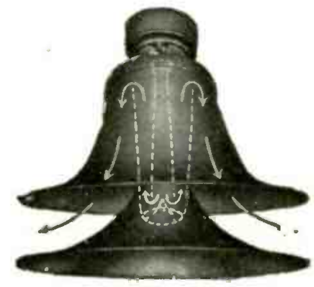

RADIAL HORN SPEAKER; a 3½' re-entrant type horn. Projects sound over 360° area. Stormproof. Made of RACON Acoustic Material to prevent resonant effects.

PAGING HORN; extremely efficient 2' trumpet speaker for use where highly concentrated sound is required to override high noise levels. Uses P.M. unit.

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Transients

TELEVISION TANGLE

★ During May the Federal Communications approved applications for new television stations in nine cities, and indicated that pending applications in twenty-seven other cities may be granted shortly. The speed with which these applications are being processed indicates that the FCC is determined to encourage the exploitation of television along present-day lines. Further, concrete evidence of renewed confidence among manufacturers is revealed by a recent survey conducted by the FCC, which indicates that over 100,000 television receivers are scheduled for production during the remainder of this year.

The sore spot in the whole television picture has been the battle against present standards conducted by the Columbia Broadcasting System. When CBS resurrected the old mechanical color scanning method and gave a successful demonstration of it under carefully controlled conditions, unquestionably present-day television suffered a severe setback. When this demonstration was followed up by national advertising under the sponsorship of CBS showing the advantage of high definition television as compared with what we have now, many resolved to wait until further technical progress had been made before plunging into the field.

The practical realization that both color and high definition television, as at present demonstrated, are subject to severe limitations, too, has tended to restore confidence among those interested in the field. In a book entitled "The Truth About Color Television", published by Allen B. Dumont Laboratories, Inc., the limitations of the mechanical color system are discussed and the advantages of electronic color television emphasized. However, it is pointed out that a minimum of 6½ years must elapse before commercial electronic color television can proceed, according to present schedules.

To most of us who have seen how fast development on complex electronic circuits can be completed under stress of wartime necessity, this length of time will appear excessive. But when it is realized that more than half this time will be required for agreeing upon standards, providing color receivers and stations, making field tests, and securing FCC approval, it does not seem very far out of line. Personally we don't think that an additional three years of research will be required before acceptable electronic scanning can become a reality — a good deal of time has already been spent on this sort of work. And we don't believe that the industry should attempt to jump the gun, and foist on the public the old, mechanical color system, which dates back to 1928. A good, black-and-white picture should be acceptable to a television-starved public for a few years, at least, and the public will appreciate not being stuck with mechanical color receivers which are more costly, and quickly become obsolete.

TOO MUCH, TOO SOON

★ Earlier howls from receiver manufacturers about the difficulty in getting components have been replaced, in some quarters, by complaints that orders for an entire year's requirements of certain parts are being dumped on the manufacturer at one time.

This is often the case when the parts ordered represent items which the component parts manufacturer has previously made for the government. The tremendous supply of surplus components now being held by many manufacturers tends to confuse the picture and give the impression that production of such items has reached astronomical proportions. This is far from being the case; the shortage of raw materials is still great, and it will probably be early Fall before enough for real mass production is available. When this condition is reached, we shall see a tremendous upswing in production and consumption of such components.

PLASTIC CONSOLE CABINETS

★ The shortage of console receivers becomes daily more acute. Because most cabinet manufacturers find that it is more profitable to make furniture rather than radio cabinets, it is difficult to obtain the latter.

This is a good time to think of switching to plastic cabinets for consoles as well as midgets. While moulds would be expensive, if radio manufacturers would get together and pool their orders, making slight variations in small areas for individuality in cabinet design, it would make possible a revitalizing of the radio receiver market, which is now flooded with midgets that aren't selling.

Plastic cabinets can be cheaper, look better, and are more easily cleaned than wooden ones. *J. H. P.*

TECHNICANA

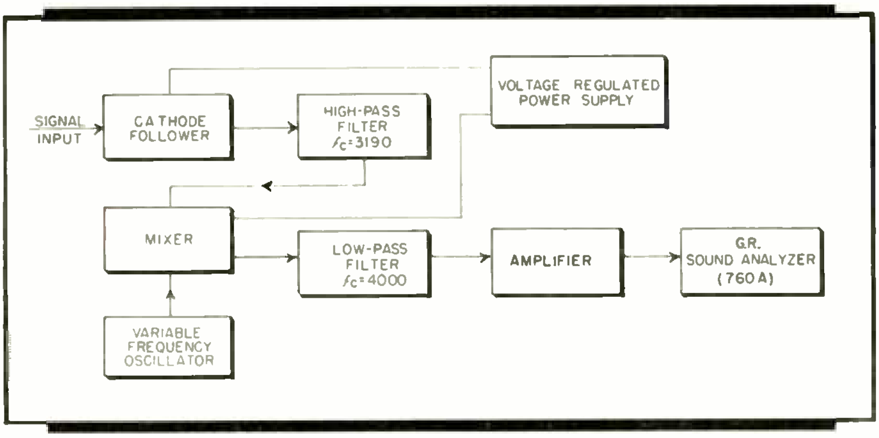


Figure 1

RANGE EXTENDER

★ Sound analyzers may have their response ranges greatly extended by means of a heterodyne system designed by J. D. Cobine and J. R. Curry, and described by them in the *Review of Scientific Instruments* for May 1946, in an article entitled "Range Extender for G-R 706 Sound Analyzer."

A block diagram of the equipment is shown in *Fig. 1*, which was constructed to extend the range of a sound analyzer to 1 mc. This analyzer was originally intended to measure complex waves with frequency components between 25 and 7500 cps. The setting of the variable frequency oscillator yields the frequency of the noise component being measured. The heterodyne system was originally designed to use the sound analyzer as a tuned detector set at 2 kc.

The schematic diagram of the instrument is shown in *Fig. 2*. The oscillator and mixer beat high-frequency signals down to the frequency accepted by the analyzer. In their article, the authors describe the circuits in considerable detail, discuss physical arrangements, and set forth techniques of calibration and application of the range extender in the study of high-frequency noise spectra.

MINIATURE MICROPHONE

★ German supersensitives microphones of very small dimensions, of possible application in the production of light-weight telephone hand sets, are described in a report by the OPB, Department of Commerce.

The supersensitive microphones were developed by Electric Acoustic, at Kiel, Germany, in connection with the production of devices for firing explosive missiles in water by means of electrical and acoustical energy. Thermodynamic and capacity type microphones were built for this purpose.

Equipment used for measurement and calibration of the microphones may also be of interest to American manufacturers of commercial telephone components, in the opinion of the investigator. The report recommends that samples of this equipment be obtained for further study but does not describe the equipment in detail. No other developments of particular interest to American manufacturers are discussed in the report.

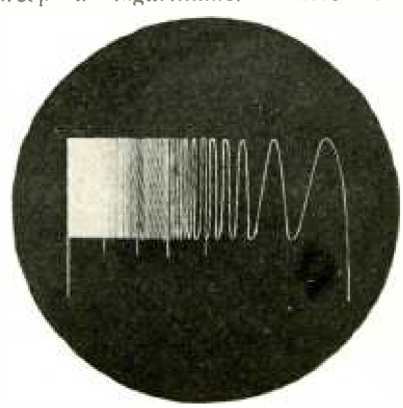
Orders for the report should be addressed to the OPB, Department of Commerce, Washington 25, D. C., and should be accompanied by check or money order for $3 payable to the Treasurer of the United States.

SWEEP FREQUENCY TRANSCRIPTION

★ A 100-10,000 cps sweep-frequency transcription developed by Wayne R. Johnson offers a new method of making instantaneous response checks of entire audio systems as well as of individual components.

This transcription when used in conjunction with an oscilloscope is claimed to give a complete response characteristic picture of all types of audio systems, amplifiers and accessory equipment such as phono reproducers, preamplifiers, filters, transformers and attenuators.

The transcription has a frequency range from 100 cps to 10,000 cps with a repetitive rate of 20 cps recorded at constant amplitude below 500 cps and constant velocity above 500 cps. The sweep is logarithmic. There is a



synchronizing pulse of 200 microseconds duration at start of sweep to lock the oscilloscope. Frequency markers are provided at alternate thousand cycles. Frequency response variations are held within plus or minus one decibel. It is made on 10" vinylite at 78 rpm and on 16" at 33 1/3 rpm, the latter recorded with N. A. B. curve. It is made by Clarkstan Corp., 11927 W. Pico Blvd., Los Angeles 34, California.

[*Continued on page* 8]

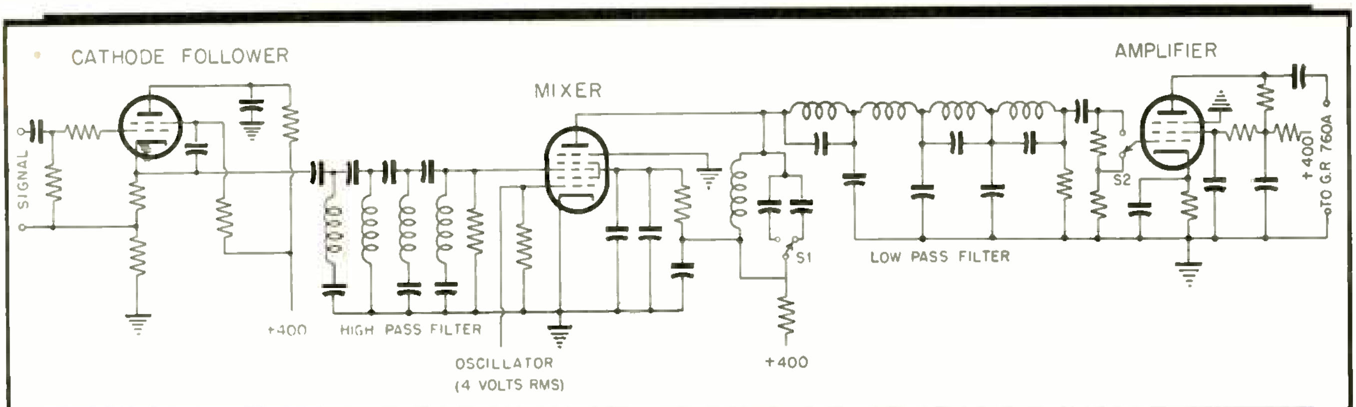


Figure 2

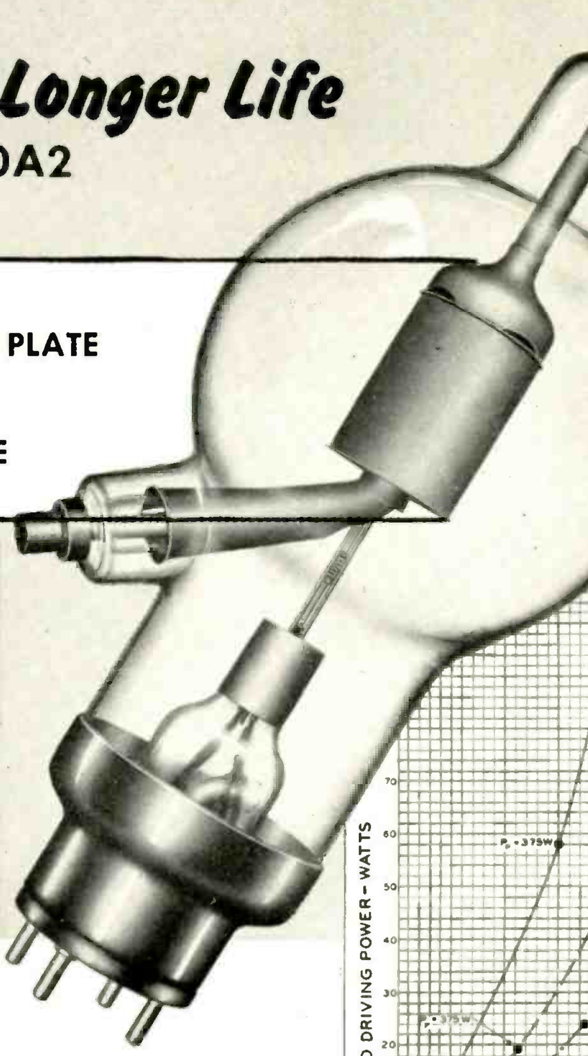
THE COUNTERSIGN OF DEPENDABILITY IN ANY ELECTRONIC EQUIPMENT

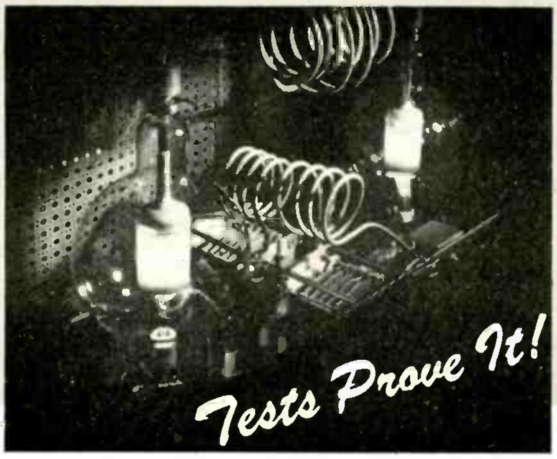
Tests Prove 100% Longer Life

in this New Eimac 3-750A2

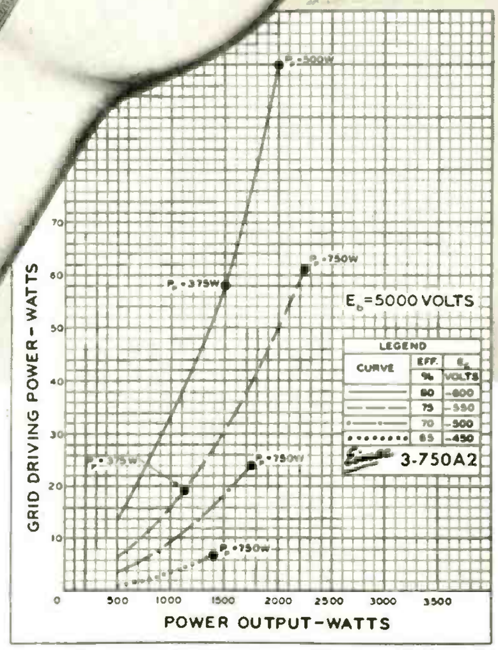
made possible by:

NEW COOLER OPERATING PLATE
NEW NON-EMITTING GRID
NEW FILAMENT STRUCTURE





Two Eimac 3-750A2 Tubes undergoing life tests in the Eimac testing laboratory



Typical of the outstanding performance of this new triode is its high power gain. With 5000 volts on the plate, the new Eimac 3-750A2 will deliver 2000 watts output with only 50 watts driving power, at a plate efficiency of 75%. (See above chart.)

Repeated tests of the new Eimac 3-750A2 in the Eimac testing laboratory show 100% longer life than previous models operated under the same conditions.

This increase in life expectancy is a result of continuing research, culminating in this new version of the 750TL triode. Among its many new features are a new cooler operating plate, new non-emitting grid and a new filament structure.

The new 3-750A2 *is a power triode, interchangeable with the previous model* 750TL, *and is but one example of the constant effort made at Eimac to furnish better tubes at lower cost. For further information and complete engineering data on Eimac tubes, write direct or contact your nearest Eimac representative.*

ELECTRICAL CHARACTERISTICS

Filament: Thoriated tungsten		
Voltage	7.5	volts
Current	21.0	amperes
Amplification Factor (Average)	15	
Direct Interelectrode Capacitances (Average)		
Grid-Plate	5.8	uuf
Grid-Filament	8.5	uuf
Plate-Filament	1.2	uuf
Transconductance (I_b=1.0 amp., E_b=5000, e_c= −100)	3500	umhos
Frequency for Maximum Ratings	40	mc

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M. B. PATTERSON (W5CI), 1124 Irwin-Keasler Building, Dallas 1, Texas. Phone: Central 5764.
ADOLPH SCHWARTZ (W2CN), 220 Broadway, Room 2210, New York 7, New York. Phone: Courtland 7-0011.
HERB BECKER (W6QD), 1406 S. Grand Ave., Los Angeles 15, California. Phone: Richmond 6191.
TIM COAKLEY (W1KKP), 11 Beacon Street, Boston 8, Massachusetts. Phone: Capitol 0050.

TECHNICANA

[from page 6]

H-F MEASUREMENTS

★ Techniques of radio measurements in the decimeter and centimeter region are discussed by R. J. Clayton, *et al.*, in the *Journal of the Institution of Electrical Engineers*, Part III, (quarterly) March, 1946.

A review of coaxial and waveguide theory is presented as an introduction, with a discussion of high-frequency oscillators of the triode, magnetron, and velocity-modulation types. Measurement of frequency, power, impedance, and voltage are described; frequency measurement by heterodyne methods, and resonant cavity wavemeters are dealt with. Designs are outlined for bolometer and calorimeter circuits for powers between one microwatt and several hundred watts. Measurements of reflection coefficient and Q are considered in the impedance section.

Derived measurements are developed in the final portion of the article; receiver sensitivity, antennas, and field strength are treated. Thermal noise voltage is adopted as a standard in establishing receiver sensitivity, and suitable designs for signal generators and sensitivity measuring equipment are discussed. Shielding requirements of the signal generator are noted, with diagrams of suitable leak-proof r-f joints using gasket or biting seals.

Measurements of antenna gain and impedance are set forth, with methods of plotting polar diagrams. The article concludes with a discussion of equipment for absolute measurement of field strength at centimeter wavelengths.

QUADRANT ANTENNA

★ A new wide-band, omni-directional, horizontal V type of antenna is described by N. Wells in the *Marconi Review* for January-March 1946. As

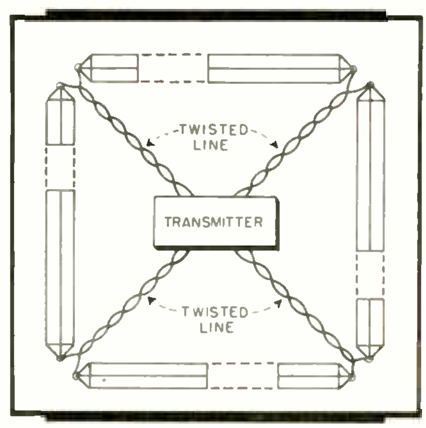

Figure 3

sketched in *Fig. 3*, the antenna consists of a group of four poles carrying four

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TECHNICANA

[*from page* 8]

antennas which together cover a frequency band of, e.g., 2 to 30 mc.

