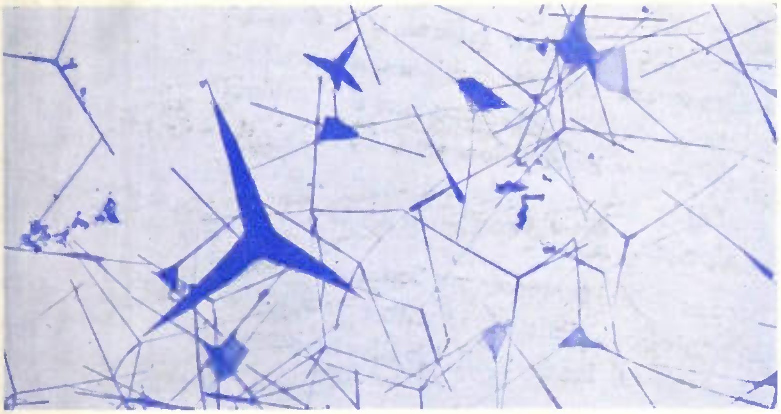
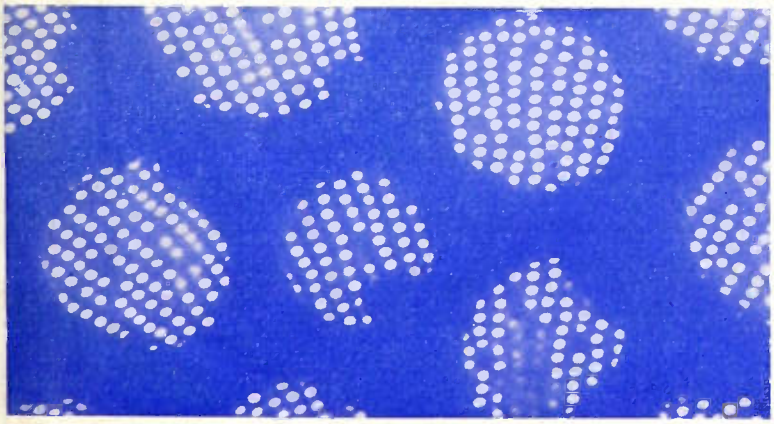
Proceedings of the I·R·E

AUGUST 1943

VOLUME 31 NUMBER 8





Radio Corporation of America Photographs

THROUGH THE RADIO-AND-ELECTRONIC EYE OF THE ELECTRON MICROSCOPE

Nature's minute and otherwise invisible world yields its secrets.
Above: Zinc-oxide smoke.
Below: Diatoms.

Institute of Radio Engineers

Proceedings of the I·R·E

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VOLUME 31 — *August, 1943* — NUMBER 8

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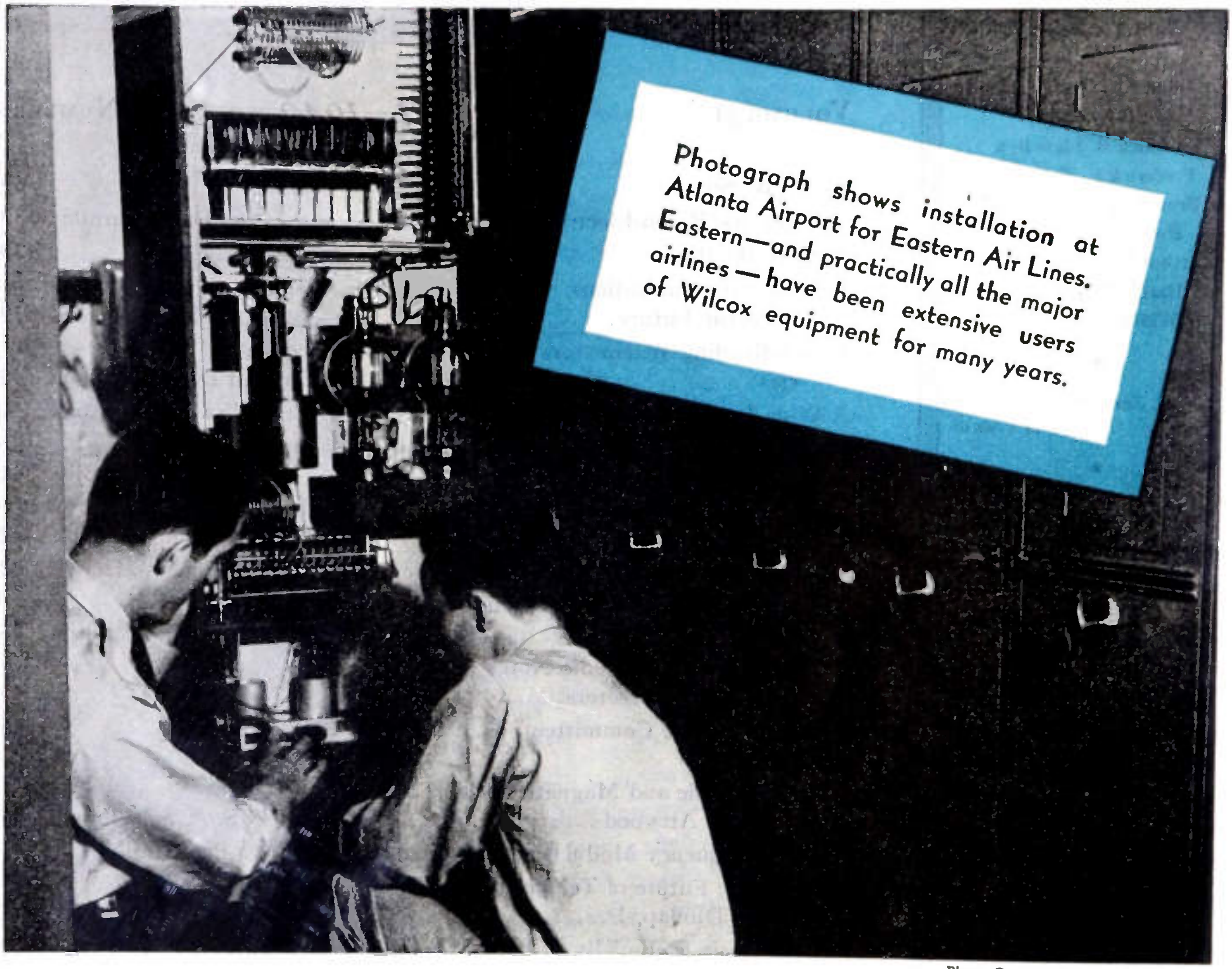
WILCOX EQUIPMENT
Proves Dependable for Eastern Air Lines



Photo, Courtesy Eastern Air Lines

Communication Receivers
Aircraft Radio
Transmitting Equipment
Airline Radio Equipment