The antenna is designed as a right-angled horizontal V or quadrant, and is

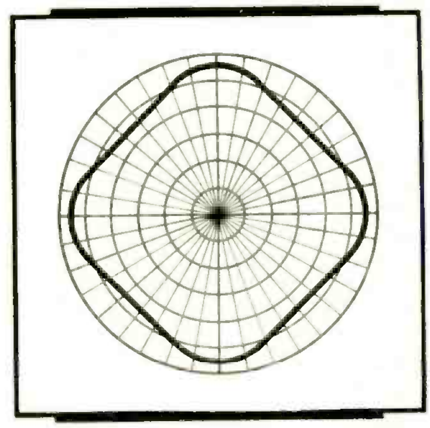

Figure 4

fed at the corners by a 400-600 ohm open line. For reception purposes only, the open line may be replaced by a coax and transformer assembly.

When a frequency band of from 2 to

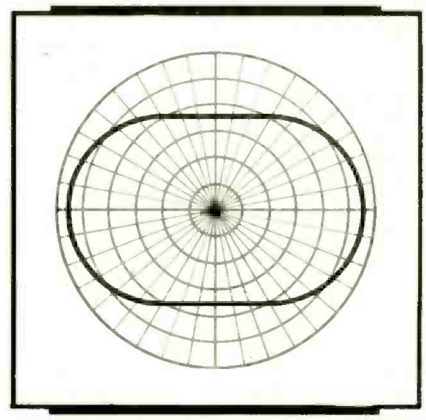

Figure 5

20 mc is to be covered, for example, the total frequency spread is divided between the four antennas: one antenna for 2-4 mc, another for 3.5-7 mc, a third for 6-12 mc, and the fourth for 10-20 mc. Poles for this assembly are set on

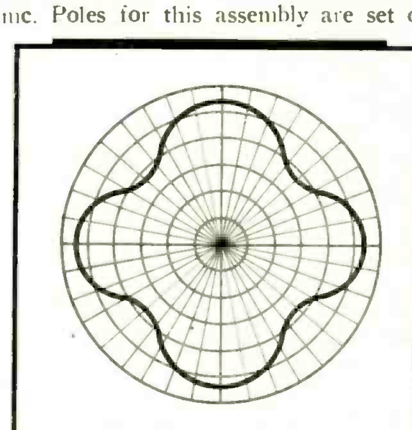

Figure 6

[*Continued on page* 22]

AMPHENOL

The Perfect Parallel Line Lead-in Wire

Amphenol Twin-Lead is a new type of radio frequency transmission line which combines the low cost of an open line with the excellent dielectric qualities of Polyethylene as a continuous spacer and insulator for the line. It is light and flexible — it can be tacked to a wall and is easy to lead in under a window sash. Its resistance to moisture, cold and heat is far superior to the usual rubber insulated, woven-braid-covered twisted pair used for antennas prior to the war.

Twin-Lead is made in three impedances that serve numerous applications. Selection of type is a simple matter. The *300 ohm* line is the most universal in use, particularly for FM and Television reception. Amateurs are using this line for both antenna and lead-in. The *150 ohm* type is excellent for antennas used mostly for short-wave broadcast reception, and is useful as a link between stages of a transmitter. The *75 ohm* line, originally designed for amateurs who operate in narrow bands of frequency, is also many times better for broadcast reception than the conventional rubber covered or cotton covered wire generally used.

It is to be emphasized that Amphenol Twin-Lead should not be thought of as exclusively for use at ultra-high frequencies. It is THE antenna lead-in for all frequencies.

AMERICAN PHENOLIC CORPORATION
CHICAGO 50, ILLINOIS
In Canada • Amphenol Limited • Toronto

ELECTRICAL DATA

300: Amphenol "Twin-Lead" Transmission Line is available in 300-ohm impedance value. RMA standardized on 300-ohm lead-in line for Television as the most efficient over broadband operation.

150: Amphenol also supplies 150-ohm twin-lead to those interested in particular applications and experimental work.

75: Designed especially for amateurs who operate in very narrow bands of frequency or one particular frequency. Ideal for dipoles with a nominal impedance of 72 ohms at the frequency for which they are cut. This line is also excellent for broadcast reception.

Dielectric constant of Polyethylene — 2.29. Capacities (mmf per ft.): "300" — 5.8; "150" — 10; "75" — 19.

Velocity of propagation (approximately): "300" — 82%; "150" — 77%; "75" — 69%.

Power factor of Polyethylene — up to 1000 Mc — .0003 to .00045.

ATTENUATION — FM AND TELEVISION BAND

Megacycles	300-ohm DB per 100 Ft.	150-ohm DB per 100 Ft.	75-ohm DB per 100 Ft.
25	0.77	0.9	1.7
30	0.88	1.03	2.0
40	1.1	1.3	2.5
60	1.45	1.8	3.4
80	1.8	2.25	4.3
100	2.1	2.7	5.0
200	3.6	4.7	8.3

Depend upon AMPHENOL Quality

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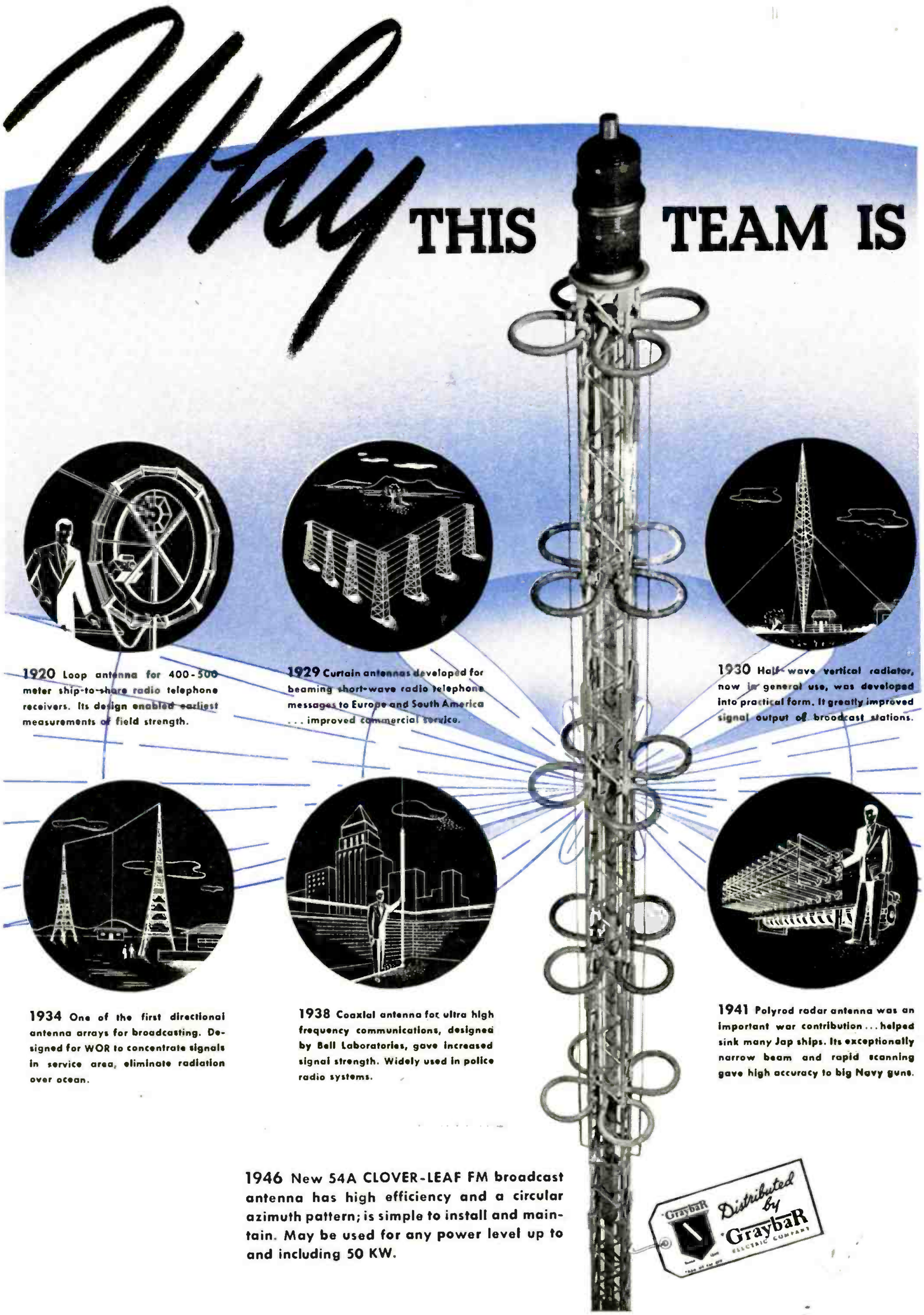
Why THIS TEAM IS
1920 Loop antenna for 400-500 meter ship-to-shore radio telephone receivers. Its design enabled earliest measurements of field strength.
1929 Curtain antennas developed for beaming short-wave radio telephone messages to Europe and South America . . . improved commercial service.
1930 Half-wave vertical radiator, now in general use, was developed into practical form. It greatly improved signal output of broadcast stations.
1934 One of the first directional antenna arrays for broadcasting. Designed for WOR to concentrate signals in service area, eliminate radiation over ocean.
1938 Coaxial antenna for ultra high frequency communications, designed by Bell Laboratories, gave increased signal strength. Widely used in police radio systems.
1941 Polyrod radar antenna was an important war contribution . . . helped sink many Jap ships. Its exceptionally narrow beam and rapid scanning gave high accuracy to big Navy guns.
1946 New 54A CLOVER-LEAF FM broadcast antenna has high efficiency and a circular azimuth pattern; is simple to install and maintain. May be used for any power level up to and including 50 KW.
Graybar
Distributed by
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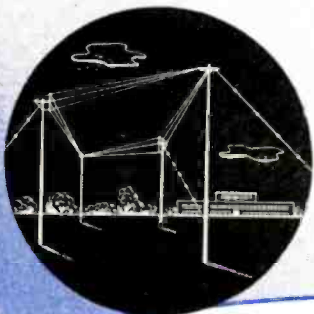
ELECTRIC COMPANY

Up ON ANTENNAS

As pioneers and leaders in radio, Bell Telephone Laboratories and Western Electric have been vitally concerned with the development of improved antennas for more than 30 years.

From the long-wave days of radio's youth, right through to today with its microwaves, this team has been responsible for much of the progress in antenna design.

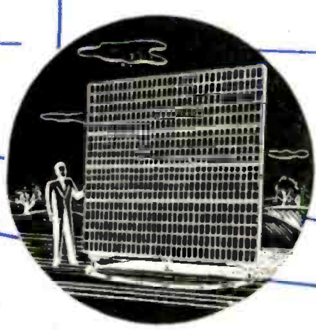


1930 Rhombic (diamond-shaped) antenna for 14-60 meters. It covers wide frequency range without adjustment. Still standard for this band.

Progress based on Research

Following their long-established method of attack, Bell Laboratories scientists are continually *observing*, *investigating* and *measuring* the action of radio waves in space. Their research has covered wave lengths ranging from hundreds of meters to a fraction of a centimeter. In over a quarter-century of intensive study, they have learned how radio waves behave, day and night, under all sorts of weather conditions.

Out of this fundamental research have come such outstanding developments as the rhombic antenna, musa antenna, vertical half-wave radiator, curtain antenna, directional array, the polyrod and other improved radar antennas, the metal lens for microwaves and the new CLOVER-LEAF antenna for FM broadcasting.



1944 Metal lenses, another Bell Laboratories development, focus microwaves like light. One type has a beam width of only 0.1°—or less than that of a big searchlight.

What this means to YOU

Whether you are interested in AM or FM—equipment for broadcasting, point-to-point, aviation, mobile or marine use—here's the thing to remember. Every item of radio apparatus designed by Bell Laboratories and made by Western Electric is backed by just such thorough scientific research as has been given to antennas. It's designed right and made right to give you years of high quality, efficient, trouble-free service.

BELL TELEPHONE LABORATORIES

World's largest organization devoted exclusively to research and development in all phases of electrical communications.

Western Electric

Manufacturing unit of the Bell System and the nation's largest producer of communications equipment.

FM BY FEDERAL

COVERS THE FIELD FROM ANTENNA DOWN!

Federal can provide your new FM station with the finest transmission equipment available—complete in every detail, from microphone to antenna. This outstanding "one-source" service means completely matched components for the entire system—all precision engineered, all of the highest quality, all designed to work together as a single perfected and coordinated FM system.

8 SQUARE-LOOP FM ANTENNA WITH NOMINAL POWER GAIN OF 9!

This remarkable new antenna, with the highest power gain ever available in FM service, provides *90-kw* effective output with a *10-kw* transmitter . . . *180-kw* with a *20-kw* transmitter . . . *450-kw* with a *50-kw* transmitter.

Antenna is built for use over entire FM range—from 88 to 108 mc. A single adjustment per loop tunes for any frequency in this band. Coaxially-fed loops radiate power in every direction of the *horizontal* plane, with very little power lost to sky or ground waves. Complete antenna and rugged supporting tower designed to withstand high wind velocities and heavy icing loads. Blaw Knox, Ideco, Wincharger towers available to mount the FTR Square-Loop array.

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WITH THE NEW "FREQUEMATIC"* MODULATOR

The new "FREQUEMATIC" modulator—an exclusive feature of Federal's complete line of FM transmitters—assures outstanding fidelity and mean carrier stability, with unsurpassed dependability and economy.

• Maintains center-frequency stability within .001% • Signal-to-noise ratio reduced to 5600-to-1 • Linear modulation of all signals between 50 and 15000 cycles • Uses simple all-electronic circuits with standard receiver tubes • Extreme ease of initial alignment and minimum maintenance.

Federal gives you complete service, too—providing factory-trained engineers to supervise installation, tune the equipment, and instruct your personnel in its operation and maintenance—all without extra charge.

NEW POWER TRIODES, developed by Federal, especially for FM service, are used in the power amplifier stages of the Transmitter, contributing to long life and stable operation.

* Trade mark

For information on Federal's complete FM equipment, write: Federal Telephone and Radio Corporation, Newark 1, New Jersey.

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An Improved Method of Testing Loop Receivers

W. J. POLYDOROFF
Consulting Engineer

The author describes a method of testing loop receivers which overcomes difficulties associated with previously used techniques. The procedure discussed has been adopted by many important factors in the industry

IN MY RECENT ARTICLE* the "antenna effect" which predominates in unbalanced or unshielded loops, has been described and its harm emphasized.

Although this effect has been discussed before,** practical methods of measurements of loop reception did not permit the evaluation of this and other loop properties.

When the loop is applied to broadcast reception the engineer is guided by R.M.A. technical bulletin #3 of Dec. 18, 1940, which describes the technique developed by the Hazeltine Corporation. The recommended method makes use of induction between two loops. A standard shielded loop of 10″ diameter is connected to the output of a signal generator (*Fig. 1*) and acts as a radiator.

The distance between two loops determines the amount of signal pick-up by a receiving loop in terms of "figure of merit" which, in terminology adopted by the British is called the "pick-up factor of the loop" and is numerically equal to

$$h_{eff} \times Q$$

We have so many "figure of merit" expressions that one more simply confuses the terminology.

The procedure described in the above bulletin applies only to:

1. High-impedance loops (directly tunable), although many receivers use low-impedance coupling.
2. The physical size of the loop under test *must be substantially* that of the standard radiator loop. Large errors are experienced if loops differ in size or if they are not placed on common centers. (This error is particularly noticeable with large console, or small iron core loops.)
3. The distortion of induction field due to the walls of a screened room is very pronounced.
4. Examination of induction pattern practically excludes the possibility of determining null point. If two loops are placed in "minimum" position, the combined directivity of the loops is measured.
5. The method is costly inasmuch as it involves extra equipment; a standard loop and a low impedance laboratory precision signal generator, although the results are far from precision.
6. The formulae and methods require several different loops for different frequency ranges.

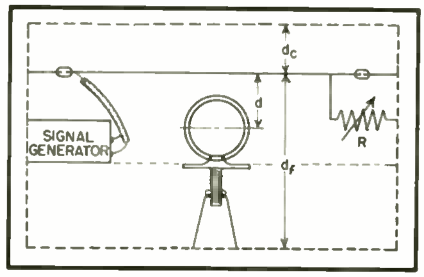


Fig. 3. Setup for improved method of testing loop receivers.

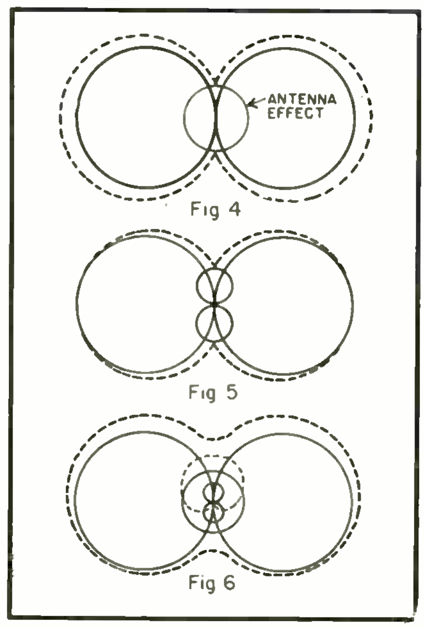


Fig. 4. Vertical antenna effect and figure-eight pattern of loop. Fig. 5. Displacement current may be eliminated by juxtaposition of turns, as indicated in the pattern shown. Fig. 6. Displacement currents produce a small figure-eight pattern at right angles to the main pattern. If an antenna effect is present, the polar diagram is further distorted, as shown.

7. The recommendation to shield the standard loop is merely a precaution against direct electrostatic pick-up (which is negligible at a distance of 20 inches) but does not exclude pick-up due to antenna effect from the proper loop pick-

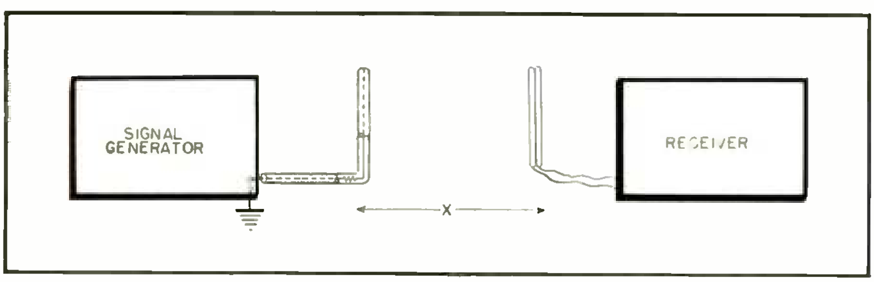
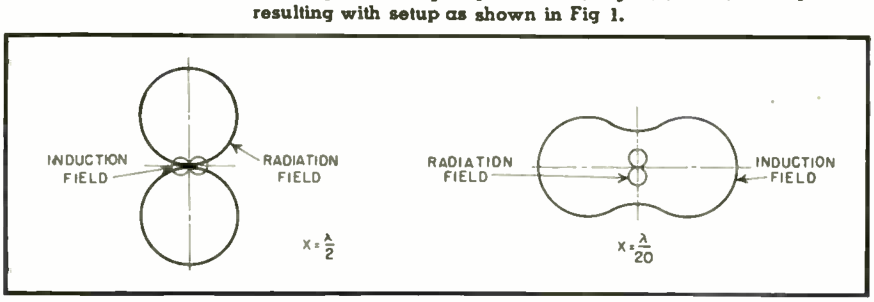


Fig. 1. (above) Conventional setup for testing loop receivers. Fig. 2. (below) Field patterns resulting with setup as shown in Fig 1.



*Inductively Tuned Loop Circuits, W. J. Polydoroff, RADIO, April, May, 1946.
***Journal Frank. Inst.*, Vol. 188. pp. 289-362.

up. There are no provisions to differentiate either type of pick-up. Before the war a new testing method was developed by the Army and adopted by most aircraft equipment makers, the Air Force, Navy, in fact, by all engineers who are interested in the proper testing of the loop under laboratory conditions approximating actual use.

This new simple equipment, already installed in many screened rooms, consists of a piece of wire and a resistor and is adaptable to any kind of signal generator (including servicemen's high impedance instruments).

Transmission Line (Test Line) Method

The method initiated by the U. S. Army has been described in two recent books,* and requires a screened room. *Fig. 3* shows a front view. Along its longest dimension, preferably in the center of the room, a copper wire transmission line, 10-16 gauge, is strung on two insulators. The characteristic impedance of such a transmission line is

$$Z = 138 \log_{10} \frac{4\,d_c}{a} \text{ (ohms)}$$

where d_c is the distance between the line and the ceiling of the screen room and a is the diameter of wire, both being in inches. In practice Z is of the order of several hundred ohms. Numerical example: $d_c = 8''$; wire #12, $d = .1''$; $\log_{10} 320 = 2.5$; $Z = 350$ ohms.

One end of the line is connected to the signal generator through a *shielded* cable. The other end terminates with a resistance, R, to ground (wall of the room). In order to prevent formation of standing waves in the line it should be terminated with $R = Z$, (350 ohms, in our example).

In order to find a proper termination, the transmission line is connected to a signal generator which is set at its highest frequency, 20-50 mc, at full output. A high-frequency vacuum tube voltmeter probe slides along the line from left to right. If there are different readings at different points of the line, the line is not properly terminated and the resistance should be adjusted. It pays to employ a variable 1000-ohm carbon resistor which is easily adjusted until the nodes disappear. If a sensitive voltmeter is not available, a high frequency receiver can be used with its antenna wrapped once around a small doughnut insulator which slides on the transmission line. Observe receiver output and adjust the variable resistance until the output is uniform from end to end of the line. After thus equalizing the output of the line at the highest operating frequency we may safely assume that its output may be uniform at lower frequencies.

*Aeronautical Radio Engineering, by P. S. Sandretto, pp. 140-145. Radio Direction Finders, Bond, pp. 221-231.

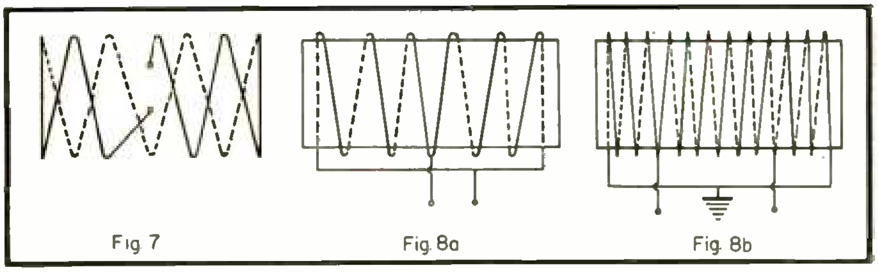


Fig. 7. Juxtaposition method of winding to eliminate displacement current. Fig. 8. (a) Displacement current may be balanced out by using two windings in parallel, as shown. (b) When the outer turns are brought to ground potential, as shown, balancing is still more effective.

The radiation of such a transmission line simulates the field of a *horizontally propagated vertically polarized radio wave.*

In such case, the field strength in microvolts per meter in the center of the loop is

$$E = 2{,}360 \frac{E_L}{Z} \left\{ \frac{1}{d} + \frac{1}{2\,d_f - d} - \frac{1}{2\,d_c + d} \right\}$$

Where E_L is the line voltage, numerically equal to the voltage of signal generator in microvolts; Z in ohms is already described; d, d_f and d_c in inches are the dimensions, respectively, line to the center of the loop, line to the floor and line to the ceiling of the room. For a numerical example we may take $d_c = 10''$; $d_f = 80''$ and $d = 20''$. In that case

$$E = E_L \times \frac{2360}{350} (.05 + .007 - .025)$$

$$= E_L \times \frac{2360 \times .032}{350} = E_L \times .2 \text{ or } E_L/5$$

In this example field strength at the center of the loop is equal to the signal output of the SG divided by 5. One can work out in advance several such round figure reductions and set the center of the loop at one of the predetermined distances from the transmission line.

In the above formula the signal generator's impedance is assumed to be of the order of several ohms as compared with Z, which is much larger, and $E_L = E_g$. With a high impedance signal generator output in the above formulae $Z = Z_g + Z_L$, where Z_g is the impedance of the generator. Obviously, if $Z_g = 350$ ohms the voltage delivered to the line will be half the voltage of the previous case of a low impedance SG.

The loop can be placed on a rotatable table graduated in degrees with a stationary index to facilitate observations of polar diagrams. In elaborate installations the bottom part of a barber's chair is used to adjust the loop to the desired height, but an ordinary swivel chair's bottom has been used to advantage.

In the case of a high-impedance loop receiver, the whole receiver could be placed on top of the table (the console set will of course set on the floor). In

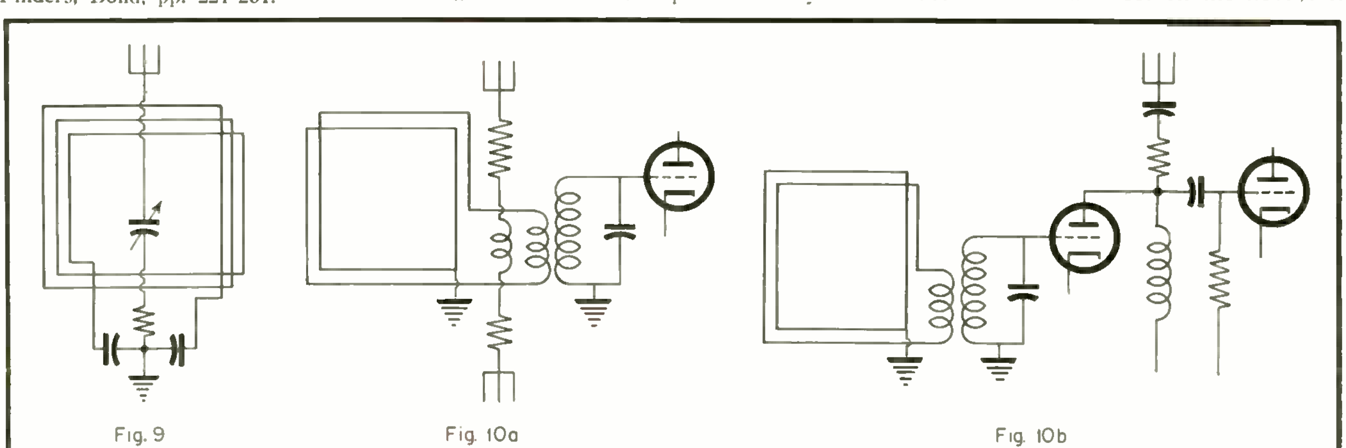


Fig. 9. For sense determination, antenna effect is deliberately introduced in the manner indicated. Fig. 10. Methods of introducing antenna effect in low-impedance loops.

low-impedance loop systems where the loop is connected to a receiver by a shielded cable the receiver may stay on the bench.

Measurement Procedure

Set the loop at maximum signal position, corresponding to the position shown in *Fig. 3*, and adjust receiver signal output, using 30% modulation. Turn off the modulation and measure the noise level in the output meter. If sensitivity at a signal-to-noise ratio of 4-1 is required and the noise is more than one-fourth the signal plus noise of previous reading, *increase* output of SG and *decrease* volume control to get same output. Again turn off the modulation and observe noise output. After 2 or 3 trials one will arrive at a signal and volume level which will produce maximum sensitivity at 4-1 signal-to-noise ratio.

Measuring Directivity

To measure the directivity or degree of attenuation at null points, rotate the loop at right angles to the maximum signal position and increase signal output of SG in decimal steps. Readjust the loop position until output meter reads minimum. Again increase output of SG and readjust the loop position. When final position is found adjust signal generator to read standard output (signal+noise at 30% modulation). The ratio of signal generator readings at maximum and minimum positions is the measure of directivity, usually expressed in db.

If the loop is carefully balanced and shielded, and all the leads are shielded and grounded, the antenna effect for all practical purposes is entirely excluded and directivity reaches the order of 50-60 db. In some cases directivity of the order of 100 db was observed; at such a degree of attenuation even a slight change in operator's position affects the measurements. The final turning of the loop should be by remote control as the approach of the hand to the loop produces tremendously strong signals. Once more we can observe the evidence of "human antenna" when signal from the line is re-radiated from the body of a person at a different angle if that person has moved his position after the balance.

Reciprocity

In a well-balanced and symmetrical antenna two null points should be 180° apart. In a large screen room this is true within half a degree. In a small room and especially when line is not symmetrical to the wall the reciprocity may be in error up to several degrees.

Effects of Loop Construction

The ideal loop in which all the wires are simultaneously hit by the wave is

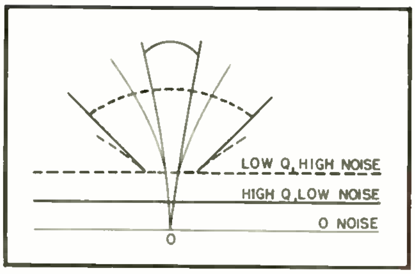


Fig. 13. A high-Q loop with low noise level produces a sharper null than a low-Q loop with high noise level.

a spiral or pancake type. Theoretically it should produce a true figure-eight pattern.

The inner circle shown in *Fig. 4* represents the vertical antenna effect which, when added to the figure-eight pattern, distorts the pattern as shown by dotted lines and the minima are masked, showing the amount of antenna effect. In many broadcast receivers the polar diagram is distorted to such extent that the ratio between maximum and minimum is less than 10 db.

Incidentally, a flat-type antenna is hard to balance by a grounded center tap alone because the electrical balance does not coincide with the electrostatic balance. In this respect solenoidal loops are preferable because a true center tap can be easily found (the end turns are at equal potential difference to the center tap). Such loops, however, have another kind of distortion due to displacement currents. Displacement currents produce a small figure-eight pattern at right angles to the main pattern, and if an antenna effect is present the polar diagram is further distorted, as shown in *Figs. 5* and *6*.

By juxtaposition of the turns, displacement current may be eliminated. This method is described in U. S. Patent #1,747,262. See *Fig. 7*.

By juxtaposition of the turns, dis- in opposition (*Fig. 8*), the displacement currents are balanced out; the second method (*Fig. 8b*) brings the

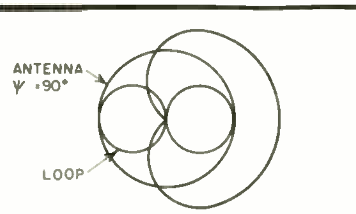


Fig 11

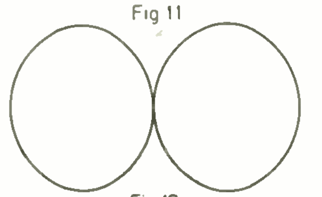


Fig 12

Fig. 11. Cardioid pattern resulting from circuit of Fig. 10b.
Fig. 12. Polar diagram of German iron-core loop.

outer turns to ground potential, making the balancing still more effective.

Generally speaking, balancing the loop by a center tap and the shielding of the loop entirely eliminates the antenna effect and greatly contributes to better loop reception.

There are the cases, however, where antenna effects are deliberately introduced in order to provide uni-directional reception essential for sense determination. To do this the vertical antenna signal is introduced at 90° phase difference to the loop signal. This can be done by the orthodox method shown in *Fig. 9*, or in case of low-pedance loops by one of two ways shown in *Figs. 10* and *11*. In *Fig. 10*, a symmetrical dipole antenna is connected to the coupling transformer, and in *Fig. 11* the output load of a tube is made reactive (so-called choke-amplifier) which produces a 90° phase shift of the loop signal, at which point a signal from the vertical antenna is injected. This results in a pattern of a so-called "cardioid". Such a pattern, while slightly reducing the signal strength at maxima, considerably sharpens the null points versus degree of rotation.

In conclusion we may also consider the influence of the Q of the loop on the directivity. Again theoretically the figure-eight pattern holds true. However, we already know that the signal-to-noise ratio is affected by Q: The lower the Q, the higher will be the noise. So if we draw a part of the polar diagram in its near zero region at a certain signal strength and draw two reference lines for noise levels (*Fig. 13*), we shall see a high Q loop with low noise level produces better aural null determination than a low Q loop with inherent high noise level, frequently used in direction finding apparatus which simultaneously indicates direction and the sense of the signal.

All of the above described patterns can be easily studied and observed when using the method and equipment above described.

Theoretically the shape of the loop, whether round or rectangular, should not affect the polar diagram pattern. The computation of signal in the square loop differs from the round one by a correction:

$$E_{av} = E \frac{d}{2L} \log_e \frac{d+L}{d-L}$$

where a square loop has sides in inches equal to $2L$.

The theory which derives figure eight pattern assumes that the loop has no axial dimensions.

In elongated (solenoidal) loops this does not hold true. Instead of the field strength gradient working in the *plane* of the loop we now have the same

[*Continued on page 36*]

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An Analysis of Cascode Coupling

R. G. MIDDLETON

Operation of individual tubes and input-output relations of the cascode amplifier may be determined by means of a graphical analysis on the plate family curves.

Cascode coupling as shown in *Fig. 1* affords a unique, direct-coupled high-amplification circuit which avoids the phase reversal encountered in a cascode amplifier. It will be observed that the cathode of the output stage is tied to the plate of the input stage, with both electrodes floating. Signal voltage e_i is applied between both grids and the cathode of V_1. Output voltage iR_L is developed across the load resistance R_L.

Circuit Variations

Various modifications of the basic circuit include the use of R-C grid coupling for a-c operation, with individual grid biases to each grid through grid leaks; bias cells may be included in either grid branch of the basic circuit to alter the operating points; self-bias may be used in the cathode lead of V_1; a-c coupling and shunt feed may be used in the output branch. The cascode circuit shown in *Fig. 3* operates with negative bias on the grid of V_1 and positive bias on the grid of V_2. The latter compensates for the plate drop through V_1 under no-signal conditions.

Graphical Analysis

Plate characteristics are shown in *Fig. 2* for a 6C5 tube, with circuit voltages indicated for an assumed plate current of 2 ma. A tube current of 2 ma results in a 100-volt drop across the plate load resistor R_L, noted on *Fig. 2* as i_1R_L. The tubes have been chosen as similar types, and the one set of characteristics applies to both. Current i_1 is common to both tubes, and under equilibrium conditions e_{c1} and e_{c2} both lie on the i_1 ordinate.

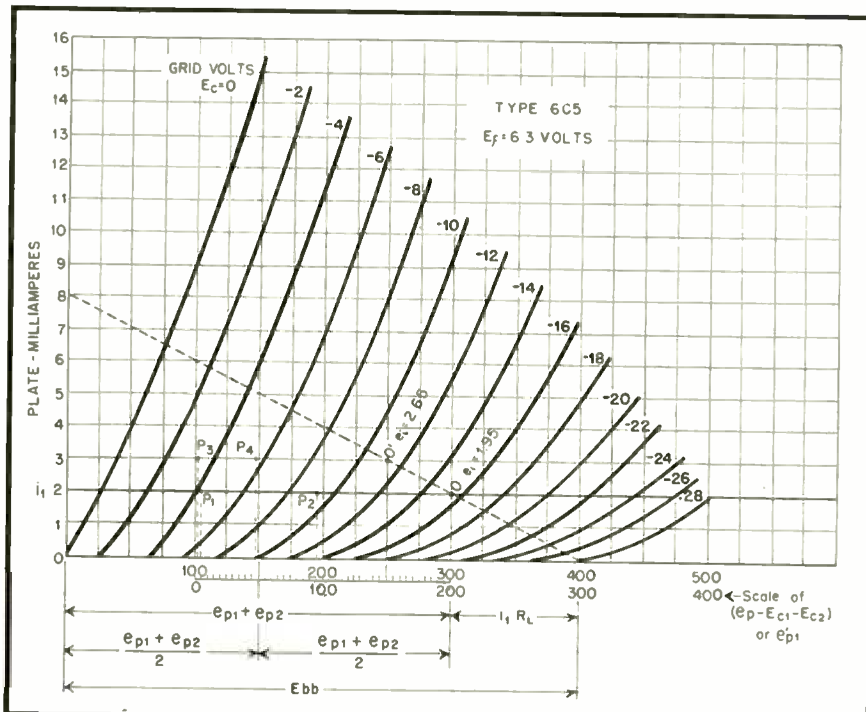


Fig. 2. Graphical relations which yield terminal characteristics of the individual cascode tube loads, and the terminal characteristic of the output resistance, with input voltage calibrations.