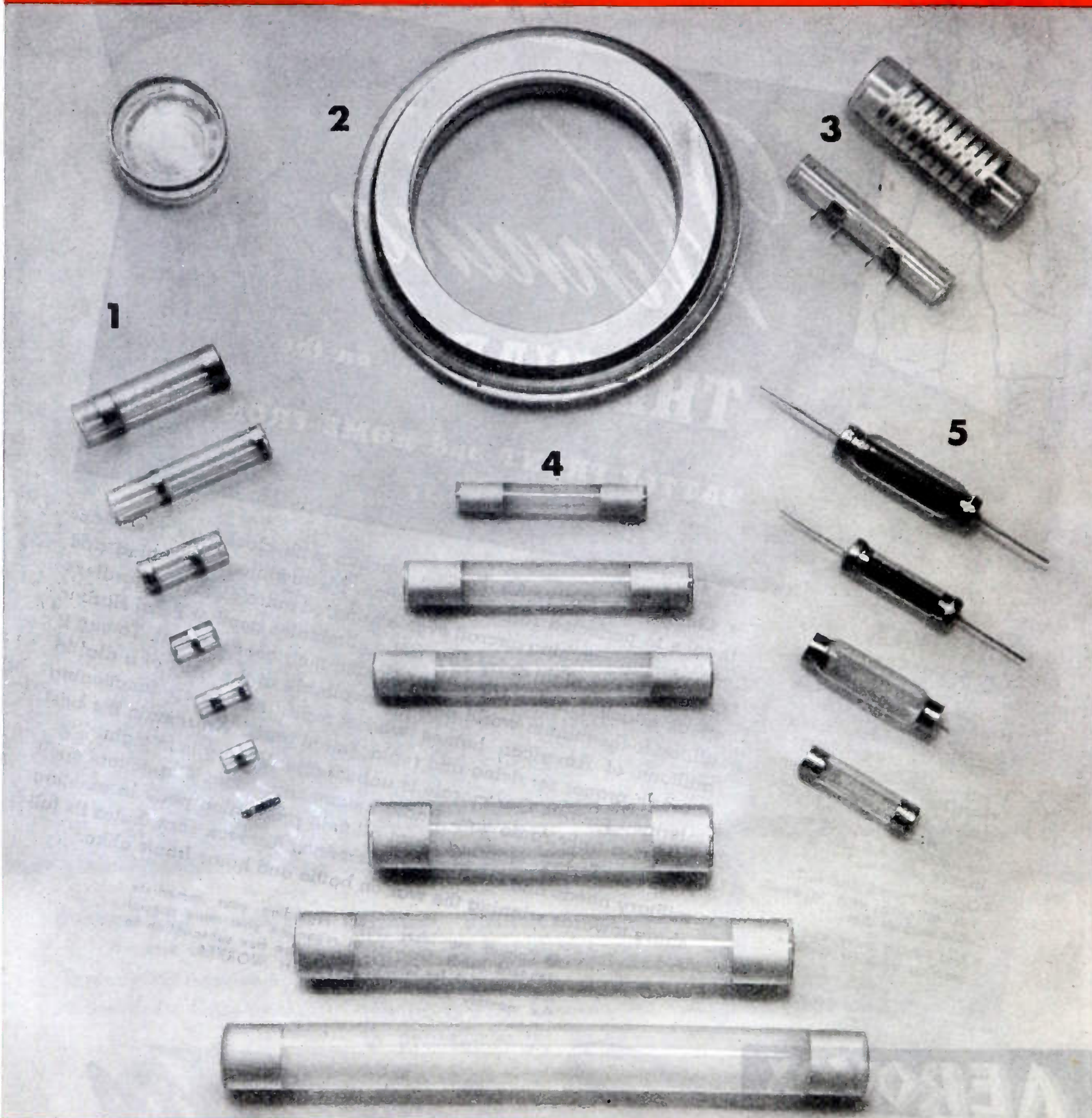
Today, the experience of years in manufacturing flight control radio equipment is turned to production for military needs. Tomorrow, this added experience with present developments will be reflected in greater radio advantages for a peacetime world.

WILCOX ELECTRIC COMPANY

Quality Manufacturing of Radio Equipment

14th & CHESTNUT **KANSAS CITY, MO.**

THERE MUST BE DEPENDABLE COMMUNICATIONS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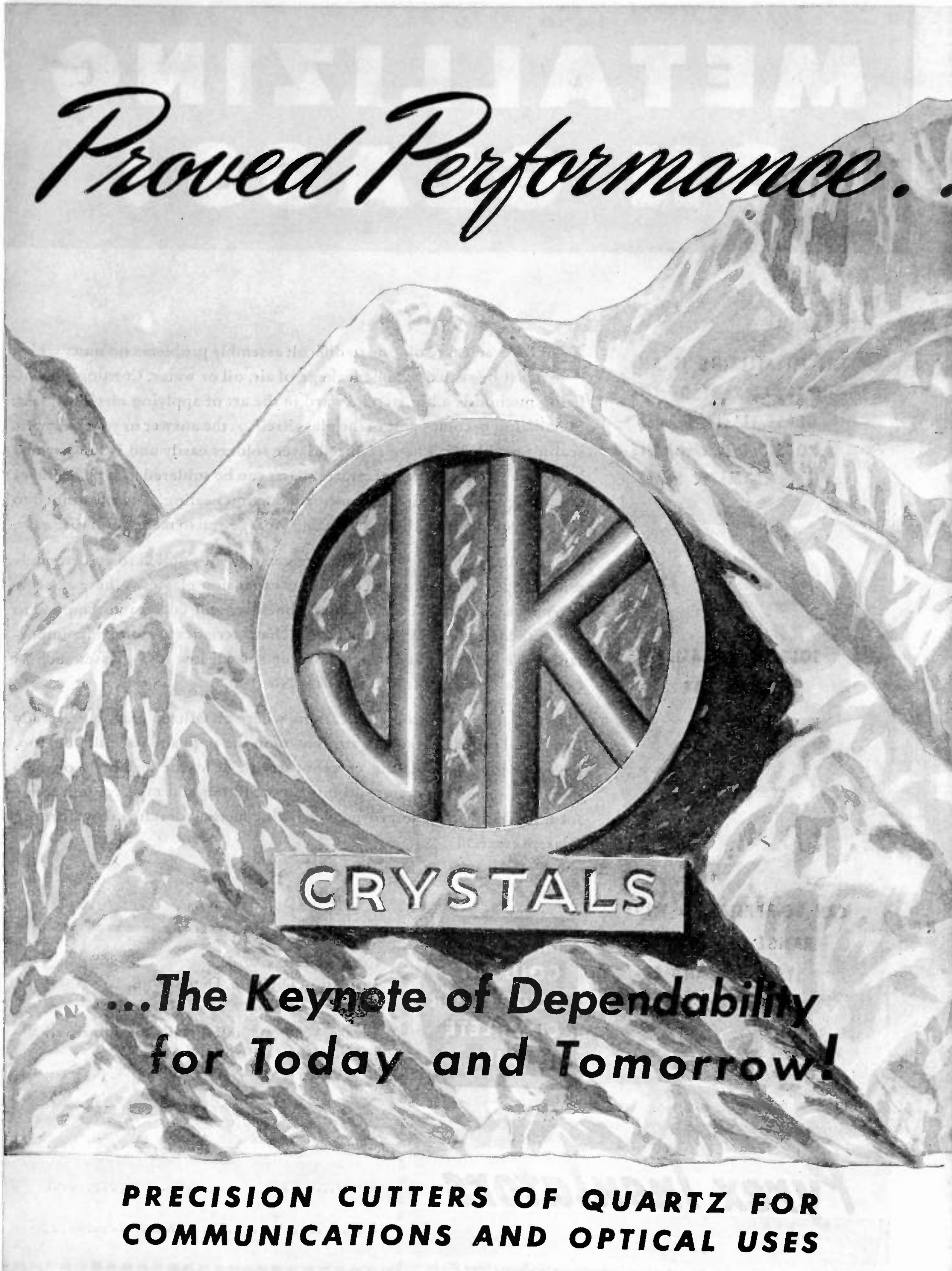

Proved Performance...
JK
CRYSTALS
...The Keynote of Dependability for Today and Tomorrow!
PRECISION CUTTERS OF QUARTZ FOR COMMUNICATIONS AND OPTICAL USES


SOUTHERN CROSS
REMEMBER THE "SOUTHERN CROSS"?
CAPTAIN Charles J. Kingsford-Smith on the morning of May 31, 1928, lifted the *Southern Cross* off the runway of the Oakland Airport and headed home for Australia. Eight days later his tri-motored Fokker monoplane landed at Brisbane after a three-legged trans-Pacific hop of 7362 miles.
"This trip," according to the *New International Year Book*, "was notable for the accuracy of the navigation and the fact that the radio operator was constantly in communication with either shore stations or ships throughout the flight."
The transmitter then aboard the *Southern Cross*, and now on display in the Smithsonian Institute, was designed by Ralph M. Heintz, co-founder of Heintz and Kaufman, Ltd. Its signal on 33.1 to 33.5 meters never faltered during the 83 hours and 19 minutes of flying.
The experience gained by Heintz and Kaufman, Ltd. while pioneering memorable events in radio history is reflected in the Gammatron tubes today serving the military, naval and air arms of the United Nations. The efficiency of modern Gammatrons at high frequencies, their long operating life and high stability, spring from 17 years of continuous and often brilliant research and development by Heintz and Kaufman engineers.
HEINTZ AND KAUFMAN, LTD.
SOUTH SAN FRANCISCO • CALIFORNIA, U. S. A.
GAMMATRON TUBES ... The grids and plates in Gammatron tubes are made of tantalum, a unique metal which eliminates the need for unstable getters, and protects Gammatrons from the release of gas even when heavily overloaded. Other Gammatron advantages: low driving power, easy neutralization, freedom from parasitics, and high efficiency at radio and very high frequencies.
HK-454
LEADERS IN ELECTRONICS YESTERDAY, TODAY AND TOMORROW

EVERY TAYLOR TUBE
IS Custom Built!
Individual Treatment Assures Uniform Dependability....
The performance of a high power transmitting tube depends largely on the purity of the metals used. Metals vary — each requires more or less or even different treatment if the utmost in uniform dependability is to be attained. That's why every Taylor Tube is Custom Built!
Individual heating, evacuating, bombarding and flashing must be scientifically perfect to insure a perfect tube. A final "OK" on a Taylor Tube is never given lightly. It is a token of our sense of responsibility. Where tubes get hard service and a chance to be overloaded, it's the Custom Built tube that can be depended upon to "deliver the goods".
810
Taylor HEAVY CUSTOM BUILT DUTY Tubes
TAYLOR TUBES, INC., 2312-18 WABANSIA AVE., CHICAGO, ILLINOIS

A miracle metal... Magnesium serves a vital wartime need in tracer ammunition, parachute flares, and illuminating signals. To reduce this difficult, extremely hazardous metal to powder, Magna Manufacturing Company developed special machinery and precision processes. Now, in three Magna plants... the largest facilities of their kind... we reduce Magnesium to a uniform dust in compliance with rigid U.S. Army and Navy standards.