The values of e_{c1} and e_{c2} may be readily determined for the assumed tube current i_1 and the bias voltages E_{c1} and E_{c2} from the following considerations: Applying Kirchhoff's law to the plate branch,

$$E_{bb} = i_1R_L + e_{p1} + e_{p2}$$

or

$$e_{p1} + e_{p2} = E_{bb} - i_1R_L$$

Since the plate supply voltage was chosen as 400 volts, and the i_1R_L drop equals 100 volts, the sum of the unknown plate voltages is necessarily equal to 300 volts, as noted on *Fig. 2*. Furthermore, e_{p1} lies as far to the left of the median (150 volts) as e_{p2} lies to the right. This follows from elementary geometrical considerations; when $e_{p1} = e_{p2} = 150$, both terminate at the median voltage, but when e_{p1} decreases by a certain amount, e_{p2} increases by the same amount to maintain the sum constant as required by Kirchhoff's law.

Applying Kirchhoff's law to the grid side of the circuit, it is seen that $e_{c2} - e_{c1} = E_{c2} - E_{c1} - e_{p1}$, or noting the values and polarities of grid biases, $e_{c1} - e_{c2} = e_{p1} - 100$. It is not necessary, but when desired, a direct graphical reading of the right-hand side of the equation may be obtained by displacing the e_p scale to the right by an amount of the constant (in this case 100 volts), as shown by e'_{p1} in *Fig. 2*.

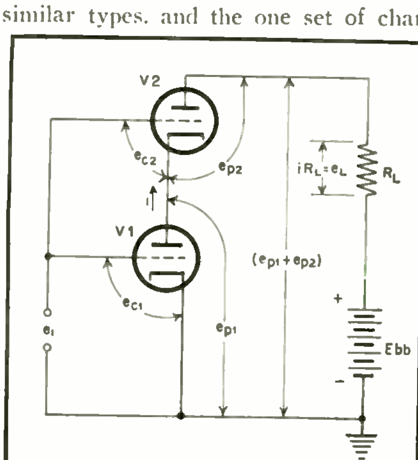


Fig. 1. Basic circuit of cascode amplifier, with d-c coupling as used in voltage regulating circuits.

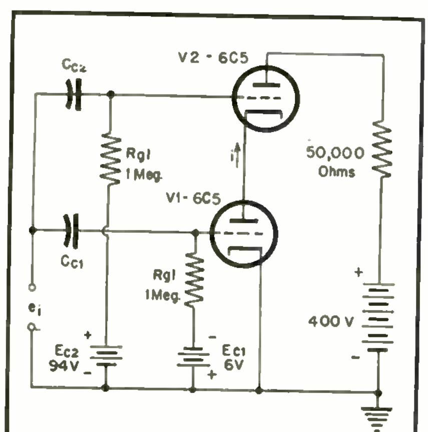


Fig. 3. Cascode circuit adapted for alternate current coupling, and with individual grid biases

[*Continued on page 32*]

DESIGN DATA AND CHARACTERISTICS OF HIGH-FREQUENCY CABLES

KARL ZIMMERMANN
Federal Telephone and Radio Corp.

New design factors and manufacturing methods to meet the rigid mechanical and electrical specifications required in modern applications of high-frequency cables are described

As previously indicated, a stranded conductor increases the cable attenuation unless very fine strands are used, which is impracticable from both a design and a manufacturing viewpoint. It is generally agreed that at u.h.f. and v.h.f. the conductor with the minimum periphery is the best design, and when it is desirable to use a stranded conductor to achieve increased flexibility, a few large strands are preferable to many fine ones.[7] Actually a seven strand conductor is almost always used, in which case R_1 has been experimentally determined to be 1.25 at about 1 mc.[5,6] However, this value undoubtedly increases somewhat with frequency.

Another variation in the strand resistance occurs after the cable is flexed; the contacts between strands change and if the copper is dirty or corroded the resistance may vary considerably. For cables where the attenuation must be kept low and constant it is frequently advisable to silver-plate the inner conductor, which not only reduces the attenuation but also keeps the resistance more constant. This is true whether a stranded or a solid inner conductor is used.

Outer Conductor—Braiding

As the reader probably knows, the outer conductor is very rarely if ever solid tubular in cables. Actually a braid is used as the outer conductor in virtually all transmission lines where flexibility is desired. The attenuation due to the braid, R_2, has been calculated to be[5]:

$$R_2 = R_m \times R_b \qquad (25)$$

where

$$R_m = \frac{\text{Resistance of outer conductor metal}}{\text{Resistance of copper}}$$

$$R_b = \frac{\text{Resistance due to braid}}{\text{Resistance due to tubular}}$$

and R_b at 1 mc is

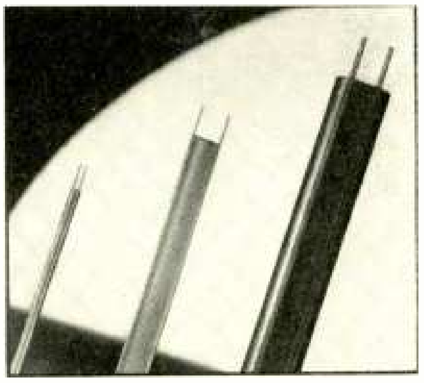
$$R_b = L^2/F \qquad (26)$$

where

L = length of braid wire per unit length of cable
F = coverage of the braid

According to this relation, the attenuation increases with length of braid, and decreases with increase in braid coverage. The limiting position of the braid wires would be, for minimum braid loss, in a lay parallel to the cable axis and closely packed to obtain 100 per cent coverage. However the cable gets stiffer as the length of braid wire per unit length of cable decreases (L should not be confused with "braid lay" which is the length of cable covered by one revolution of braid wire, so as L decreases, the braid lay increases). A compromise between electrical and mechanical requirements must therefore be made and as a result of this compromise a fairly short lay[7] is used.

CONCLUSION



Open-wire line represents the low-cost trend; designed for f-m and t-v lead-in use, with characteristic impedances of 100, 200 and 300 ohms. Attenuation of 200-ohm line is 0.4 db/100' at 30 mc and 0.82 db/100' at 100 mc; 300-ohm, 0.85 db/100' at 30 mc, 2 db/100' at 100 mc.

—Courtesy American Phenolic Corp.

Dielectric Loss

The second part of equation (18), $\frac{1}{2}G(L/C)^{-\frac{1}{2}}$ represents the power loss due to the dielectric. $(L/C)^{\frac{1}{2}}$, of course, is Z_o which may be written in the form

$$Z_o = (1016\ k^{\frac{1}{2}}\ 10^{-12}/C)$$

where

C = capacitance of cable in farads per foot
k = dielectric constant of dielectric; 2.25 for polyethylene

and $\quad G = \omega C P$

where

$\omega = 6.28\ f$, the frequency in cycles
P = power factor of the dielectric; 3×10^{-4} for polyethylene

Therefore the dielectric loss of a polyethylene insulated cable in db/100 feet is

$$\alpha_c = (4.34\ \omega \phi P \cdot 1016 k^{\frac{1}{2}} \cdot 10^{-11}/\phi) = 12.6 \times 10^{-4} f \text{ db/100 ft} \qquad (27)$$

where f is in megacycles

Jacketing Materials

Jacketing material is dictated mainly by physical requirements. However, when a constant, low-attenuation transmission line is desired the electrical properties of the jacketing compound must be considered; for under certain conditions, discussed in full detail in an article by Mr. Warner,[9] the plasticizer in this compound will migrate into the polyethylene, causing a rise in its power factor and subsequently an increase in attenuation of the cable. This can be overcome by use of a "non-contaminating" jacket, such as IX-102, which was developed by Federal Telephone and Radio Corporation.

Summarizing the total attenuation of a coaxial transmission line:

$$\alpha = (0.434 f^{\frac{1}{2}}/Z_o)(R_1/d + R_2/d) + (2.77\ P\ k^{\frac{1}{2}} f) \text{ db/100 ft} \qquad (28)$$

Attenuation of Balanced Lines

Calculation of attenuation of shielded balanced transmission lines is a much more complicated problem due mainly to the addition of one element and the proximity of the parallel conductors. As in the computation of the characteristic impedance only approximate results can be obtained which can be used for the design of an experimental cable.

The attenuation due to the conductors of a shielded pair of balanced lines (see *Fig. 8*) is given as[5]

$$\alpha_c = [(3.1 \cdot 10^{-3} k^{\frac{1}{2}} f^{\frac{1}{2}})(R_1 R_p/d + 4\rho^2 R_2/d)]/[\{\log_{10} 2V(1-\rho^2)\}/\{1+\rho^2\}] \text{ db/100 ft} \qquad (29)$$

where

k = dielectric constant of the insulation
f = frequency in megacycles per second
$V = 2h/D$; $h = \frac{1}{2}$ the separation of conductors in inches (*Fig. 8*)
$\rho = 2h/D$

$$R_1 = \frac{\text{Resistance parallel conductor metal}}{\text{Resistance of copper}}$$

$$R_2 = \frac{\text{Resistance of outer conductor metal}}{\text{Resistance of copper}}$$

R_p = Proximity factor
d = Diameter of parallel conductors in inches

D = Diameter of inner surface of outer conductor in inches

This formula is of slightly more approximate form than that given in the paper by Green, Leibe, and Curtis.[9]

In the special case where no shielding is used $\rho = 0$ and equation (29) reduces to

$$\alpha = (R_1 R_p \cdot 3.1 \cdot 10^{-3} k^{1/2} f^{1/2}) / (d \cdot \log_{10}{}^2 V) \text{ db/100 ft.} \quad (30)$$

The formula for the proximity factor, R_p, at very high frequencies has been given by Carson[10] as

$$R_p = [V^4 - 2V^2(V^2-1)^{1/2} + 1] / [2V^2(V^2-1)^{1/2} - V^4 + 1] \quad (31)$$

Calculation of the attenuation due to all other parameters is the same as shown for coaxial lines. Specifically the dielectric loss, and R_2 follow the same equations.

Minimum Attenuation

As in the case of coaxial lines an optimum value of D/d, called p, and ρ exists which results in minimum attenuation. These values have been determined graphically[8] and are:

$$p = 5.4 \text{ and } \rho = 0.46$$

If an oval-shaped shield is used the attenuation is approximately 12 per cent less than an equivalent circular shield with equal cross-sectional area and the optimum values for minimum attenuation in this type of transmission line are[6]

$$p = 3.7 \text{ and } \rho = 0.47$$

where D and h are as defined on *Fig. 9*.

Power Rating

The power rating of a cable is a function of the maximum temperature that it can withstand. In u-h-f solid dielectric coaxial lines this power carrying capability is limited by (1) the rate at which the line can dissipate heat generated internally. (2) the rate at which the heat is generated internally. The first depends on the diameter (surface exposed), the color, texture, and material of the jacket, paint, armor, and the ambient temperature, among other factors. The second depends on internal dimensions, materials and frequency, among other factors.

The maximum temperature of a cable occurs at its inner conductor and is proportional to the well known I^2R loss and inversely proportional to the less known heat flow or the magnitude of the heat that is radiated from the conductor. Obviously if the conductor were perfectly insulated, and heat flow equal to zero, a continuous current would cause a continuously rising conductor temperature. On the other hand if there were infinite heat flow then the conductor temperature would remain constant regardless of what the I^2R loss is.

The theory of heat flow is similar to the theory of current flow in electric circuits as expressed by Ohm's Law. Temperature rise (potential) is equal to the product of heat flow (current) and thermal resistance. Thus the amount of heat that will radiate from the inner conductor will depend on the temperature gradient and the thermal resistances of the dielectric, braid, jacketing material and jacket surface to air.

The thermal resistance of a material is a function of its thermal resistivity and its dimensions. The recommended values for the thermal resistivity of polyethylene and black vinylite as determined empirically, using a center conductor temperature of 175°F, are[11]

K_d = 325°F./watt/inch/inch²
K_j = 850°F./watt/inch/inch²
H = 300°F./watt/inch/inch² for cables of 0.5 inch or less outside diameter
350°F./watt/inch/inch² for cables larger than 0.5 inch O. D.

where

K_d = thermal resistivity of polyethylene
K_j = thermal resistivity of black vinylite (jacketing material)
H = thermal resistivity of jacket surface to air

The relation between thermal resistivity and thermal resistance in h-f cable can be expressed as:

$$t_d = .3665\ K_d \log_{10} D/d = 110 \log_{10} D/d \quad (35)$$

where

t_d = thermal resistance of polyethylene
D and d are as defined in (7)

$$t_j = .3665\ K_j \log_{10} D_s/D_b = 314 \log_{10} D_s/d_b \quad (36)$$

where

t_j = thermal resistance of the jacketing material
D_s = overall diameter of the cable
d_b = diameter over the braid

for cable with D_s less than 0.5 inch

$$t_s = 0.32\ H/[(T_s - T_a)^{(a-1)} D_s] = 51.5/D_s$$

for cables with D_s larger than 0.5 inch (37)

Fig. 8 (left): Pair of balanced lines in round shield; Fig. 9 (right): Balanced pair in oval-shaped shield, with 12% less attenuation.

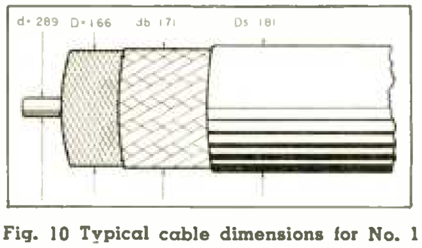


Fig. 10 Typical cable dimensions for No. 1 AWG inner conductor and 50-mil jacket.

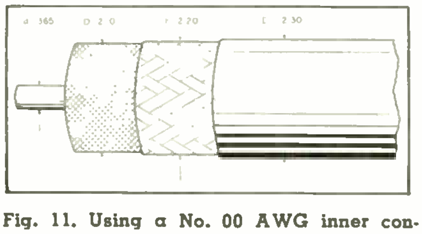


Fig. 11. Using a No. 00 AWG inner conductor the overall cable dimensions increase (see text).

where

T_s = temperature of the jacket in °F.
T_a = ambient temperature in °F.
a = a constant

$(T_s - T_a)^{a-1}$ = 1.85 when T_a = 104°F.
1.95 when T_a = 68°F.

From these values we obtain the rate at which the cable can dissipate heat generated internally. Now we must calculate the rate at which the heat is generated internally. This can be done by use of the following relations

$h_c = 2W\alpha_c$ = heat generated in the center conductor in watts/inch (38)
$h_b = 2W\alpha_b$ = heat generated in braid in watts/inch (39)
$h_d = 2W\alpha_d$ = heat generated in dielectric in watts/inch (40)
$W_h = 2W\alpha_t$ = total heat generated and dissipated in watts/inch (41)

where

W = total average power input watts
α_c = attenuation of the center conductor in nepers/inch (R_1 component in equation (28) divided by 1.04 × 10⁴ to convert from db/100 ft to nepers/inch).
α_b = attenuation of the braid (R_2 component of equation (28) divided by 1.04 × 10⁴) in nepers/inch.
α_D = attenuation due to the dielectric in nepers-inch (equation (27) divided by 1.04 × 10⁴).
α_t = total attenuation of the cable in nepers per inch (equation (28) divided by 1.04 × 10⁴).

If the thermal resistance and the heat generated in the material are known then its temperature rise can be expressed in terms of watts input by the following expressions:

Center conductor to braid—

$$T_c - T_b = t_d(\tfrac{1}{2}h_d + h_c) = t_d W(2\alpha_c + \alpha_D) \quad (42)$$

Braid to surface—

$$T_b - T_s = t_j W_h = 2t_j W\alpha_t \quad (43)$$

Surface to air—

$$T_s - T_a = t_s W_h = 2t_s W\alpha_t \quad (44)$$

therefore center conductor to air is equal to

$$(T_c - T_a) = W[t_d(2\alpha_c + \alpha_D) + 2\alpha_t(t_j + t_s)] \quad (45)$$

Equation (45) can be rewritten in the form

$$W = [T_c - T_a]/[t_d(2\alpha_c + \alpha_d) + 2\alpha_t(t_j + t_s)] \quad (46)$$

where

W = maximum average power input
T_c = maximum temperature that the cable can withstand

If the results must be accurate then the values used for attenuation should be corrected to include the effects of temperature rise and any mismatch that may occur.[12]

Typical Design Problem

The following design problem illustrates some of the typical calculations necessary to engineer a h-f cable:

Required: A 70-ohm coaxial line to transmit 100 megacycle. 10 kw average power at an ambient temperature of 104°F.

Solution: Since the characteristic impedance of the line must be 70 ohms, its D/d ratio is fixed and can be calculated from equation (7)

[Continued on page 32]

TECHNICANA

[*from page* 10]

the corners of a 270-foot square. A height of 100 feet has been found suitable, giving an effective height of about one wavelength at 30 meters, and of one-half wavelength at 60 meters.

The principle of operation is a superposition of two appropriate figure-eight patterns at right angles to each other and in space phase quadrature. The radiation pattern at medium frequency is shown in *Fig. 4*, with the deviation at lower and upper frequency limits of the 2-1 band depicted in *Figs. 5* and *6*, respectively.

HUM AND NEGATIVE FEEDBACK

★ Negative feedback is not a panacea for hum troubles, as discussed by an author who signs himself "Cathode Ray" in an article entitled "Negative Feedback and Hum" in the *Wireless World* for May, 1946. The author considers what happens to hum introduced by the power supply into the stage with feedback. To simplify the investigation, hum introduced by stray coupling is neglected. Results were obtained upon an analytical basis, and checked by experiment.

General conclusions are: Unless filtering is very good, feedback should not be taken from the plate when the load is transformer coupled. With triodes, parallel feed yields lowest hum across the load. When using tetrodes and pentodes, it is extremely desirable to use additional filtering in the screen supply.

The two typical circuits shown in *Figs. 7* and *8* gave the best performance on the basis of the investigation. Less satisfactory circuits are also presented by the author in his article. For *Fig. 7*, the hum voltage across the load in percentage of hum voltage of the source was found: without feedback (dotted line) 24%; with feedback, 6%. For *Fig. 8*, without feedback, 2%; with feedback 1%. However, when the screen

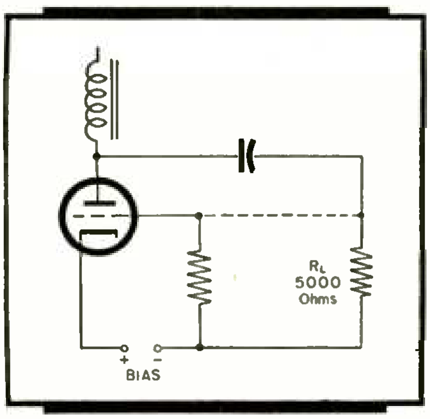


Figure 7

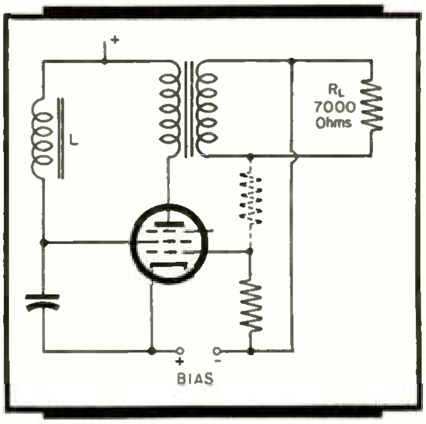


Figure 8

filter was omitted, these figures increased to 180% and 16%, respectively.