BUT TOMORROW...

The facilities that Magna perfected open new possibilities for the utilization not only of magnesium but also *all* other types of disintegration-resisting metals and other materials such as ceramics, plastics and pigments.

Magna capacities are today entirely occupied with production for victory. However, Magna engineers are prepared to council forward looking industries on the potentialities of powdered metals and other materials in postwar product planning.

An Invitation...

Manufacturers, design engineers, and technicians are invited to investigate Magna facilities and to work experimentally with our staff to develop new and better products... through the practical disintegration of desirable materials heretofore unavailable in powdered form. Your inquiry is invited.

MAGNA

MANUFACTURING COMPANY, INC.

MANUFACTURERS OF MAGNAFLAKE METAL POWDERS

444 MADISON AVENUE, NEW YORK 22, N. Y.

OF LEADERSHIP

CONSTANT PIONEERING . . . the unceasing search beyond present horizons for a Better Way . . . this is the responsibility of Leadership.

As man reaches for the stars, through the science of electricity . . . the miracles of Radio and Television . . . the amazing new world of Electronics, we at American Lava Corporation have paced each new achievement with insulation engineered to the new requirements.

When war came upon us with its demands for tremendously expanded production . . . its requirements for higher performance in Communications . . . we were ready with the KNOW HOW gained from 40 years' experience in pioneering, developing and perfecting steatite ceramic insulation.

Our research and engineering staffs will gladly cooperate on today's blue print . . . tomorrow's production.

AMERICAN LAVA CORPORATION

CHATTANOOGA, TENNESSEE

ARMY E NAVY

AWARDED JULY 27, 1942

Where stability is an important requirement, AlSiMag Steatite ceramics are unsurpassed for lending rigidity and permanence of alignment to electronic circuits.

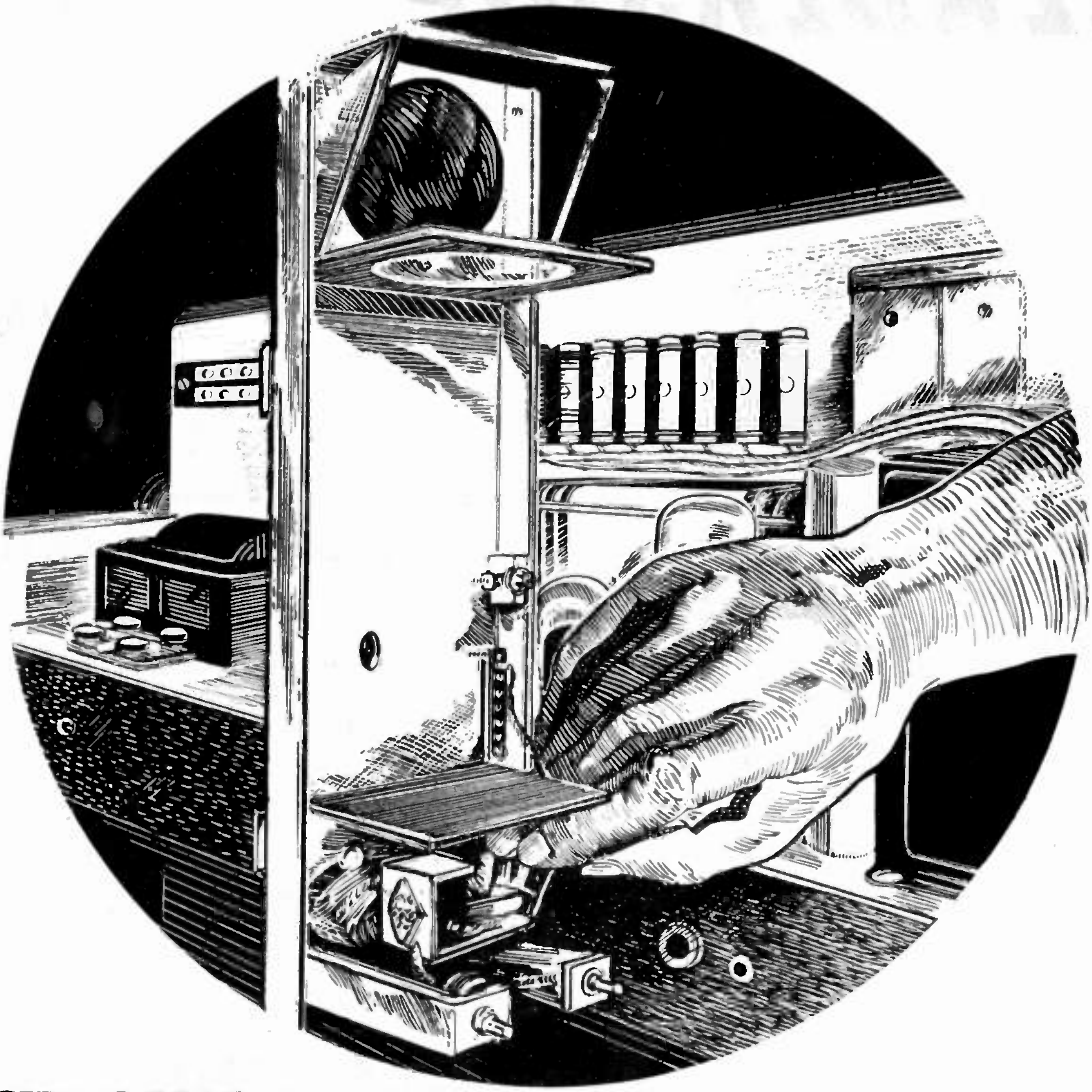
ALSIMAG

TRADE MARK REGISTERED U. S. PATENT OFFICE

STEATITE CERAMIC INSULATORS

CHARACTERISTICS TAILORED TO YOUR REQUIREMENTS

ON GUARD AGAINST SABOTAGE!



ELECTRONIC TUBES AID PLANT PROTECTION

America's war production plants are carefully guarded . . . night and day electronic tubes guard against intrusion. Units that transmit a virtually invisible light beam are powered by tubes that are constantly on the job . . . guarding against sabotage! . . . Raytheon's experience during its war time production of tubes for our armed forces will prove an invaluable factor when these new developments can be released for general domestic uses.

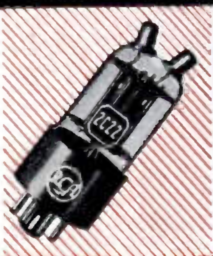
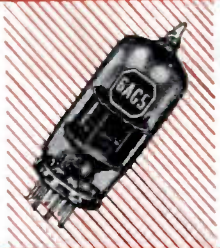
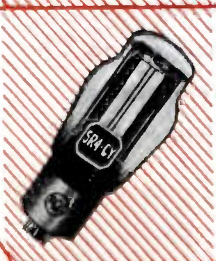
RAYTHEON

Raytheon Manufacturing Company

Waltham and Newton, Massachusetts

DEVOTED TO RESEARCH AND THE MANUFACTURE OF TUBES AND EQUIPMENT FOR THE NEW ERA OF ELECTRONICS

4 NEW RCA TUBES!

TYPE	MECHANICAL DATA	ELECTRICAL DATA	APPLICATION DATA
2C22 AMPLIFIER TRIODE 	Max. overall length......3¼" Bulb..................T-9 Caps (2): Skirted miniature Base: Intermediate shell octal 8-pin Mounts in any position	Heater volts..........6.3 AC or DC Heater amps..................0.3 Max. plate volts..............300 Plate dissipation..........3.3 watts Plate milliamps..................11 Mu......20 Rp......6600 ohms Transconductance...3000 micromhos	Designed for special high-frequency applications including transmitter service. Approx. resonant frequency of grid-cathode circuit is 335 megacycles.
6AG5 MINIATURE R-F AMPLIFIER PENTODE 	Max. overall length......2⅛" Bulb..................T-5½ Base: Miniature button 7-pin Mounts in any position	Heater volts..........6.3 AC or DC Heater amps..................0.3 Max. plate volts..............300 Plate dissipation............2 watts Transconductance...5000 micromhos	Sharp-cut-off type designed for compact, light-weight equipment as an r-f amplifier, and as a high-frequency intermediate amplifier.
6J6 MINIATURE TWIN TRIODE 	Max. overall length......2⅛" Bulb..................T-5½ Base: Miniature button 7-pin Mounts in any position	Heater volts..........6.3 AC or DC Heater amps..................0.45 Max. plate volts..............150 Max. plate dissipation.....1.5 watts* Max. plate milliamps...........15* Class C telegraph service: At 150 volts: Output watts....3.5 Driving watts..0.35 Cathode is common to both triodes. * Each triode	Designed for either push-pull or parallel operation up to 600 megacycles. With push-pull grid circuit and plates in parallel, the 6J6 makes a good mixer at frequencies up to 600 megacycles. Also useful as an oscillator.
5R4-GY FULL-WAVE HIGH-VACUUM RECTIFIER 	Max. overall length...5-5/16" Bulb..................ST-16 Base: Medium shell octal 5-pin, Micanol Mounts vertically, or mounts horizontally with pins 1 and 4 in vertical plane.	Filament volts..................5.0 Filament amps..................2.0 Max. peak inverse volts (No-load conditions)........2800 Peak plate milliamps per plate...650 Max. DC output amps.........0.25 at 2400 peak inverse volts.	

FREE

RCA

RCA ELECTRON TUBES

RCA Victor Division

RADIO CORPORATION OF AMERICA

Camden, N. J.

FOR VICTORY BUY UNITED STATES WAR BONDS AND STAMPS

RCA Commercial Eng. Section
Radio Corp. of America
447 So. Fifth St., Harrison, N. J.
Gentlemen:
Please send me your handy 16-page booklet (Form 1275B) which shows "Characteristics and Socket Connections of RCA Receiving and Allied Special Tubes" in easy-to-read tabular form.

Name..............................

Street..............................

City.............................. State..............................

EVERY MINUTE *Counts!*

PRODUCTION LINES MUST NOT SLOW DOWN!

Lafayette Radio is strategically located to give you *quick* deliveries on radio and electronic parts and equipment. Millions of items have been shipped from Chicago, the shipping hub of the nation, to industrials, training schools and all branches of the armed services. Lafayette's procurement and expediting service has helped to prevent work stoppages on many vital war production lines.

Many instances are on record wherein Lafayette has made immediate delivery on hard-to-find key items, eliminating costly delays in giant armament programs. This is because Lafayette handles the products of every nationally known manufacturer in the radio and electronic field. A single order to Lafayette, no matter how large or how small, will bring prompt delivery of *all* your requirements.

Free—130 page Catalog—"Radio and Electronic Parts and Equipment." Write 901 W. Jackson, Chicago, Ill. Dept. 8H3.

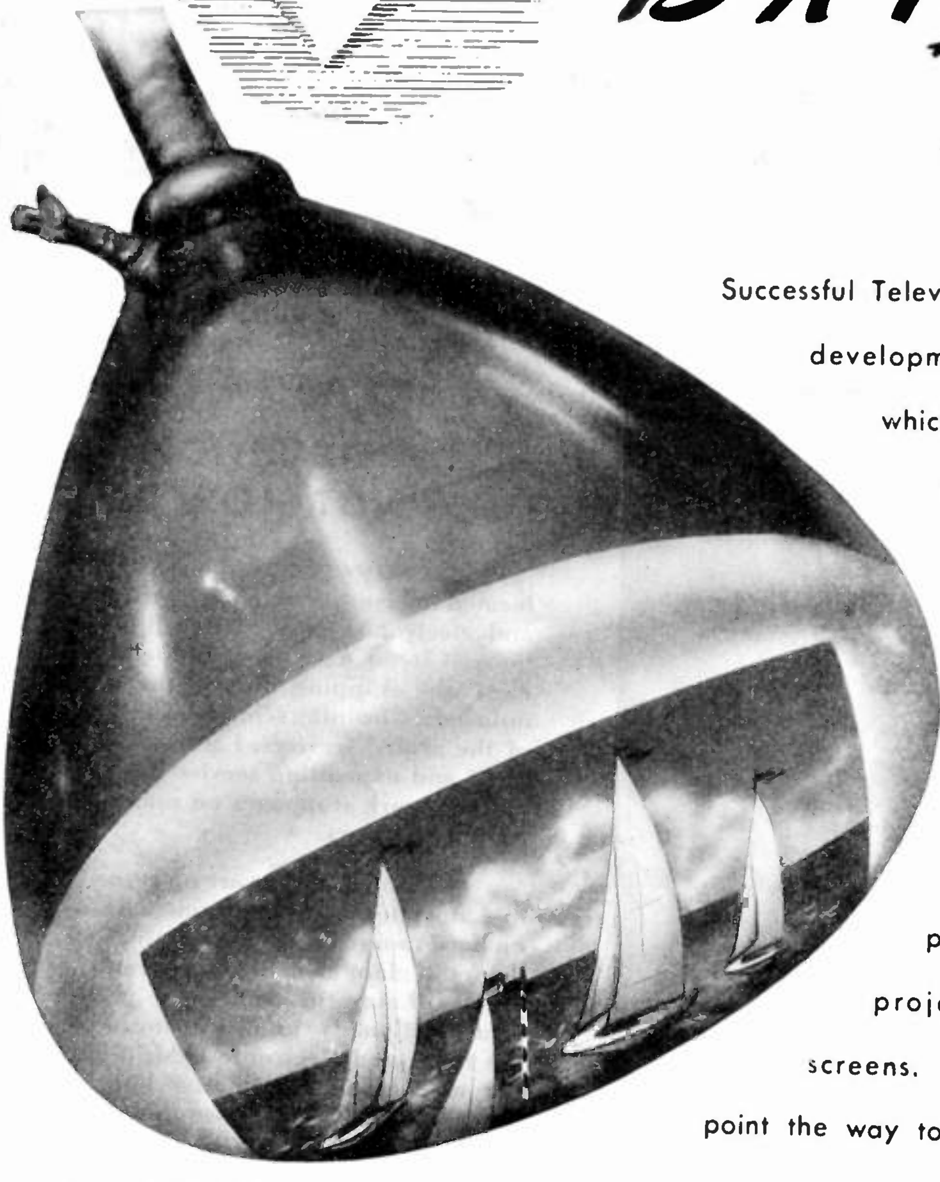
LAFAYETTE RADIO CORP.
CHICAGO

★ ★ ★ ★ ★ ★ ★

"Quick Delivery of Radio and Electronic Parts and Equipment"

THE HEART OF TELEVISION

IS A TUBE!



Successful Television came into view with the development of the cathode ray tube which gave it definition—translating millions of electrical impulses per second into animated reproductions of scenes occurring at other points. ★ RAULAND tubes played a major part in this epochal achievement by making possible practical television projection on 15 foot x 20 foot screens. Significant facts like these point the way to effective post-war planning.