With regard to push-pull circuits, the author notes that the circuit affords hum cancellation, but that if carried to excess, the power supply ripple may modulate the signal. Feedback from the secondary of the output transformer is indicated for push-pull circuits.

ABSOLUTE BELS

★ Adoption of 1 watt as a reference power level with resulting power ratios expressed in the form of "Absolute Bels" is advocated by F. S. G. Scott in an article appearing in the *Wireless Engineer* for May, 1946.

The author points out that many reference levels exist, such as 1 mw in 600 ohms, 6 mw in 500 ohms, etc., which leads to difficulties of computation if not confusion when the worker makes measurements on various loads. He recommends having implicit in the bel term the absolute value of P_x by comparison with a universally recognized and accepted standard

$$\log_{10} P_x/P = N_{px}$$

where N_{px} is the value of P_x expressed in "absolute bels", achieving maximum simplicity by equating reference power P to unit. This equation may be rewritten

$$\log_{10} P_x - \log_{10} P = N_{px}$$

The author provides the definition: "When any power (P_x) is compared (in bels) with one watt, the resultant answer is expressed in absolute bels." (B = 10 db = 10 absolute decibels.)

$$N_{px} = \log_{10} P_x$$

The advantages of the one-watt reference level are summarized by the author: Many other parameters than power, such as dynes, lengths, resistances, volts, etc., which condition power under various circumstances, are expressible in "absolute bels" and afford a direct indication of how these parameters will affect the power in any circuit into which they may be inserted. Furthermore, the author points out, mental conversion of the ratio of two powers into db or a difference in phons is readily effected.

GRAPHICAL CALCULATION OF DOUBLE STUBS

ROBERT C. PAINE

Circle diagrams may be elaborated for rapid solution of double-stub transmission line matching problems

> ***Note:*** **In lines operating at VHF or UHF, the magnitude of series reactance is so much greater than series resistance, and likewise the magnitude of shunt susceptance is so much greater than shunt conductance, that R = G = 0 for all practical purposes. For this condition the general line equation becomes:**
>
> $$\frac{Z_{in}}{R_r} = \frac{(R_r/R_o) + j\tan 2\pi\,(l/\lambda)}{1+j(R_r/R_o)\tan 2\pi(l/\lambda)}$$
>
> **where Z_{in} = input impedance; R_o = characteristic resistance of line; R_r = resistance (or impedance) at receiving end; l = wavelengths of line under consideration; λ = operating wavelength. In this form, the left-hand side of the equation gives the input impedance in convenient dimensionless units. When the right-hand side of the equation is separated into real and quadrature components, the loci of Z_{in}/R_o plot as families of circles. This is the basis of the circle diagram discussed in Mr. Paine's article.**
>
> ***—Ed.***

TRANSMISSION LINE impedance-admittance charts for solving line matching problems have been described by the author in a previous issue.[1] Their use in computing the position and length of single stubs for matching the load to the line was described therein. These charts are presented again with additional curves drawn in for use in computing double stubs.

For the benefit of those readers who may not have read the original discussion, these charts are again briefly described. In a transmission line not terminated by its own characteristic impedance the input impedance varies as the length of the line increases. The same values are repeated in a cyclic manner every half wavelength, or 180°. Mismatch of the load also results in a variation along the line of effective voltage (and current) which also passes through a cycle of maximum and minimum values between points along the line separated by intervals of 180°. The ratio of maximum to minimum voltage (or current) values is designated here as Q. The variation in impedance is also a function of Q.

For any given value of Q the locus of impedance values is a circle. A chart of these circles for different values of Q is shown in *Fig. 1.* For this chart the characteristic impedance is taken at a normalized value of unity and the rectangular coordinates $r \pm jx = R/Z_o \pm jX/Z_o = Z/Z_o$, the actual apparent line impedance divided by the characteristic impedance, Z_o. The arcs designated in degrees represent distances along the line. The 0° and 180° arc represents a point of maximum voltage and impedance, while the 90° arc is a point of minimum voltage and impedance. *Fig. 2* is an enlarged detail of *Fig. 1* for more accurate reading in a region where the lines of the latter are crowded together.

Practical Example

To illustrate the reading of this chart consider a 100 ohm line terminated by a load of 155 $-j130$ ohms. This load appears as the point marked *"Z"* in *Fig. 1* at $R/Z_o = 1.55$ and $X/Z_o = -1.3$ on the circle $Q = 3$ at 20°. The impedance at any point nearer the power source is found by a clockwise motion around the Q circle, in this case the first point of pure resistance (minimum)

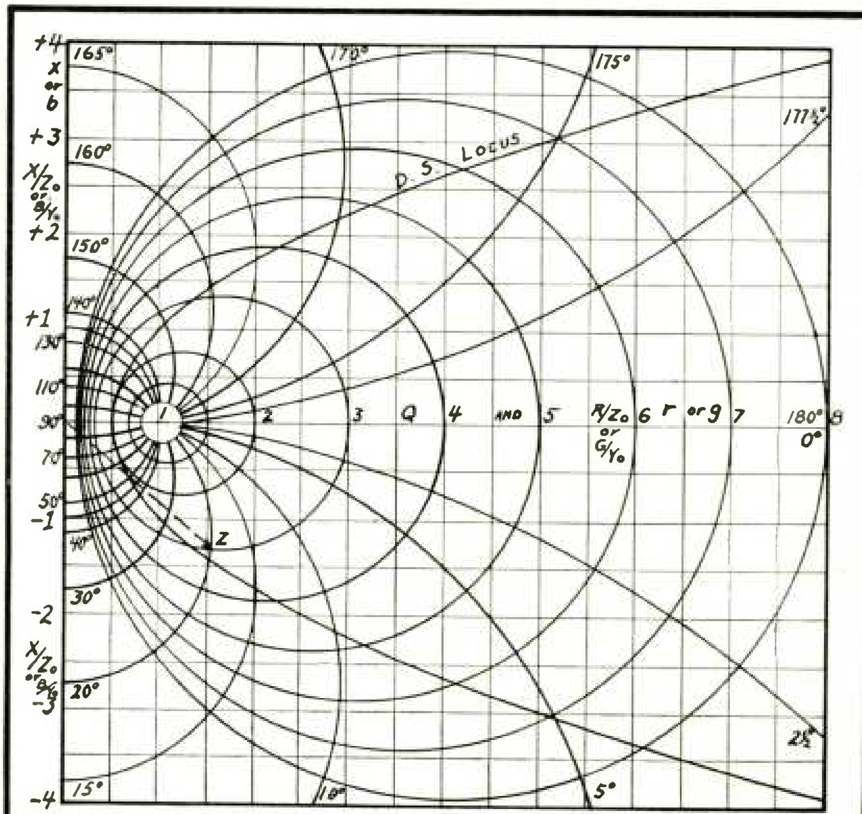


Fig. 1. Transmission line impedance-admittance matching chart with parabolis "D.S.Locus" for centers of double-stub admittance circles.

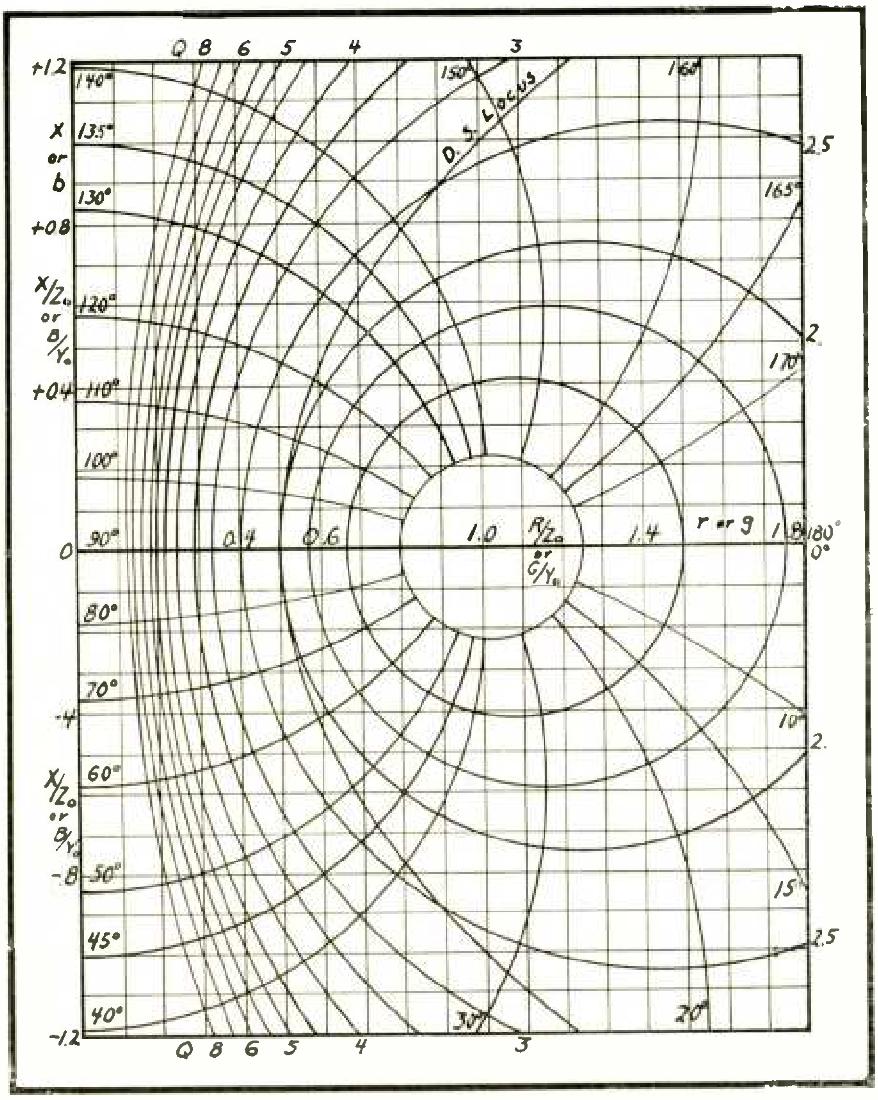


Fig. 2. Enlarged portion of Fig. 1 for more accurate readings.

is at $r = .33$ on the $Q = 3$ circle at 90°, a distance of (90° − 20°) or 70° from the load. The actual value of the line resistance at this point is .33 times 100, or 33 ohms. The next point of pure resistance maximum is 90° farther along the line, 160° from the load, and it is equal to 300 ohms.

For problems involving transmission line sections in parallel, as for the use of stubs, it is convenient to have a transmission line admittance chart in units of conductance, *g*, and susceptance, *b*. The charts of *Figs. 1* and *2* can also be used in this way, assuming the characteristic admittance, Y_0, to be unity, so that *g* and *b* of the chart respectively equal G/Y_0 and B/Y_0, the apparent conductance and susceptance of the line at any point divided by the characteristic admittance. In this case the arc 0° and 180° denotes a point on the line of maximum pure conductance on the admittance chart, corresponding to a point of minimum pure resistance at 90° in terms of the impedance chart. In the same manner, for any given value of impedance, the corresponding value of admittance can be found on the same Q circle at the angular arc 90° away. For example, take an impedance of $r - jx = .73 - j.53$ on the circle $Q = 2$ at 50°; the corresponding admittance is found as $g + jb = .9 + j.65$ on the same Q circle at 50° + 90°, or 140°.

Why Double Stubs Are Used

On transmission line installations it may not be convenient to move a stub. This is especially true of concentric lines where, due to the type of construction, stubs must be integrated as part of the whole design. In such cases double stubs are useful, since they can remain located at fixed points and, by adjustment of their effective length with sliding plugs or other convenient means, loads varying over a considerable range of values can be matched to a line.

The problem of double stub matching is illustrated in *Fig. 3*. The line with its two stubs shown in *Fig. 3a* may be considered as composed of three sections, (*3b*), the main line with an admittance, considered as unity, of $Y_0 = g + jb = 1 + j0$; (*3c*), a matching section of any given length, ϕ_2, in degrees of wavelength, the required characteristics of which are to be determined; and (*3d*), the load Y_L, either with or without a short section of line of length ϕ_1 to the point where the first stub is connected. When the first stub is connected at a distance of ϕ_1 from the load, the effect is the same as if a load of $Y_1 = g_1 + jb_1$ were connected directly at the stub.

A given length of ϕ_2, of matching section presents an input conductance of 1 (plus a susceptance b_3) to the main line when any of a certain range of loads, Y_2, is connected at its distant end. If the load as seen as Y_1 has a conductance, g_1, within this range, impedance matching can be accomplished at this point by adjusting a suitable stub at Y_1 to compensate for the difference between the susceptances b_1 and b_2. Impedance matching at Y_3 can be accomplished by adjusting the stub at this point

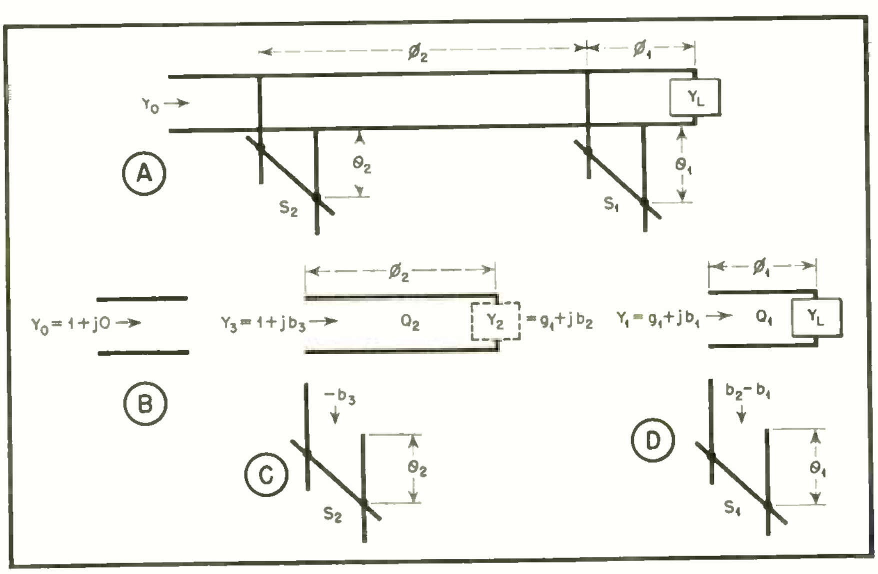


Fig. 3a. Diagram of transmission line impedance matching system with two fixed position adjustable stubs, for analysis divided into its elements. 3b., main line. 3c., matching section with stub S for compensating susceptance at input end. 3d., load section with stub S_1 for matching the susceptance of load to that of matching section.

to neutralize the susceptance of the matching section, seen here as b_s. Thus it is necessary to determine what admittance must be seen at Y_2 in order that unit conductance may be seen at Y_3. This value can be found by aid of the charts described above.

Chart Interpretation

The admittance needed to be seen at Y_2 for any given length of matching section ϕ_2 is multi-valued. It can be defined on the charts by a circle drawn tangent to the b axis at the terminus of the arc $(180° - \phi_2)$ and passing through the point $g = 1$, $b = 0$. The locus of the center of a series of such circles is a parabola, the equation of which is $b^2 = 2g - 1$. This parabola appears on both charts as the "D.S. Locus". It can be plotted from the equation and drawn in on any similar charts[2] if desired. *Fig. 4* shows a circle for a matching section ϕ_2 long. Its center is found at the intersection of the parabola with a horizontal line through the terminus of the arc $180° - \phi_2$. By beginning at an admittance represented by any point on the circumference of this circle and following any Q circle such as Q' or Q'' in a clockwise direction, the vertical line through $g = 1$ is reached in a distance of ϕ_2 degrees. For such a matching section the required load Y_1 must have an admittance equal to any of the values on the circumference of the ϕ_2 circle.

The center of the required admittance circle for a matching section ϕ_2 long can also be found by the geometric construction shown in *Fig. 5*. A perpendicular bisector is erected to the line, p_1, through the two points, $g = 1$, and $180° - \phi_2$ on the b axis. The center of the required circle then lies at the intersection of this bisector with the horizontal line pc, through the point on the b axis, $180 - \phi_2$.

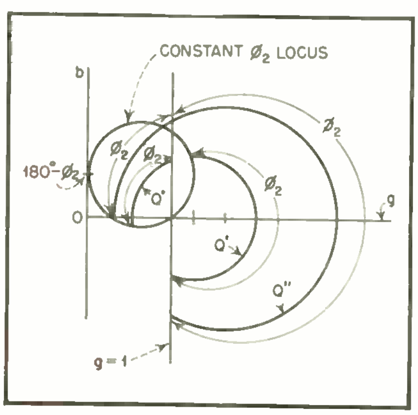


Fig. 4. Diagram showing how the circular locus of admittances for a constant length matching section, ϕ_2, can be drawn on the admittance chart of Fig. 1 for computing the value of double stubs.

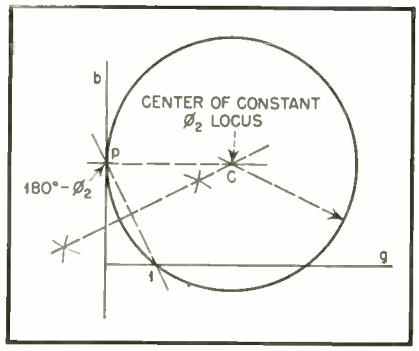


Fig. 5. Geometric method of locating the center of the admittance circle for a constant length matching section as in Fig. 4.

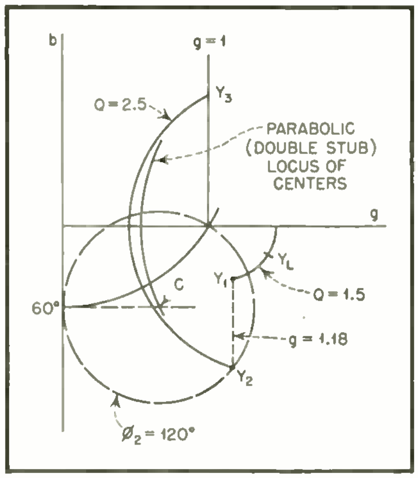


Fig. 6. Graphic solution of double stub matching, problem 1.

A Double-Stub Problem

Example 1. Given a load admittance $Y_L = 1.45 - j.2$ (in terms of a line admittance considered as unity), the first stub, S_1, at 18° from the load, and the second stub, S_2, at 120° from S_1, to find the length of the stubs. In chart 2 the load Y_L is spotted on the $Q = 1.5$ circle at the 10° arc. Following this Q circle for 18° the apparent load, Y_1, is found as $g_1 \pm jb_1 = 1.18 - j.4$ at the 28° arc. A horizontal line is then drawn through the vertical axis at (180° − 120°), or 60°, intersecting the double stub locus curve and determining the center, c, of the ϕ_2 circle, as shown in *Fig. 6*. With this point as center the $\phi_2 = 120°$ circular locus is drawn through the point $g = 1$, $b = 0$, intersecting the $g = 1.18$ vertical on the $Q = 2.5$ circle at Y_2 where $b_2 = -1.02$. This determines the susceptance of the S_1 stub as $b_2 - b_1 = -1.02 - (-.4) = -.62$. The $Q = 2.5$ circle is followed 120° to the $g = 1$ vertical to Y_3 where $b_3 = +.94$, then the susceptance of the S_2 stub is the negative of this value or −.94. From these values of stub susceptance the actual length of the stubs is then determined.

The stubs are constructed of low-loss line and appear as almost pure susceptance. They can be open or short circuited, but they are generally short circuited because of ease of adjustment and lower radiation losses. The susceptance of a shorted stub equals −cot θ, θ being its wavelength in degrees. The stub susceptance appears on the charts on the $Q = \infty$ (no loss) circle which lies on the b axis, starting at $b = -\infty$ for a 0° stub. The susceptance of an open stub equals tan θ and appears on the b axis, starting at $b = 0$ (on the 90° curve for a 0° stub, thus −90° must be added for values of open stub lengths found above the g axis and +90° for values below the g axis.

In the example given above, the S_1 stub requires a susceptance of $b = -.62$ for a closed stub, this value of b corresponds on the chart to a length of 58° (or, as computed for greater accuracy, $\cot^{-1} - (-.62) = 58.2°$. The S_2 stub requires a susceptance of $b = -.94$, corresponding on the chart to a length of 47° (computed as 46.8°). If open stubs are used, the S_1 stub requires a length of (58° + 90°) or 148° and the S_2 stub, (47° + 90°) or 137°.

Example 2. Given an admittance $Y_L = 1.95 + j0$ with stub S_1 located at the load ($\phi_1 = 0$) and stub S_2 at a distance of 40°, to find the required stub lengths. On chart 1 the center, c, of the $\phi_2 = 40°$ circle is found by drawing a horizontal line to the double stub locus curve from the (180° − 40°) or 140° point on the b axis as shown in *Fig. 7*. The circular locus of ϕ_2 is drawn in,

[*Continued on page 36*]

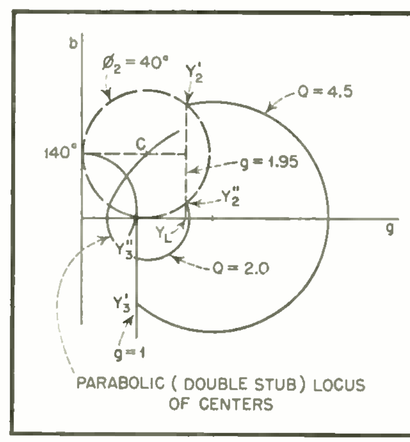


Fig. 7. Graphic solution of double stub matching problem 2, in which two possible solutions are shown.

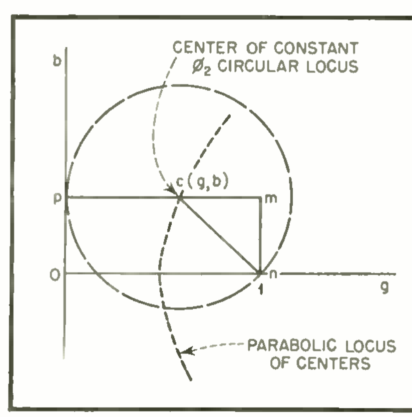


Fig. 8. Diagram illustrating the derivation of the equation of the parabolic curve of admittance circle centers of Figs. 1 & 2.

ADDITIONAL NOTES ON THE

FRANK C. JONES

VARIOUS QUESTIONS have come to light in connection with the higher fidelity parallel-tube amplifier described in the October issue of RADIO. Additional tests and construction of feedback amplifiers also have brought out the point that the more simple a circuit, the better it will function when built with various types of radio parts. In the original article, two circuits were shown with negative feedback applied over two stages. Experience with a number of different amplifiers has shown that the general type of circuit shown in *Fig. 1* is the most reliable and easiest to make operate correctly.

In any two-stage feedback system, phase shifts toward the high and low frequency ends, or beyond the useful ends of the range, will often produce regeneration or at least greatly reduce the desired degeneration. This results in an undesirable peak at the very low and often at the high frequency ends of the audio band. In several cases a peak occurred at some frequency below 30 cps with enough amplitude so a slow "motor-boating" or "breathing" effect took place especially at high signal levels. This type of trouble is difficult to stop when the feedback system is from the voice coil to the 6SJ7 cathode. The feedback system direct from 6L6 plate to 6SJ7 cathode permits the low frequency loss in the output transformer and coupling capacitor C_1, *Fig. 1*, to offset any low frequency peak within the feedback loop.

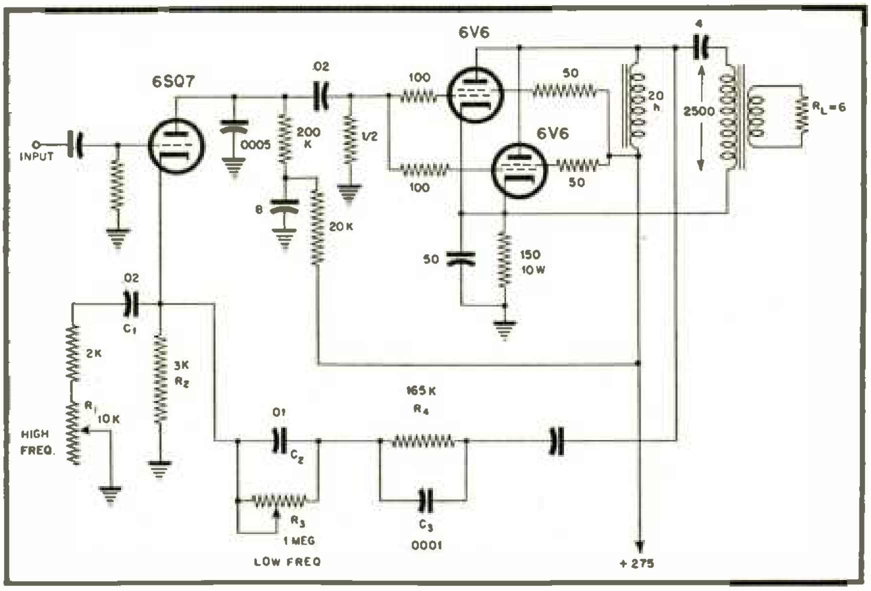


Fig. 3. Experimental amplifier with high—and low-frequency tone controls. Bass and treble response are independently variable.

The high frequency peak can be eliminated usually by shunting the feedback resistor R_2 by a small capacitor C_2. Normally the low frequency peak is much more difficult to eliminate. Values of C_1, the output coupling capacitor, of from ½ µfd up to 8µfd have been used with various amplifiers and makes of output transformers. A rule-of-thumb is to use a fairly large capacitor at C_1 for small or low priced output transformers and a small capacitor of ½ or 1 µfd for the highest grades of output transformers. This capacitor should preferably be an oil-filled, 600 volt paper capacitor. Push-pull output transformers are more available than the single type and usually half of the primary may be used with the other side left open. Good sized replacement output transformers may require the use of the full push-pull primary in order to secure enough primary inductance at low audio frequencies. All of the types of "high-fidelity" output transformers of the 15 to 45 watt sizes tested here functioned best with half of the primary left unused. A single 6L6 (with *200-ohm* cathode resistor), works well into a 2500 to 3000 ohms load. Two in parallel (with 100 ohm cathode resistor) function well with about half of that or 1500 ohms. Lower load impedances permit greater output but at increased harmonic distortion.

Parallel 807 tubes may be used to give greater output power since their plates and screens may be operated at 275 or even 300 volts. The same circuit constants are suitable for 807 tubes except for plate supply.

The amplifier shown in *Fig. 1* was built with a heavy duty power supply in order to take care of either an AM or a separate FM superheterodyne tuner..

The output of the amplifier shown

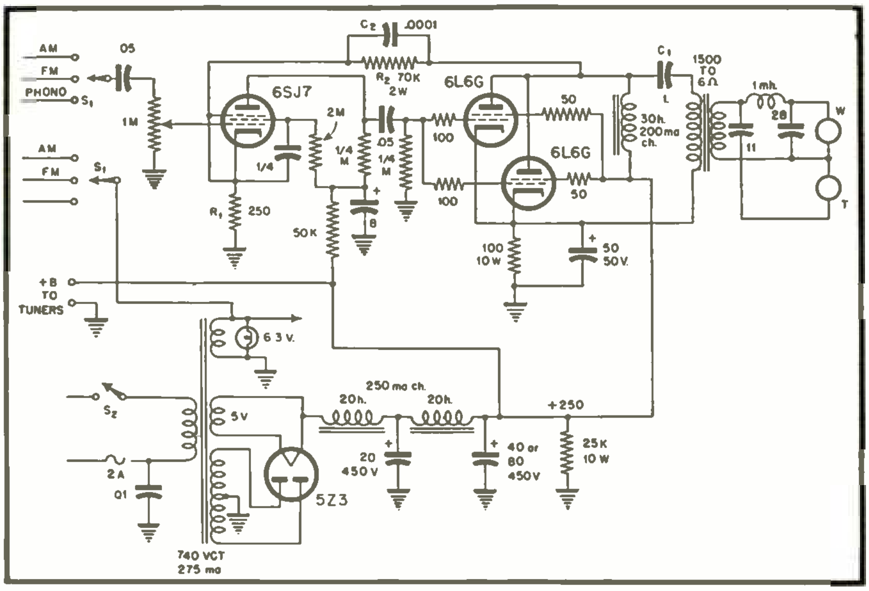


Fig. 1. Typical feedback circuit with stable characteristics. Feedback is applied over two stages.

PARALLEL TUBE AMPLIFIER

Parallel-tube amplifiers are capable of affording high fidelity and flexible bass-treble tone control

here was set to work into a pair of "6 ohm" loudspeaker, one as a 12" woofer and the other as a 6" tweeter. The cross-over filter consisted of standard paper 110 volt power factor correcting capacitors used in fluorescent lighting installations. The 1 mh filter coil consisted of a 100 foot roll of No. 14 enameled antenna wire, taped firmly, and then clamped to the cabinet with a wooden block. The two loudspeakers were phased correctly by first connecting them in series across a 1½ volt dry cell. Correct phase or series connection is when both diaphragms move in the same direction at the instant the battery connection is made or broken. The two speakers can then be connected to the cross-over filter and output transformer with this same series connection for correct phasing on audio frequencies.

The amplifier illustrated here had an overall frequency characteristic, measured across a 6 ohm terminating resistor on the output transformer, as shown in *Fig. 2*. The resultant gain from 6SJ7 grid to 6L6 plate was about 225 times, and the feedback about 12½ db. The harmonic distortion was measured with 500 cycles fundamental input and produced values of ¾% second and 2% third harmonic at 8.2 watts into a 6 ohm resistor load. At 10.7 watts into the resistor, the second measured 2% and the third 7½%. At 13½ watts output the second was 3½% and the third was 20%. This was considered more than ample for the 12 inch woofer speaker which was rated at 7 watts maximum power for its voice coil.

The same effects that cause peaks in a 2 stage feedback amplifier can be put to use where a tone control system is desired. Feedback tone control normally functions by decreasing the negative feedback at the high and low frequencies, thus increasing the gain at these two extremes. An experimental amplifier was set up with the circuit of *Fig. 3*, with high- and low-frequency tone controls. Either one could be adjusted to raise the response at one end without any appreciable effect on the middle of the range, a very desirable characteristic. The solid curve of *Fig. 4* is with the low-frequency tone control R_7 at minimum resistance and the high frequency control, R_7 at maximum resistance.

The dotted curve was obtained with R_7 set at maximum resistance of 1 megohm. The dot-dash curve of *Fig. 4* was obtained with R_7 at its minimum setting. Intermediate values were obtainable, the curves only indicating the maximum and minimum tone control effects. The tone control constants were suitable for a high mu 6SQ7 triode. However, it was found that the latter seemed to overload before the power stage, resulting in cancellation of all second harmonics at a certain input level, but quite high (and objectionable) third harmonics. A 6SJ7 pentode with a lower value of R_2 (perhaps 1500 ohms) in its cathode would have been a better driver stage.

Following up "Parallel Tube High Fidelity Amplifiers" in the October issue of Radio, Mr. Jose Velasco, of Santa Ana, El Salvador, C. A., presents some of his own experimental conclusions:

It is time that someone analyzed thoroughly, with laboratory instruments, the practical performance of power amplifiers using tubes in push-pull circuits—especially Class AB_1, AB_2, and B. It would appear that the latter has no place in high-fidelity applications.

Measurements bear out Jones' findings, namely, that parallel power amplifier tubes in a Class A circuit are vastly superior to a corresponding push-pull circuit, from the standpoint of distortion.

It cannot be too strongly emphasized that phase inverter stages introduce highly objectionable harmonic and cross-modulation products. Such inverter stages fail to remain balanced throughout the desired frequency range. All types of inverters have been tested and balanced by VTVM measurements at 50, 400, or 1000 cps with the aid of a variable resistor in one side of the circuit.

The inverter invariably exhibits serious unbalances at frequencies other than the test frequency. In some cases this unbalance reached 15%. This alone should prohibit its use in high-fidelity equipment.

[*Continued on page 37*]

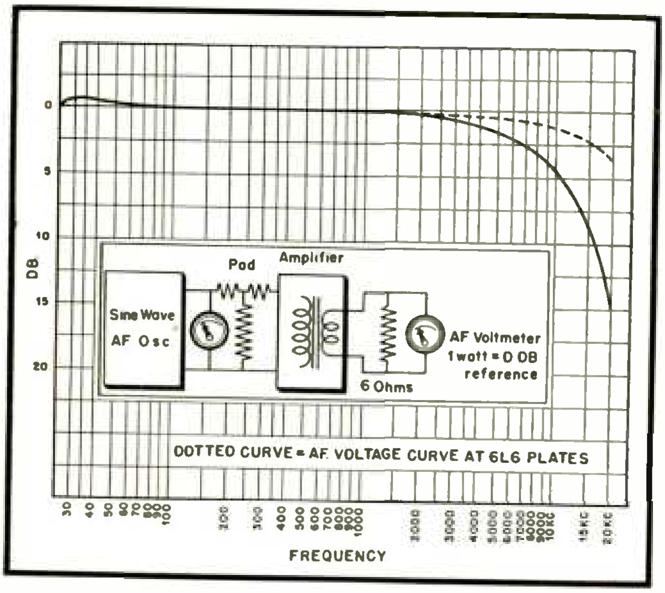


Fig. 2. Frequency characteristic of amplifier illustrated in Fig. 1. Gain is 225, with 12½ db feedback.