Illustrated above is a 15-inch Cathode Ray Tube

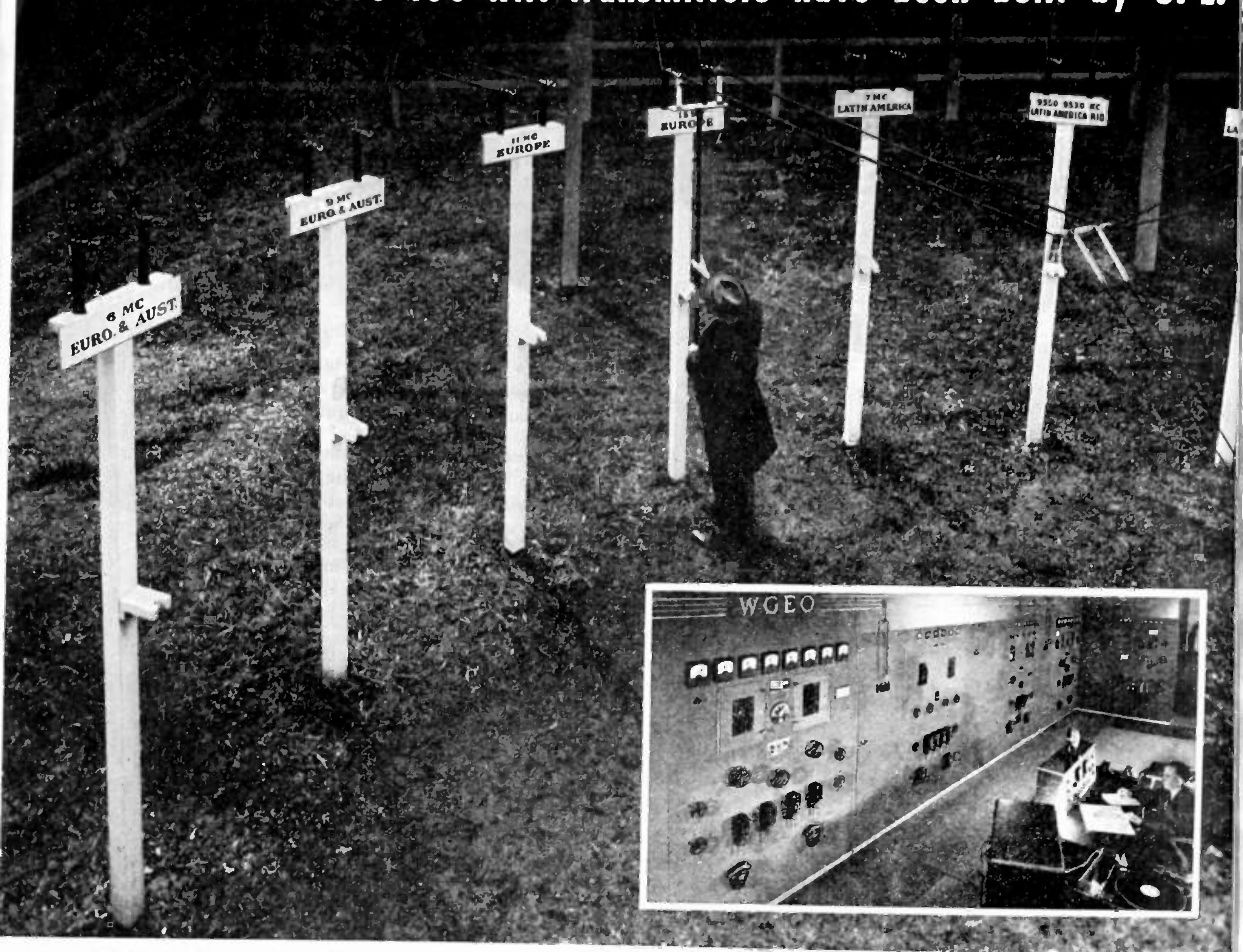
RADIO ... SOUND ... **Rauland** ... COMMUNICATIONS

Electroneering is our business

THE RAULAND CORPORATION ... CHICAGO, ILLINOIS

Buy War Bonds and Stamps! Rauland employees are still investing 10% of their salaries in War Bonds

All of America's 100-kw. transmitters have been built by G. E.



Switchyard at General Electric's 100-kw station, WGEO, in Schenectady, N. Y.

Forceful Allied propaganda is today beamed to all the Axis world by G-E international short-wave transmitters.

G-E pioneering in international short wave, begun in 1923, led to the development of nine American international stations of varying outputs up to 75 kw. Recently, G. E. added four more, two of them of 100 kw, the highest signal output of any American-built stations of that type. G. E. is now building three more giant 100-kw transmitters for the expanding American war needs.

General Electric is the only American manufacturer ever to have successfully designed and built international transmitters of such great power.

The G-E 100-kw and 50-kw transmitters for Station WGEO-WGEA, shown in the insert above, have their programs beamed by special panel-type antennae backed by ingenious dipole reflectors that step up the radiation efficiency.

In the main illustration is another G-E development, adding greatly to the flexibility and efficiency of international equipment. This antenna-feeder hookup gives quick manual switching from one directional beam to another—from one overseas work area to another. Day and night, this powerful station is working for a shorter war, a better peace.

What G-E Leadership Means to You

Informed thinking today points to changes in post-war broadcasting. It looks for a big increase in local FM stations. It foresees fewer but more powerful AM stations, and that television will grow, becoming an important factor in consumer markets.

General Electric offers any broadcaster a complete service in all three fields of FM, AM, television!

1. G. E.'s unmatched achievements in international transmitters are ample evidence of G-E ability to build new high-power AM transmitters and improved receivers after the war.

2. The fact that G. E. has built over a third of all FM broadcast transmitters and a large percentage of FM receivers is positive evidence of its continued leadership in the post-war FM field.

3. And four years of live-talent programming experiment in its own non-commercial television station, WRGB, plus its full line of television transmitters, relay transmitters, studio apparatus, and receivers provide a sum total of television equipment and experience that will be of immense value to the post-war broadcasting industry. . . . *Electronics Department, General Electric, Schenectady, N. Y.*

Tune in "THE WORLD TODAY" and hear the news direct from the men who see it happen, every evening except Sunday at 6:45 E. W. T. over CBS. On Sunday listen to "The Hour of Charm" at 10:00 P. M. E. W. T. on NBC.

• • •

42,000 hours of international service and still going strong! At WGEO, a G-E mercury-vapor rectifier tube — Type 857-B — has given faultless service since 1934.

GENERAL ELECTRIC

160-B6-6916

G-E employees are now purchasing over $1,000,000 in War Bonds weekly

FM · TELEVISION · AM

Flying blind . . . but not deaf

HIGH over obscuring clouds, through the murk of fog and the black of night, the "blind" pilot speeds to his mission . . . and returns . . . almost completely dependent upon what he hears through his headphones. A tremendous responsibility for any piece of equipment.

Making headphones to the exacting standards of the Army and Navy Air Force is one of the wartime tasks of Rola. A pioneer in Radio and later in Electronics, the technical knowledge and the manufacturing skill of this seasoned organization now is devoted exclusively to giving our Fighters in the Air the best, most effective equipment of any in the world. THE ROLA COMPANY, INC., 2530 Superior Avenue, Cleveland, Ohio.

In addition to complete headsets, Rola manufactures transformers and coils of all kinds for aerial communications. If your problem involves Electronics . . . and is important to the war effort . . . why not discuss it with a Rola engineer.

★ ROLA ★

MAKERS OF THE FINEST IN SOUND REPRODUCING AND ELECTRONIC EQUIPMENT

THE "MAGIC BRAIN" OF *Any Electronic Equipment* IS A TUBE...

and the fountainhead of modern tube production is RCA

LET'S LOOK AT THE RECORD!

To a greater degree than any other manufacturer, RCA has solved the problem of turning out in quantity, high quality Tubes expertly engineered to the requirements of the practical user.

In the field of "Receiving" Tubes, RCA is looked to by the government as well as many other critical customers as producing an eminently satisfactory product—and doing it in tremendous quantities.

In Cathode-Ray Tubes, RCA was the first to produce well-engineered units at low cost, thus making them available to industry generally.

RCA has long led in Phototube development, meeting one newer and more exacting application after another. Included here as RCA "firsts" are such outstanding developments as the RCA-931 Multiplier Tube, and other Phototubes utilizing the famous S4 surface which is highly responsive to daylight and blue light.

RCA first developed the Beam Power Tube in the "Receiving" field, subsequently carrying this important development into the field of Transmitting Tubes.

In the field of Ultra-High-Frequency Tubes, RCA was the first to bring out Acorn Tubes—and, later, many others, among them such important Tubes as the RCA-1628 (and its successor, 8012), -825, -827 and RCA-832.

In Power and Transmitting Tubes, RCA was the first to design a successful air-radiator for cooling heavy-duty units.

Time and again, RCA has anticipated the need for new or improved Tube types, often pioneering them far ahead of any commercial demand.

And always RCA has led the way with Application Engineers in the field. These practical men apply their specialized talents to the dual objective of developing better tubes for the job at hand, and finding ways to use tubes already available to better advantage.

BUY U. S. WAR BONDS AND STAMPS

RCA RADIO-ELECTRONIC TUBES

RCA Victor Division, RADIO CORPORATION OF AMERICA, Camden, N. J.

Underwood and Underwood

Ellery W. Stone

Captain Ellery W. Stone, U.S.N.R., who until his recall by the Navy, was president and director of the Postal Telegraph Company, has resigned those posts to enter active duty. Announcement of his re-entry into active Navy service was made May 24, 1943. He had been deferred from active duty during the merger negotiations with the Western Union Telegraph Company.

Captain Stone has just been promoted from the rank of Commander, having previously served frequent tours of duty in the Communications-Liaison Reserve of the Office of Naval Communications and Eastern Sea Frontier. He is now designated for overseas duty as Chief of Staff to Vice-Admiral William Glassford on the latter's mission to French West Africa.

He has been associated with the field of communications since he became a licensed radio operator in 1911 while still a high-school student in Oakland, California. He subsequently attended the University of California where he specialized in electrical and radio engineering. In World War I, he served with distinction as a lieutenant, first junior then senior grade, as District Communications Officer of the 11th Naval District on the Pacific Coast.

He was a pioneer in the radiotelegraph field and organized the Federal Telegraph Company on the Pacific Coast, which company was later acquired by the International Telephone and Telegraph Corporation and became the Mackay Radio Company of California. With the International Telephone and Telegraph Corporation, he successively held a variety of executive posts, including the Executive Vice-Presidency of Mackay Radio, Vice-President of All America Cables and Radio, and he was also in charge of all International Telephone and Telegraph Corporation radio operations. He joined the Postal Telegraph organization as Vice President in 1938.

He has been a member of various committees of the Board of War Communications. He joined the Institute of Radio Engineers as an Associate in 1914, transferred to Member grade in 1916, and became a Fellow in 1924. He is the author of two books on radio, and of various articles on communications. The following papers by Captain Stone has been published in the PROCEEDINGS.

"An Impulse Excitation Transmitter," June, 1916

"Additional Experiments with Impulse Excitation," April, 1917

"Municipal Regulations Covering Radio Stations," December, 1917

The membership of the Institute of Radio Engineers and other engineering readers of the PROCEEDINGS OF THE I.R.E. will doubtless find thoughtful suggestions and helpful inspiration, both leading to increasingly effective wartime and postwar efforts, in expressions of the thoughts of pioneers in the radio industry who have risen to leadership in that field. There is accordingly here presented a statement prepared for the PROCEEDINGS by the President of the Radio Corporation of America who is as well a former Secretary of the Institute. As seems suitable in the publication of such material, it is presented in its original form as received from the author.

The Editor

Saluting the Radio-Electronic Engineer

David Sarnoff

The radio engineer has his hand on the throttle of Victory in an engine of science empowered by the radio-electron tube. Electrons are the fuel; radio is the driving force. While soldiers, sailors, and fliers of the United Nations align their forces to smash the "Fortress Europe," radio research men, engineers, and production workers are electronically allied in the effort.

Radio in this war is a priceless ingredient. As the offensive progresses, the indispensability of radio-electronics becomes more apparent, not only in battle, but in the peace to come. Thousands of radiomen are working at top speed to rush new ideas and inventions into service, and to supply the apparatus that will assist greatly in achieving unconditional surrender of the enemy in Europe and the Orient, as it has done in Africa. The radio engineer's importance was never more significant, for he is performing two vital services—helping to win the war, and laying the foundations of a more enduring peace and a better world.

My faith in the radio research worker and in the engineer goes back a long way, yet not so long as radio's fast pace measures time. It is like looking back to yesterday that I recall how I wore the earphones in the old Marconi station on the sands of Siasconset in 1908, and marveled at the work of the radio engineers who had made it possible for the sparks to flash dots and dashes. My admiration for the men who create, design, and build the apparatus that has kept the cavalcade of radio advancing, has grown with the passing years.

I remember how we anxiously awaited the *Titanic* as the new Queen of the Seas; how we waited to hear its new spark and hoped to catch a glimpse of its wireless station when the big ship came into port. I remember as vividly as if it occurred last night, how shocked we were, how unbelievable it was to hear that the *Titanic* was in distress on her maiden voyage; and then the news that she had plunged to a sepulcher in the sea.

That memorable night in the annals of radio was a turning point for many an American youth interested in science. Wireless from mid-ocean, from a great liner in distress, captured the imagination of many an American boy. From the group of wireless amateurs that sprang up after that terrible night on the air, developed many leading radio engineers, who today look back to 1912 as the date when wireless got into their blood and became their life work.

Behind the operator's fist at the key, behind the microphone or the high-speed transmitter and recorder, stands the engineer. He put Uncle Sam at the forefront of wireless in the First World War, and helped to make the United States the communication center of the world. Since 1918, with new radio-electron tubes produced by research engineers, the operating engineer has blazed new trails through space. He has explored on the wavelengths far into the ionosphere and to every corner of the earth, from North Pole to South Pole. He has opened up new spectrums of the "ether"—electromagnetic regions once looked upon as barren. It was the ingenuity of the radio engineer that transformed them into fertile fields of communication. Today, he stands with radio on the frontier of light.

In America the scientist and engineer enjoys an atmosphere of freedom in which to think and to work. He is fortunate in having radio as a force of unlimited possibilities; a force that is unending. The engineer's work is never finished, for radio itself is measured by infinity.

We are fortunate too, to be living at a time when the Destiny of civilization challenges, and, by answering the challenge as radio engineers are doing, brings joy to their work in a gigantic task. In radio, one of the qualifications for success in achievement is for the mind to be ever active. We call the radio tube a "magic brain." But it is the brain of the engineer that puts the electrons to work—first, to the task of winning the war.

The veteran radio engineer can look back to the old spark days and quickly realize how his profession has contributed to the progress of the world since this century opened. Great are the triumphs of radio in the span of years that separate the first transatlantic signal from television and radar. The youthful engineer whose career may date back to the advent of broadcasting also knows the importance of his profession in the modern world. To him the wireless sparks of yore no doubt seem tame compared with voices that leap the hemispheres, and with television which frames into a picture the glory of the past in radio engineering. But it was in the spark gaps that radio achieved its start, and today the engineer, whose Alma Mater is the school of wireless telegraphy, marches ahead with the younger men of the Radio-Electronic Age, men of the Signal Corps, Naval Communications, and in every phase of radio operations.

A wartime salute is in order for the radio engineer of the past and present, and may his future be one of unparalleled opportunities. I have every confidence that it will be, for the man who thinks and works to keep pace with progress on the wings of radio.

THE INSTITUTE OF RADIO ENGINEERS
INCORPORATED

SECTION MEETINGS

ATLANTA	CHICAGO	CLEVELAND	DETROIT
September 17	September 17	September 23	September 21

LOS ANGELES	NEW YORK	PHILADELPHIA	PITTSBURGH	WASHINGTON
September 21	September 1	September 2	September 13	September 13

SECTIONS

ATLANTA—Chairman, C. F. Daugherty; Secretary, Ivan Miles, 554—14 St., N. W., Atlanta, Ga.

BALTIMORE—Chairman, G. J. Gross; Secretary, A. D. Williams, Bendix Radio Corp., E. Joppa Rd., Towson, Md.

BOSTON—Chairman, R. F. Field; Secretary, Corwin Crosby, 16 Chauncy St., Cambridge, Mass.

BUENOS AIRES—Chairman, J. P. Arnaud; Secretary, Alexander Nadosy, Florida St. 1065, Dept. C-16, Buenos Aires, Argentina.

BUFFALO-NIAGARA—Chairman, Leroy Fiedler; Secretary, H. G. Korts, 432 Potomac Ave., Buffalo, N. Y.

CHICAGO—Chairman, A. B. Bronwell; Secretary, W. O. Swinyard, Hazeltine Electronics Corp., 325 W. Huron St., Chicago, Ill.

CINCINNATI—Chairman, Howard Lepple; Secretary, J. L. Hollis, 6511 Betts Ave., North College Hill, Cincinnati, Ohio.

CLEVELAND—Chairman, F. C. Everett; Secretary, Hugh B. Okeson, 4362 W. 58 St., Cleveland, 9, Ohio.

CONNECTICUT VALLEY—Chairman, W. M. Smith; Secretary, R. E. Moe, Radio Dept., General Electric Co., Bridgeport, Conn.

DALLAS-FORT WORTH—Chairman, H. E. Applegate; Secretary, P. C. Barnes, WFAA-WBAP, Grapevine, Texas.

DETROIT—Chairman, F. M. Hartz; Secretary, E. J. Hughes, 14209 Prevost, Detroit, Mich.

EMPORIUM—Chairman, R. K. Gessford; Secretary, H. D. Johnson, Sylvania Electric Products, Inc., Emporium, Pa.

INDIANAPOLIS—Chairman, S. E. Benson; Secretary, B. H. Rinehart, 1920 Park Ave., Indianapolis, Ind.

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LOS ANGELES—Chairman, Lester Bowman; Secretary, R. C. Moody, 4319 Bellingham Ave., North Hollywood, Calif.

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WASHINGTON—Chairman, C. M. Hunt; Secretary, H. A. Burroughs, Rm. 7207, Federal Communications Commission, Washington, D. C.

Electric Communications, the Past and Present Illuminate the Future*

A Suggestive Interpretation

LLOYD ESPENSCHIED†, FELLOW, I.R.E.