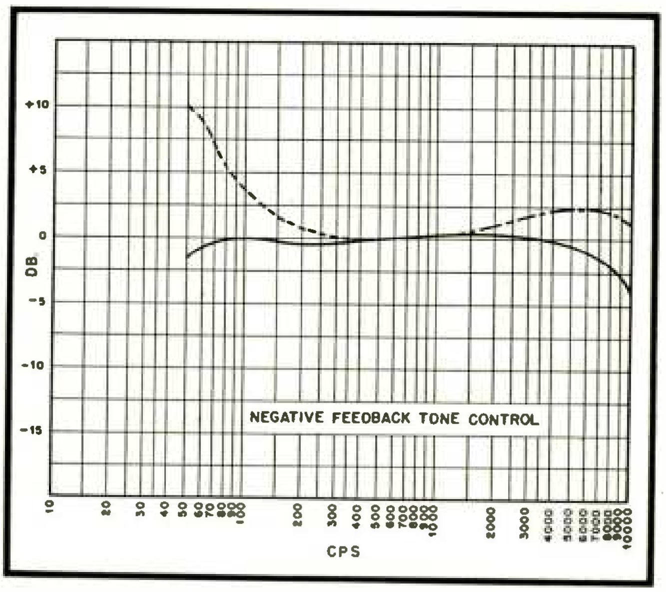
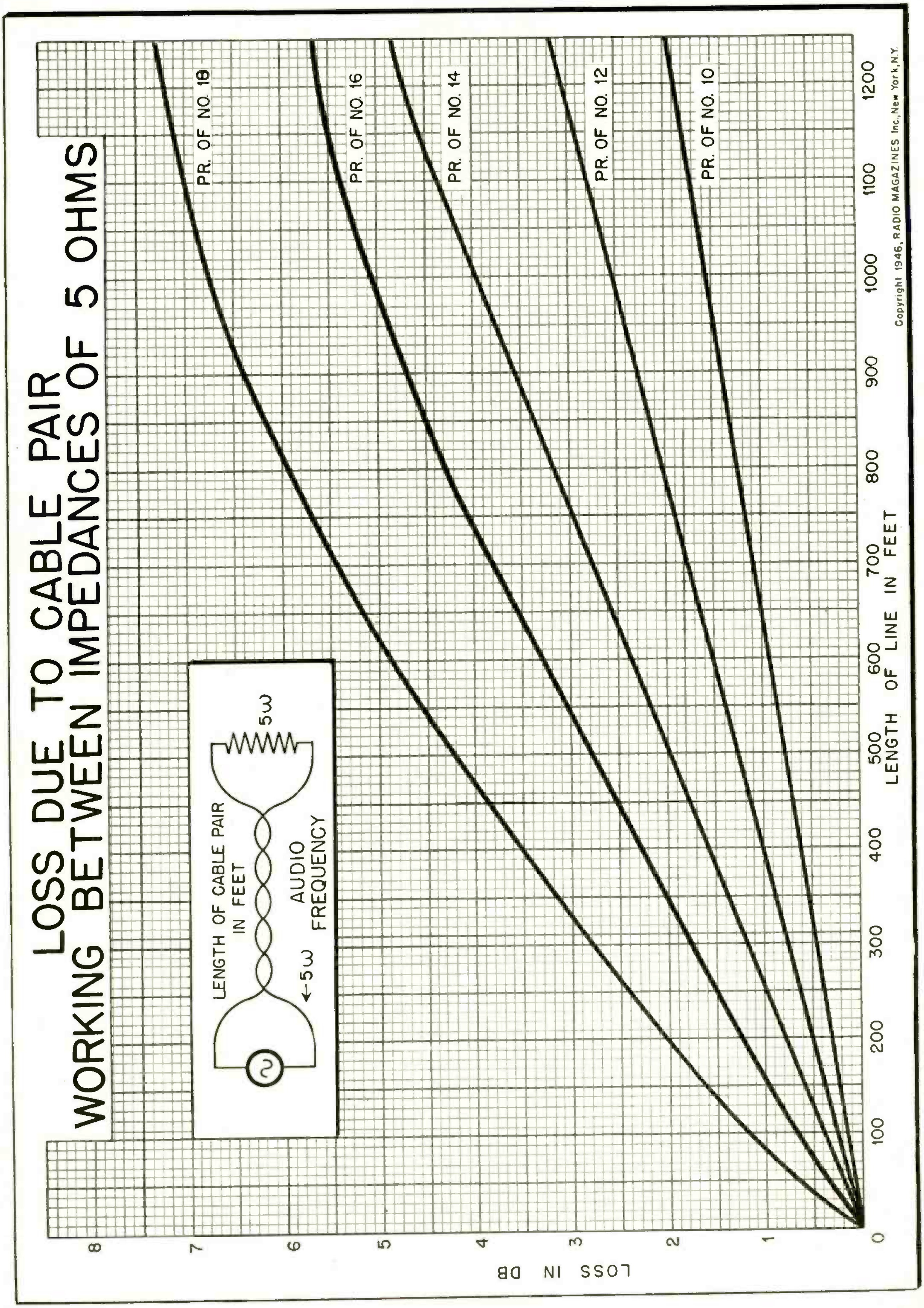


Fig. 4. Negative feedback tone control characteristics. Treble variation is 2½ db, bass variation 10 db.


LOSS DUE TO CABLE PAIR
WORKING BETWEEN IMPEDANCES OF 5 OHMS
PR. OF NO. 18
PR. OF NO. 16
PR. OF NO. 14
PR. OF NO. 12
PR. OF NO. 10
LENGTH OF CABLE PAIR IN FEET
AUDIO FREQUENCY
5ω
←5ω
LOSS IN DB
0
1
2
3
4
5
6
7
8
LENGTH OF LINE IN FEET
100
200
300
400
500
600
700
800
900
1000
1100
1200
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RADIO DESIGN WORKSHEET

NO. 49 – PERVEANCE

Conventional vacuum tubes are designed to operate with space-charge-limited current; there is in reserve at all times a potential emissivity greater than the power requirements of the anode circuit. The resulting space-charge concept was first formulated by C. D. Child and further studied by Irving Langmuir. For electrode geometry represented by a parallel-plane diode of infinite extent (no boundary conditions), Child's law states:

$i_p = 2.336 \times 10^{-6} e_p^{1.5}/d^2$ amp/cm² of surface, where d = cathode-anode spacing in cm.

Accordingly, space-charge-limited current is a non-linear function of voltage, being proportional to the 1.5 power of the latter. The i_p-e_p relation for the cylindrical-electrode diode (see *Fig. 1*) is formulated:

$$i_p = (14.68\ l \times 10^{-6} e_p^{1.5})/R_a\beta^2$$

where l = length of emitting cathode, R_a = the radius of the anode, and β is a function of cathode and anode radius, as given by the series:

$\beta = L - 2/5\ L^2 + 11/120\ L^3 - 47/3300\ L^4 + \ldots$ where $L = \log_\epsilon R_a/R_k$.

The variation of β^2 with R_a/R_k is graphed in *Fig. 2*. For ratios of radii in excess of 8/1, β is essentially unity.

The i_p-e_p relation for the cylindrical diode is thus a non-linear function, as

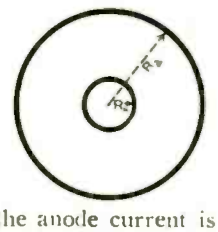


Figure 1

the anode current is again proportional to the 1.5 power of the anode voltage. For a given electrode geometry, $i_p = ke_p^{1.5}$. Rewriting the equation shows k as a form of conductance:

$$k = i_p/e_p^{1.5}$$

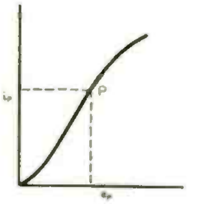
To point up the conductance property of the constant k, the notation may be changed to G, and termed a perveance:

$$G = i_p/e_p^{1.5}$$

Since the perveance is constant for a fixed electrode geometry,† and is independent of electrode voltages and currents, it constitutes a useful design factor and may be regarded as a figure of merit. The perveance of a given tube structure is increased if the anode voltage required to maintain a given space current is decreased. A perveance of 1×10^{-6} corresponds to 1 μa at 1v, 1 ma at 100v. or 1 amp. at 1,000v. The definiton of "perveance" appears to have been first stated by Kusunose.*

Under conditions of incomplete space charge, the perveance will no longer be a constant: at point P in *Fig. 3*, the perveance begins to vary. When more than one potential is involved, the perveance will remain constant under conditions of complete space charge, if the ratios between the potentials are maintained constant. The perveance will be

Figure 3



seen to be independent of the physical size of the electrodes, being a function of their shape and spacial relationships. The derivation of the perveance for any but the simplest electrode geometries is exceedingly difficult, and resort is usually made to empirical measurement.

It has been observed that the independence of perveance with respect to physical size holds for all designs in which the 1.5 power law is obeyed. Practical vacuum tubes frequently show considerable departure from the 1.5 power law. Fringing, emission velocity, and fields from electrode supports are all responsible. In the case of filament-type tubes, a further departure is traceable to the potential drop along the filament.

The variation of perveance G with R_a/R_k for R_a = 1 cm. is shown by the dotted curve in *Fig. 2*. It is not necessary, of course, that R_k be a physical cathode: as in the case of space-charge-grid tubes, a virtual cathode may be established by means of a positive grid, and this virtual cathode accordingly increases the perveance of the tube.

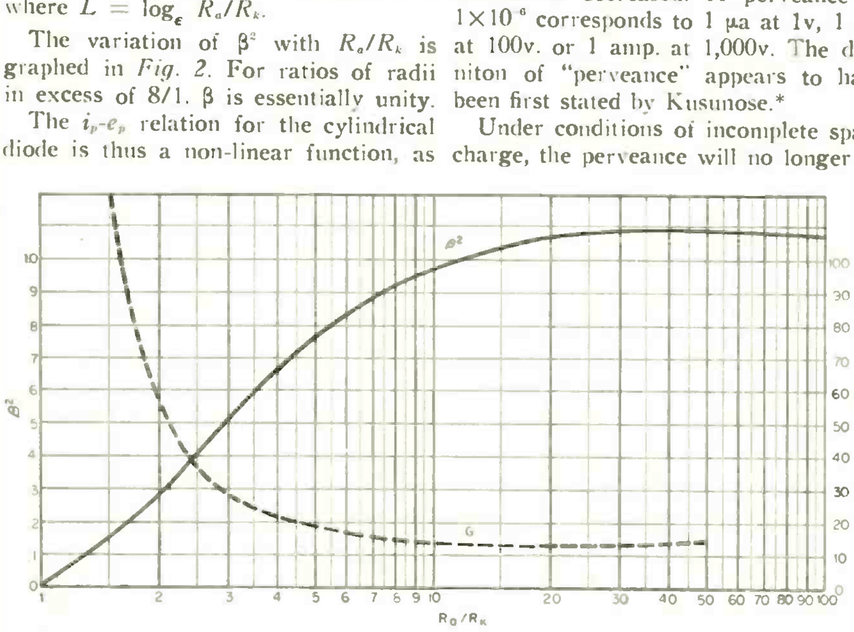


Figure 2

†Design of Electron Guns, A. L. Samuel, *Proc. I.R.E.*, Vol. 33, No. 4, p. 237. April, 1945.

*Design of Triodes, Yuziro Kusunose, *Proc. I.R.E.*, Vol. 17, No. 10, p. 1706. Oct., 1929.

New Products

LIMITS BRIDGE

Accurate testing of great quantities of resistors is claimed for the new low range limits bridge, Model 81-A, announced by Associated Research, Inc., 231 S. Green

St., Chicago 7, Illinois. Resistance from a fraction of an ohm to 20,000 ohms may be checked to tolerances of ± 1 per cent to ± 20 per cent. Plus and minus tolerances are separately adjustable.

The bridge is completely self-contained and uses three No. 6 dry cells. This internal 4½ volt battery suffices for intermittent duty for a wide range of resistances. Provision is made with binding posts for an external battery connection where higher sensitivity is required.

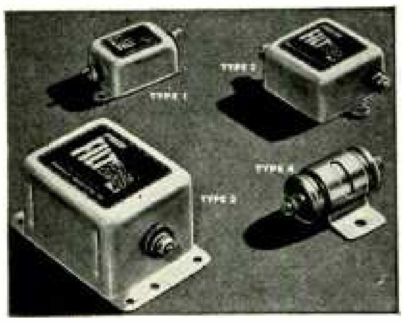
Full details may be had by addressing the manufacturer. Ask for Model 81-A, Bulletin 300.

EQUALIZER

A new mixer control claimed to reduce contact noise to a low value is announced by Cinema Engineering Co., 1510 West Verdugo Ave., Burbank, Calif. It uses a wedge-shaped roller riding on a plastic arm, contacting the resistance wire and a metallic track.

INTERFERENCE FILTERS

Filterol radio interference filters recently announced by the Sprague Products Company, North Adams, Mass., are small, completely self-contained units and are applicable to any electrical device within their ratings. Designed for installation in series with the power line or interfering device, Filterols should be mounted on the



frame of the device or in a grounded junction box as close to it as possible. Their basic circuit is a three terminal network of which the can is one terminal. Four available types include 115 volts a-c or d-c ratings from 1 to 35 amperes, and one unit for 220 volts a-c or d-c is rated at 20 amperes.

NEW HIGH FREQUENCY PROBE

A practical answer to the problem of measuring voltages in very high frequency circuits is offered, for the first time, in the Model 29 high frequency probe, recently developed by Alfred W. Barber Laboratories.

Designed to meet the need for a radio frequency probe with an extremely low input capacity, the Barber innovation, with an input capacity of ½ to 1 micro-microfarad, extends the range of measurements ten times, from 50 to 500 megacycles.

No multiplier is required to measure voltages up to 1000 volts. Frequency range 0.5 to 500 megacycles.

Descriptive Bulletins available from Alfred W. Barber Laboratories, 34-14 Francis Lewis Blvd., Flushing, N. Y.

NEW CODE BEACON

Designed and built by Andrew Co., Chicago 19, this 300 code beacon is 32¼ inches high and is required by the CAA for lighting radio towers of 150 feet or higher in order to minimize aviation hazards.

Two 500-watt prefocus lamps give an

32¼ inch high Code Beacon

intense light which passes through red pyrex glass filters and is then radiated in a circular horizontal beam by cylindrical fresnel lenses. Hardware is corrosion resistant bronze and metal parts are made of light-weight cast aluminum.

This code beacon is one of the tower lighting equipment items made by Andrew Co., 363 E. 75th St., Chicago 19. Andrew Co. is pioneer manufacturer of a complete line of antenna equipment.

BLIND RIVET "GUN"

An inexpensive, light-duty blind rivet "gun", the Jr. riveter, has been added to

the Cherry line of blind riveting tools.

It is a one-hand, plierlike tool that installs the rivet with a simple "pull", and is made especially for the many small fastening jobs arising constantly in any shop. It installs a new 3/32" diameter blind rivet, provided in three grip lengths. The rivet is the tight-clinching, pull-through hollow type having generous shank expansion.

For further information, write the Cherry Rivet Company, 231 Winston Street, Los Angeles 13, California.

CONTACT RECTIFIER

Latest addition to the Bradley line of copper oxide rectifiers, Model CO12D4F, is conservatively rated at either 6 volts a.c., 4½ volts d.c. or 35 ma, d.c., and is designed for control applications. The unit is completely sealed, with a special plastic compound, against moisture and corrosive vapors. Pre-soldered lead wires prevent damage from overheating during assembly. Manufactured by Bradley Laboratories, Inc., 82 Meadow St., New Haven 10, Conn.

NEW RACON PAGING SYSTEM

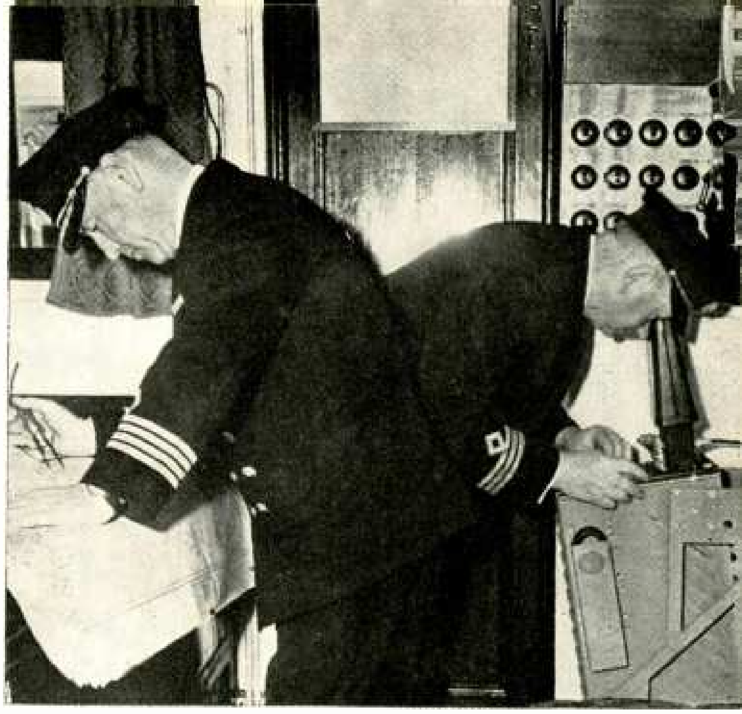
A new type of re-entrant paging horn having over 9½" of air column encompassed in a speaker only 4" in diameter and 2" in depth is announced by Racon Electric Co., 52 E. 19th St., New York City.

Where only voice reproduction is required in monitoring, intercom or paging installations, an Alnico-5 dwarf p-m driving unit is used. However, for installations distribution, the same speaker housing may which also require music as well as voice be used with a cone loudspeaker driver.

The speaker's small size and exceptional range makes it suitable for multiple-speaker installations or in cases where only small space may be provided for the sound distributors.

4 inch diameter paging horn

This Month



First Officer C. W. Johnsson operates RCA loran receiver as Captain John Nordlander charts the ship's position in the chart room aboard the Swedish-American liner Drottningholm, making a loran-guided Atlantic crossing.

OHMITE MEETING

After completion of the Radio Parts Show, Ohmite held a sales meeting at the Stevens Hotel, in Chicago. Attending the meeting were Ohmite sales representatives from over the United States, officers of the company, and members of the sales and engineering departments.

The meeting was opened with a talk by Mr. Siegel, president, who gave a shore history of the company and its operations. Mr. Lair, vice-president, sales, discussed sales policies and distribution problems. He also described briefly the many new items under development in the laboratory, some of them soon to be announced.

Mr. Howe, vice-president, discussed present and new products from the engineering and production standpoint. Mr. Buehling, advertising counselor, gave a brief outline of the future advertising program.

After the meeting the group attended a luncheon and in the afternoon were taken on an inspection trip to the Ohmite plant where the latest production methods were seen and the greatly enlarged laboratory was visited.

DESIGN NEW FM EQUIPMENT

More than 150 working radio men — stations owners, chief engineers and rank-and-file operators assisted in designing the new line of Westinghouse f-m transmitters, according to C. J. Burnside, manager of the industrial electronics division. Their preferences and suggestions, running all the way from improved eye-appeal to more efficient operation, were recorded in a nationwide survey of 91 stations in 56 cities of 22 states and drawn upon in establishing the basic design.

First of the new f-w line — the one and three-kilowatt units — already are in production. These are to be followed by the 10-kilowatt model to be ready about the end of the year, and later by the top-of-the-line 50-kilowatt unit.

SOUND STUDIO

The availability of completely equipped service facilities for independent and major producers in the sound recording field was announced by Hazard E. Reeves, President, Reeves Sound Studios, Inc., to be opened at 304 E. 44th St., New York City.

Inside the modern facade are five floors devoted to the production of motion pictures, radio transcriptions, sound films and television shorts. The studios, with original headquarters at 1600 Broadway, will be under the direct supervision of Chester L. Stewart, executive vice president and general manager.

RAYTHEON-SUB. SIGNAL MERGER

The merger of Submarine Signal Co. into Raytheon Manufacturing Co. was consummated on May 31, pursuant to the agreement of merger, dated April 24, between the two companies, after adoption by the required vote of the stockholders of each company, it was announced by Laurence K. Marshall, president of Raytheon.

W-E EXPANDS

Western Electric Company has leased the Kenmore plant of the Curtiss-Wright Corporation in the town of Tonawanda and the city of Buffalo, New York, according to C. G. Stoll, president of Western Electric. Comprising 760,000 square feet of floor space, the plant will manufacture switchboard cable and other telephone products and is expected to employ 3,000 people when in full production.

RIDER NAMED CONSULTANT

Lt. Col. John F. Rider, prominent radio engineer and writer, has been retained by the RCA Victor Division of the Radio Corp. of America as a consultant on test equipment, it was announced by Meade Brunet, vice president in charge of engineering products activities. Col. Rider will work in cooperation with the test and measuring equipment section, according to Mr. Brunet's announcement.

PERSONAL MENTION

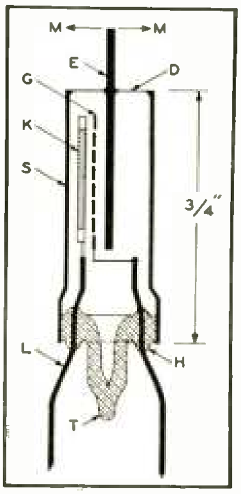
Glenn E. Webster, formerly in charge of speech equipment for Collins Radio Company, has been appointed chief engineer of the Turner Company, according to a recent announcement by Renald P. Evans, president. In his new position Webster will head the engineering department of the microphone and electronic division.

Mr. Webster comes with a background of twenty year's experience in radio engineering. He is a member of the Institute of Radio Engineers and the American Radio Relay League.

Ralph Bown

Dr. Ralph Bown, assistant director of research at Bell Telephone Laboratories since 1944, has been named director of research for that organization, effective today, Dr. Oliver E. Buckley, president of the laboratories, announced last night.

Dr. Bown, internationally recognized for pioneer research and development work in the broad field of communication engineering, has been associated with the Bell System since 1919. He succeeds Dr. M. J. Kelly who has been serving both as director of research and executive vice president of the laboratories and who will continue in the latter capacity.



The vibrotron, RCA's new tiny metal electron tube, converts motion directly to electron flow, weighs 1/15oz. and is 1 x 1¼". Adapted for use in pickups, microphones, and industrial control. (Crossection): Motion of movable electrode E affects electron flow of triode with cathode K, grid G, and movable anode E. Contained in metal envelope S with leads L through header H and exhaust tip T.

CASCODE

[*from page* 19]

Determining e_i

The output voltage for tube current i_1 is of course directly beneath 0 and its value is 300 volts to ground, or $e_L = 100$ volts. Point 0 is evidently the point of intersection of the R_L load line, (shown dotted), and the i_1 ordinate. To find the input signal voltage corresponding to point 0, it should be observed that the mu of a triode, while non-linear with respect to plate current, is approximately linear with respect to plate voltage for all negative values of grid voltage, and for moderately positive values.

By inspection, the tube mu for i_1 is 18 (compare with the published value of 20, current unspecified). Point P_1 determines a voltage $e'_{p1} = 2\Delta e_c$ (recall that $e'_{p1} = e_{c1} - e_{c2}$). By direct proportion $e_{p}'_{1} = 5.1$ volts; point P_2 is a mirror image of P_1 with respect to the median. Observe that $e_{c1} - e_{c2} = 5.1$ volts along the i_1 ordinate as required. Now P_1 is a point on the load line of V_1 and P_2 is a point on the load line of V_2, while 0 is a point on the load line of R_L. All three points correspond to a signal voltage $e_i = e_{c1} - E_{c1} = 1.95$ volts. As the output is taken across R_L, point 0 is of final concern, and it may be identified as 1.95.

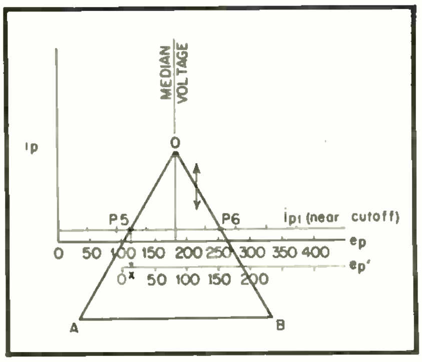


Fig. 4. When ultimate graphical precision is desired, or when working in the vicinity of cut-off, the method of successive approximations with sliding-rule technique may be used.

Operation Repeated

To find another point on the R_L load line with its corresponding signal voltage, i_1 may be chosen at 3 ma, resulting in a drop of 150 volts across R_L; accordingly, the output voltage is located at $0'$, on the 250-volt e_p abscissa as shown in *Fig. 2*. The median voltage is now 125 volts, and by inspection the mu is seen to be 19.2 (compare with the published value of 20); by proportion accordingly $e'_{p1} = 2.05$ volts. The signal voltage required is $e_i = 2.65$ volts, which is the value of input voltage associated with point $0'$.

Successive Approximations

Depending on the geometry of electrodes, departures from linearity of mu with respect to e_p are encountered in the vicinity of cutoff. When graphical analysis is of interest in limiting regions, accuracy is attainable by the method of successive approximations.

The application of the method is shown in *Fig. 4*. For greatest convenience, a draftsman's triangle is moved up or down the median voltage line to determine the condition of equilibrium. To maintain the triangle in its proper position, a line is ruled or scored to bisect the angle AOB. By placing the median line and this line of bisection over one another, the triangle is always properly positioned.

As in the proportional method, P_5 and P_6 bound an interval of grid voltage along i_{p1} which is equal to the interval of voltage OX along e'_{p1}. This is the condition of equilibrium sought.

H-F CABLES

[*from page* 21]

$70 = (138 \log_{10} D/d)/(1.5)$ therefore $D/d = 5.75$ (polyethylene insulation is at present standard for the industry).

A line whose power rating must be

ANOTHER NEW

Jensen *Coaxial*

The most significant postwar loud speaker development yet announced is the new Jensen family of Type H *Articulated* Coaxial Speakers. The latest member is Model HNP-51, an all ALNICO 5 design — in which low-frequency and high-frequency speakers are employed coaxially in an articulated assembly. The 15-inch l-f cone acts as an extension of the h-f speaker horn. The two loud speakers are electrically and acoustically coordinated into a system achieving brilliant and natural response through the entire useful frequency range (l-f performance depends upon the baffle or enclosure used). Frequency-dividing network has variable control in range above 4,000 cycles.

HNP-51 is recommended for FM receivers, high quality phonograph reproduction, television, review rooms, monitoring and home and public entertainment generally.

Coaxial Models HNP-50 and HNF-50 (for manufacturers) and HNP-51 (for general use), are now nearing quantity production. All Type J Jensen Coaxials (3 models) are now in production. Write for complete information.

TYPE H SPECIFICATIONS

MODEL HNP-51 (15-inch) with ALNICO 5 in both l-f and h-f units. Power rating, 25 watts maximum in speech and music systems. Input impedance, 500 ohms. List price approximately $125.

MODEL HNF-50 (15-inch) ALNICO 5 design h-f unit, field coil in l-f unit; otherwise same as HNP-51. List price approximately $115.

COMPLETE REPRODUCERS. Model HNP-51 Speaker is offered in 2 cabinet models to form complete reproducers. Model "CR" Reproducer employs beautiful Jensen Imperial Walnut cabinet. Model "RA" Reproducer employs attractively finished general utility cabinet.

Jensen SPEAKERS WITH ALNICO 5

JENSEN RADIO MANUFACTURING CO., 6615 S. Laramie Ave., Chicago 38, Ill.
In Canada: Copper Wire Products, Ltd., 137 Oxford Street, Guelph, Ontario

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high must have a very low attenuation and therefore a large standard-sized inner conductor should be used. Arbitrarily a #1 AWG solid copper inner conductor is selected. This makes d = .289″ and D, the outer diameter, becomes .289 × 5.75 or 1.66″. Using a typical value for the braid width and a 50 mil jacket, the dimensions of the cable (*Fig. 10*) become:

$$d = .289$$
$$D = 1.66$$
$$d_b = 1.71$$
$$D_s = 1.81$$

Knowing these values we can calculate the following parameters:

$$\alpha_c = (0.434 \cdot 100^{1/2})/(70 \cdot 0.289 \cdot 1.04 \cdot 10^4) = 20 \cdot 10^{-6}, \text{ using } R_1 \text{ of (28)}$$
$$\alpha_D = (12.6 \cdot 10^{-4} \cdot 100)/(1.04 \cdot 10^4) = 20 \cdot 10^{-6}, \text{ using (27)}$$
$$t_d = 119 \log_{10} 5.57 = 90$$
$$t_j = 314 \log_{10} 1.06 = 7.6$$
$$t_s = 60.3/1.81 = 33.3$$
$$W = 10{,}000 \text{ as defined}$$

If 185°F. is assumed to be the maximum safe temperature that the cable can withstand then α_t becomes:

$$2\alpha_t(7.6+33.3)+90(40\times10^{-6}+12\times10^{-6}) = (185-104)\times10^{-4}, \text{ using (46) ((A))}$$
$$2\alpha_t(41) = (8100-4600)\times10^{-6}$$
$$\alpha_t = 42.5\times10^{-6}$$

knowing what α_t must be, we can calculate α_b:

$$\alpha_b = \alpha_t-(\alpha_c+\alpha_D) = 42.5\times10^{-6}-(32\times10^{-6}) = 10.5\times10^{-6}$$

If α_b is to be 10.5×10^{-6}, then R_2 can be calculated from equation (28):

$$10^{-6}\times10.5 = (0.434 \cdot 100^{1/2} \cdot R_2)/(70 \cdot 1.66 \cdot 1.04 \cdot 10^4); \quad R_2 = 2.7$$

This can be accomplished by use of a #30 AWG copper braid that would give an R_b of approximately 2.5.

However, the selection of a safe maximum temperature of 185°F. may be too high in some applications and a figure of 175°F. is frequently used. If 175° is substituted in ((A)) then α_t becomes:

$$\alpha_t = 2500 \cdot 10^{-6}/82 = 30.2 \cdot 10^{-6}$$

which of course is too low if these dimensions are to be used for

$$\alpha_c+\alpha_d = 32.5\times10^{-6}$$

In order to reduce the attenuation a larger sized inner conductor must be used. Using a #00 AWG copper conductor whose diameter is .356″ we obtain the following characteristics of the cable (*Fig. 11*):

$$d = .365 \qquad c = .16\times10^{-6}$$
$$D = 2.10 \qquad D = .12\times10^{-6}$$
$$d_b = 2.20 \qquad b = .07\times10^{-6} \text{ using } R_2 = 2.5$$
$$D_s = 2.30$$
$$t_d = 90$$
$$t_j = 6.3$$
$$t_s = 26$$

$$W = 71/\{[90(30+12)+70(6.3+26)] \cdot 10^{-6}\} = 11.5 \text{ kw}$$

References [*Continued from Part 1*]

5 "The Theoretical Attenuation and Power Rating (Thermo) of the R. F. Cables," *Intra-Service-British Technical Note No. 146.*

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LOOP RECEIVERS

[*from page* 17]

applied over the entire *volume* occupied by the loop. Thus from the concept of voltage-plane gradient we may change to energy-volume gradient. While computing effective heights of the antenna we may arrive at certain values regardless of what concept we use. But we have already observed in recent years with the introduction of elongated loops (iron core type) that the shape of the polar diagram undergoes a noticeable change as the loop length is increased. For example, in the German iron-core loop where the length is three times greater than its diameter the polar diagram takes the shape of a "coffee bean" as shown in *Fig. 12.*

It is evident that a high Q loop of high h_{eff}, or high signal strength, produces sharper null determination than a low Q loop of low h_{eff} or a weak signal.

The observations in the screen room equipped with a transmission line established this rule beyond a doubt.

Thus, with this new technique of loop measurement, one can obtain all the data on loop sensitivity, antenna effect, directive properties and signal-to-noise ratio under conditions approximating actual reception.

DOUBLE STUBS

[*from page* 25]

intersecting the $g = 1.95$ vertical (there is no section ϕ_1, so $Y_1 = Y_L$) at two points, $Y_2' + Y_2''$, where $b_2' = +2.1$ and $b_2'' = +.25$ making possible two solutions. Since the Y_1 has no reactance, the S_1' stub must equal the value of b_2', or +2.1 for the first solution. This value falls on the $Q = 4.5$ circle which intersects the $g = 1$ vertical 40° farther along the arc at $Y_s' = -1.65$. Then the susceptance of the S_s' stub must equal the opposite of this value or +1.65 for the first solution.

For the second solution stub S_1'' must equal the value of b_2'', or +.25. This value falls on the $Q = 2$ circle which intersects $g = 1$ at Y_s'' where $b_s'' = -.7$, so the S_s'' stub for the second solution must equal +.7. The required stubs for the solution of this problem then are:

Required length of stubs for *example 2* is shown in Table 1.

For *example 2* it is seen that the first solution results in a higher Q or standing wave ratio for the matching section as well as longer stubs, so probably the second solution would be more desirable.

Theoretical Considerations

The proof that the locus of admittances for any angular spacing of stubs, ϕ, is a circle may be of interest. It is shown in many textbooks that the input admittance of a length ϕ (1) of a transmission line is $Y_1 = (Y_L + j \tan \phi)/(1 + jY_L \tan \phi)$ where Y_L is the load admittance. Substituting rectangular coordinates of conductance and susceptance, g and b, for Y_L we obtain

$$[g+j(b+\tan\phi)] / [(1-b\tan\phi) - jg\tan\phi] \quad (2)$$

rationalizing the above

$$[g(1+\tan^2\phi) + j(1-b\tan\phi)(b+\tan\phi) - jg^2\tan\phi] / [(1-b\tan\phi)^2+g^2\tan^2\phi] \quad (3)$$

Since this expression must equal $1 + jb_s$, the real parts can be set equal, then

$$1=g(1+\tan^2\phi)/[(1-b\tan\phi)^2+g^2\tan^2\phi] \quad (4)$$

clearing of fractions and transposing,

$$g^2 \tan^2 \phi - g(1 + \tan^2 \phi) + (b^2 \tan^2\phi - 2b \tan\phi + 1) = 0 \quad (5)$$

	Shorted		Open
S_1'	$\cot^{-1} - (+2.1) = 154.5°$		$\tan^{-1} + 2.1 = 64.5°$
S_2'	$\cot^{-1} - (+1.65) = 148.8°$		$\tan^{-1} + 1.65 = 58.8°$
	or		or
S_1''	$\cot^{-1} - (+.25) = 104°$		$\tan^{-1} + .25 = 14°$
S_2''	$\cot^{-1} - (+.7) = 125°$		$\tan^{-1} + .7 = 35°$

Table 1

Dividing through by $\tan^2\phi$ and adding $[(1+\tan^2\phi)/2\tan^2\phi]^2$ to both sides we obtain

$$[g-(1+\tan^2\phi)/2\tan^2\phi]^2+(b-1/\tan\phi)^2 = [(1+\tan^2\phi)/2\tan^2\phi]^2 \quad (6)$$

since $(1+\tan^2\phi)/\tan^2\phi = \csc^2\phi$, this may be written

$$(g-\csc^2\phi/2)^2+(b-\cot\phi)^2=(\csc^2\phi/2)^2 \quad (7)$$

which is the equation of a circle. The center of this circle is at $g = \csc^2\phi/2$, $b = \cot\phi$, which is the parametric equation of the parabolic locus of the centers of all such circles.

The equation of the parabolic locus for the centers of the circles can also be derived by geometric means as shown in *Fig. 8*. All such circles are tangent to the b axis and pass through the point $g = 1$, $b = 0$ at which point $Q = 1$ and its corresponding circle becomes a point. In the figure $cn = \sqrt{mn^2+cm^2}$, but the center c has the coordinates g and b, therefore $mn = b$, $cn = pc = g$, and $cm = 1-g$. Then $g = \sqrt{b^2+(1-g)^2}$ Squaring both sides and transposing, we obtain $b^2 = 2g-1$, which is the equation of the parabolic locus in rectangular coordinates.

The charts of *Figs. 1* and *2* have been used here for computing admittances and susceptances in the design of stubs to be connected in parallel with the line. The same charts are used for calculating impedance problems and can be used for computing the values of series reactive elements to be used in series with the line for standing wave correction in a manner analogous to the parallel reactive stubs described above.[3] However, such arrangements are less commonly used.

Some other interesting methods of computing double stubs for impedance matching have been presented elsewhere.[3, 4]

The circle diagram may be easily drawn on a drafting board with the aid of the following rules:

Note that the circles of constant R_r/R_o intersect the axis of abscissas at reciprocal points; i.e., the $R_r/R_o = 3$ circle intersects the axis at 3 and at 1/3.

Each of the semicircles has its center on the axis of ordinates, at a distance of contangent $2\beta x$ from the origin. Calibration of this semicircle is then βx degrees (or radians). Radius of the semicircle is fixed by the requirement that it pass through (1, 0).

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AMPLIFIERS

[*from page 27*]

Speaker Distortion

Numerous tests of power output into voice coils of various loudspeakers fail to show better than 50% of rated undistorted power output. While the voice coils easily handle rated output, sufficient non-linear distortion is observed that high-fidelity requirements can be met only by reduced input levels.

Experimental Amplifier

In an attempt to achieve a true high-fidelity power amplifier, the following equipment has been built and tested: It consists of a dual-channel amplifier and elaborate loudspeaker system, using a low-frequency unit for the low-frequency amplifier, a middle-frequency unit for the high-frequency amplifier, and two high-frequency units working out of a filter.

The amplifiers make use of electronic volume expanders and separate booster circuits for low and high frequencies; the low-frequency channel uses four type 6A5G's in a parallel push-pull circuit with self-bias which is adjustable. Each cathode circuit includes a jack for a milliameter to assist in balancing the circuit.

From the input to the ouput stage, negative feedback is used, and to excite the power stage a pair of good audio transformers are used, rather than a phase inverter. The cross-over frequency can be adjusted from 560 cps to 2400 cps and the attenuation of each side of the cross-over point is about 8 db per octave. At present, a cross-over of 880 cps is being used.

The high-frequency channel uses a pair of 6L6's in the power stage. A highly degenerative link is used, and high-quality audio transformers are also used to excite the 6L6s.

Frequency response of the amplifiers is from 20 cps to 15000 cps; the bass boost is 14 db at 50 cycles and the treble boost is approximately 10 db at either 4500 or 8800 cps. Or course, the response can be made flat within these limits, or the bass and treble may be attenuated as desired. A filter cutting off sharply past the cut-off frequency of 5800 cps is used in the high-frequency channel.

Projected Amplifier

It is now planned to construct a six-tube power stage parallel-amplifier using 6A5G's in Class A. At least 15 watts output are expected, since the plate load will be approximately 500 ohms, with a voice coil impedance of 8 ohms. This is an impedance ratio of 62.5 to 1, and will make for high efficiency of transformation, using a top grade output transformer.

Comparisons will be made with six 6L6's in Class A, using the same system. Since the plate load resistance of both the 6A5 and 6L6 tubes will be the same (2500 ohms) the benefits of high transformation efficiency will be realized.

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