Summary—The paper is intended to stimulate thought. The art of conveying intelligence to a distance is looked at broadly in terms of the underlying physical dimensions of energy, time, and space. These dimensions connect each individual with the outside world. The original information, which is to be transmitted, itself partakes of them as does the system over which the information is to be projected.

One sees in the free electron in vacuous space the ideal means for supplying the appropriate carrying energy and of renewing it en route. In the time dimension, or its inverse, the frequency spectrum, one can trace an underlying trend which courses through the whole history of electric communications and that now leads to very high frequencies and an enormous accretion in intelligence-carrying capacity. In the spacial dimension, of distance and direction, one recognizes not only the conquering of distance but also the conserving of space by the directing and guiding of the radiation, representing further advance in the kinds and amounts of intelligence that may be conveyed. Altogether the trends toward new electronic instrumentalities, the higher frequencies, and wave propagation more sharply demarked in space, spell enlarged opportunities and new horizons.

"May-be the things I perceive, . . . are only apparitions, and the real something has yet to be known."
(Walt Whitman: "Of the Terrible Doubt of Appearances," in the "Leaves of Grass.")

THUS has a great poet alluded to that interesting tentativeness which always resides in our conceptions of the day and to the complementary element of discovery and growth of new knowledge. This is, of course, in the very nature of progress and applies as well to our perception of things at a distance electrically, to the impressions we get and how we get them, to the physical systems we employ, and our interpretation of them. For truly our conception of the art of electric communications, of the means being brought within reach, of what it is that we undertake to do, the scope of it as both a science and a service, is continually changing and rapidly these days.

In considering the subject of electric communications, then, let us not confine ourselves to the existing practices as we ordinarily think of them, telegraphy, telephony, radio, etc., but rather try to get outside and deal with the art as a more general whole in terms of the physical principles and dimensions which underlie it and relate it to our surroundings. By getting a fresh point of view we might see the interesting field in which we work in a somewhat different light, discern trends more nearly in their true perspective, and gain some suggestion, perhaps, of that real something that has yet to be known.

The Underlying Dimensions

We start by recognizing that the projection of intelligence to a distance electrically involves three of the major aspects or dimensions of the physical universe; namely,

Energy or power—in the sense of its being that something which does the carrying of the meaningful pattern.

Time—in the ordinary sense of transition, involving the velocity of propagation and the rate at which the signal patterns are formed.

Space—in the sense of three-dimensional space defining distance and direction.

We may think of these entities as dimensions which define a great imaginary structure or universe centering about each individual. Thereby the individual may gather information of the world about him in scope beyond the capabilities of his ordinary senses, and similarly send out information. Actually it appears that there is much more to take in than there is to send out! In each case the requirement is that the manifestations as related to man must be in such form as to link up with him, as to be capable of stimulating his senses inbound and of permitting him to express himself outbound. Naturally it is desired to link up electrically with as many of man's senses as possible, giving him a general and natural coupling with his environment; from which will be seen something of the great opportunity which exists for development.

If one looks rather carefully at what the art has to tell when viewed in this manner, through the underlying physical facets of the structure as it were, certain flowlines of development will be discerned and a certain basis in reality established for judging the various practical embodiments of the art and even perhaps of future trends. Of course discovery and invention are so much a matter of intermodulation between the physical world and human affairs, and mankind is so very unpredictable, that one cannot hope to foretell the relative values and the actual occurrences of the future. Streaming in from other directions than those which we can now encompass in our consideration are other forces related to man's activity, which in combination with the older trends determine the new

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† Bell Telephone Laboratories, Inc., New York, N. Y.

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In attempting to develop this, a more universal-physical conception of electric communications, we shall observe in the first place the nature of the *original information* which is to be transmitted, in terms of the dimensions of energy, time, and space. Then we shall go on and consider in terms of each of these dimensions the distance-covering telecommunications link itself. In doing this we shall sweep over comprehensively but briefly the whole history of electric communications down to the present time in respect to these elementary elements of energy, time, and space. From this the imaginative ones may be able to discern something of how these dimensional cross sections are likely to reassemble themselves to form the electric communications art of the future.

The Original Message Information

Let us think of the information which we are to send and receive as in the nature of a package which can be picked up, sent over the intervening medium, and delivered. What is the form and nature of this information package in its original appearance and as it must needs be delivered? It appears to be a pattern which exists principally in one of two senses:

(1) In space, as the object itself exists, with all of its parts present simultaneously. Such is the kind of information we pick up with our senses of touch and vision;

(2) In time, a fleeting pattern which develops from instant to instant and which we might say is disposed along the time axis. Such is the manipulating of a telegraph key, the exercise of our vocal cords in speaking.

These are not watertight divisions, for we find that the two forms flow into each other. If in speaking I gesticulate with my hand, your impression will start with a space pattern and merge with one of a time character. In hearing my voice you receive what is primarily a time-disposed pattern and yet through the binaural effect, as well as through your vision, you will place me in space. In that most elementary of our senses, the tactile sense of feeling, the full object is present in a space-pattern sense but we "take it in" in succession as we do visual images.

We realize that our perception of time and space are intimately related, different aspects of our consciousness. When we think of traveling to a distance, the two elements of spacially disposed things and of arriving at them in time succession are intimately associated. But we should not let this intimacy of our two primary dimensions of time and space obliterate the distinction between them in defining the character of the original message pattern itself, whether it is primarily a spacial one or a time-developed one, for this constitutes a useful "handle" with which to grasp an otherwise elusive subject.

Nor need we, in doing this, be unduly perplexed by the fact that the way in which we describe the sending-end information is the way in which it presents itself through our ordinary senses at the receiving end. We simply admit that there is no such thing as the *a priori* existence of this sending-end information in an absolute sense, that our description of it is necessarily limited to the way it stimulates our receiving senses; and then we simply say that in terms of *life*, of our own reactions, this is what the information really *is*.

Then we go on and see that the pattern itself, whether disposed in space or in time, is constituted basically of departures or changes. An infinitely extended uniform surface conveys no information, nor does an indefinitely prolonged uniform sound. It is the changes that mark off the patterns; in the world of intelligence we deal essentially with the transients of space and time. The sharpness of definition, the rate at which intelligence is dispatched, is a matter of the abruptness of the transitions, the width of the spectrum of harmonic frequencies.

Now picture the world of information as it is disposed about the observer. There is in the first place the spacially disposed information composed for example of reflected light, which in terms of energy is there in the first place ready for us to come along, pick up, and interpret. Myriads of rays come pouring in wherever one goes from a world of tremendous detail, in large part a fixed world from which one may help himself at will to this and that portion. Then in this background of fixity there play the forces of motion and of life. We hear noises; we see the motions of clouds, the movement of the leaves; in fact the entrance of the observer himself into the field of view represents the setting-up of motion and in turn the generation of the time pattern.

The observer is attracted by animate life, and particularly by his fellow man who proves to possess the most highly developed kind of instant-to-instant-generated intelligence. By his appearance, expressions, and gestures, and more especially by his vocal cords and writing, this fellow man conveys his own reaction to his surroundings, his *thoughts*, by a process of codification which we call language. Although this man-made information is no different in its physical form from that which we gather from the surrounding world, it is more highly developed in meaning because attached to the most complex assembly of mechanism the world affords, namely, man himself. We are, of course, tremendously interested in this man-to-man kind of

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THE INSTITUTE OF RADIO ENGINEERS

INCORPORATED

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ATLANTA	CHICAGO	CLEVELAND	DETROIT
September 17	September 17	September 23	September 21

LOS ANGELES	NEW YORK	PHILADELPHIA	PITTSBURGH	WASHINGTON
September 21	September 1	September 2	September 13	September 13

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communication, itself of both the space-disposed and the time-disposed kind, especially the latter. But note also that this is only a part of the information world with which we are surrounded. Thus far the practice of electric communications has been concerned principally with this man-generated language kind of information, outstandingly the aural form of it which is essentially time-disposed information. Such man-made information is transferred out of the body through muscular actions—facial expressions, gestures, actuation of vocal mechanism, and writing—mechanical actions the fundamental period of which is relatively slow, of the order of one tenth of a second. Perhaps the fact that we are limited by these muscular actions in what we send out is a wise provision of nature, for it seems that we cannot really masticate our ideas in the first place any more rapidly, and it would be unfortunate were we able to flood the world with undigested brainstorms! Or is it that we are at present actually encumbered by having to express our thoughts thus in these mechanical forms and that communication would be freed by the avoidance of this obstacle? At least these are considerations we shall need to keep in mind when it comes to tapping the brain for more direct electrical transmission.

The Carrying Energy

In undertaking to transmit intelligence to a distance one seeks immediately some effect or source of energy which naturally propagates rapidly to a distance with not too great loss and which is capable of "taking" the information. Light is such a medium or vehicle, and in accordance with our modern ideas of it, as a wave-propagation-corpuscular phenomenon, is electrical. In fact it was the discovery of the remarkably high speed of propagation of electricity along wires, approximating the velocity of light, which first excited interest in electricity as a messenger of intelligence in the 1700's.

On the basis of the identity of light and electricity we can regard the early visual or semaphore type of telegraph, from which the word telegraph has come, to be the first electrical system for reading graphic signs or writing at a distance.[1] Indeed we can go a step farther and say that our everyday vision is an electrical system of communication. The messages we receive are those set up all around us in a spacially disposed world by the reflection of waves in a certain portion of the electromagnetic-wave spectrum.

[1] In its original meaning the word telegraph applied to the writing or displaying of graphs or signs at the *sending* station which was viewed from the receiving end by a telescope. In its electrical adaptation it has been skewed around to apply to the writing down of the information at the receiving end. This has occasioned some ambiguity. In accordance with this interpretation if the receiving operator does not record the message but merely receives by ear and passes it along by telephone, there has been no writing at a distance! Actually in the sense of taking the original graph, a surface-disposed pattern, and reproducing it for reading at the receiving end, the facsimile or television type of transmission is more truly in accordance with the original meaning of the word than is one which operates on the basis of the time-disposed Morse code and which permits of being read by ear rather than the eye.

The form of energy which man himself generates in expressing his thoughts through muscular exertions of one kind and another is essentially one of mechanical motion. These motions are taken off and fed to our senses either

(1) Through the medium of light, wherein we note that they become time modifications of the space-pattern kind of intelligence, or
(2) As continued outward mechanical impulses, as in punching keys or in beating the air with our vocal cords, wherein the intelligence pattern is developed along the time axis.

The latter way of continuing the man-made information beyond the body, through mechanical waves, propagates relatively slowly and with a very high attenuation and is not ordinarily useful for distances beyond a few tens of feet. This is in fact the very information in which we are most interested to transmit to a distance for it conveys so much of man's thought. To do the conveying to considerable distances we must ride this mechanical-action kind of time pattern upon a more fleet and far-reaching carrier—that of electricity, and we do it electromechanically.

We recognize that what gave birth to the age of electric communications was the discovery of electricity in a form which could be reasonably conserved in being propagated to a distance, namely, the galvanic or voltaic electricity of the low-impedance battery as distinguished from the preceding electrostatic machines of impossibly high impedance. With this kind of electricity came the discovery of electromagnetism whereby there became available the electromechanical instrument with which the silent messenger is linked with man's mechanical actions and senses.

In the absence of other instrumentalities the art continued for years on the basis of using as the carrying energy a direct current taken from a battery. And even today, after having learned how to develop high-frequency carriers and implant the intelligence on them, we seem always to start out with the primary source of electricity one of direct current or very low frequency. Once having started our technique with electricity generated at the low-frequency end of the scale, it seems we cannot do other than start there anew even though we wish to land high up in the frequency spectrum. In our engines for getting this low-frequency energy in the first place we start with heat, which itself represents high-frequency oscillations—of the molecules. How long is it going to take us to make a short cut from the energy of heat, directly to the frequencies that we want for transmission? Certainly everyone knows today that the power pack, etc., the primary supply for vacuum tubes, is the bane of one's existence and one of the reasons why the vacuum-tube amplifier is still rather expensive and clumsy in use.

The idea of transmitting intelligence in relays, through a chain of stations, must be as early as the marathon race itself. Before the advent of electric

events. Naturally one is always doing a certain amount of wondering and predicting about the future especially in such a lively and promising art as the present one, and under wartime stimulation. Each is free to do this on his own account; but the writer on his part forbears being a prophet. It should be mentioned also that the interpretation of the art which is here offered by its very nature is a personal one to the writer and not anything for which the communications agency with which he is associated should be held responsible!

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(1) In space, as the object itself exists, with all of its parts present simultaneously. Such is the kind of information we pick up with our senses of touch and vision;

(2) In time, a fleeting pattern which develops from instant to instant and which we might say is disposed along the time axis. Such is the manipulating of a telegraph key, the exercise of our vocal cords in speaking.

These are not watertight divisions, for we find that the two forms flow into each other. If in speaking I gesticulate with my hand, your impression will start with a space pattern and merge with one of a time character. In hearing my voice you receive what is primarily a time-disposed pattern and yet through the binaural effect, as well as through your vision, you will place me in space. In that most elementary of our senses, the tactile sense of feeling, the full object is present in a space-pattern sense but we "take it in" in succession as we do visual images.

We realize that our perception of time and space are intimately related, different aspects of our consciousness. When we think of traveling to a distance, the two elements of spacially disposed things and of arriving at them in time succession are intimately associated. But we should not let this intimacy of our two primary dimensions of time and space obliterate the distinction between them in defining the character of the original message pattern itself, whether it is primarily a spacial one or a time-developed one, for this constitutes a useful "handle" with which to grasp an otherwise elusive subject.

Nor need we, in doing this, be unduly perplexed by the fact that the way in which we describe the sending-end information is the way in which it presents itself through our ordinary senses at the receiving end. We simply admit that there is no such thing as the *a priori* existence of this sending-end information in an absolute sense, that our description of it is necessarily limited to the way it stimulates our receiving senses; and then we simply say that in terms of *life*, of our own reactions, this is what the information really *is*.

Then we go on and see that the pattern itself, whether disposed in space or in time, is constituted basically of departures or changes. An infinitely extended uniform surface conveys no information, nor does an indefinitely prolonged uniform sound. It is the changes that mark off the patterns; in the world of intelligence we deal essentially with the transients of space and time. The sharpness of definition, the rate at which intelligence is dispatched, is a matter of the abruptness of the transitions, the width of the spectrum of harmonic frequencies.

Now picture the world of information as it is disposed about the observer. There is in the first place the spacially disposed information composed for example of reflected light, which in terms of energy is there in the first place ready for us to come along, pick up, and interpret. Myriads of rays come pouring in wherever one goes from a world of tremendous detail, in large part a fixed world from which one may help himself at will to this and that portion. Then in this background of fixity there play the forces of motion and of life. We hear noises; we see the motions of clouds, the movement of the leaves; in fact the entrance of the observer himself into the field of view represents the setting-up of motion and in turn the generation of the time pattern.

The observer is attracted by animate life, and particularly by his fellow man who proves to possess the most highly developed kind of instant-to-instant-generated intelligence. By his appearance, expressions, and gestures, and more especially by his vocal cords and writing, this fellow man conveys his own reaction to his surroundings, his *thoughts*, by a process of codification which we call language. Although this man-made information is no different in its physical form from that which we gather from the surrounding world, it is more highly developed in meaning because attached to the most complex assembly of mechanism the world affords, namely, man himself. We are, of course, tremendously interested in this man-to-man kind of

communication, itself of both the space-disposed and the time-disposed kind, especially the latter. But note also that this is only a part of the information world with which we are surrounded. Thus far the practice of electric communications has been concerned principally with this man-generated language kind of information, outstandingly the aural form of it which is essentially time-disposed information. Such man-made information is transferred out of the body through muscular actions—facial expressions, gestures, actuation of vocal mechanism, and writing—mechanical actions the fundamental period of which is relatively slow, of the order of one tenth of a second. Perhaps the fact that we are limited by these muscular actions in what we send out is a wise provision of nature, for it seems that we cannot really masticate our ideas in the first place any more rapidly, and it would be unfortunate were we able to flood the world with undigested brainstorms! Or is it that we are at present actually encumbered by having to express our thoughts thus in these mechanical forms and that communication would be freed by the avoidance of this obstacle? At least these are considerations we shall need to keep in mind when it comes to tapping the brain for more direct electrical transmission.

The Carrying Energy

In undertaking to transmit intelligence to a distance one seeks immediately some effect or source of energy which naturally propagates rapidly to a distance with not too great loss and which is capable of "taking" the information. Light is such a medium or vehicle, and in accordance with our modern ideas of it, as a wave-propagation-corpuscular phenomenon, is electrical. In fact it was the discovery of the remarkably high speed of propagation of electricity along wires, approximating the velocity of light, which first excited interest in electricity as a messenger of intelligence in the 1700's.

On the basis of the identity of light and electricity we can regard the early visual or semaphore type of telegraph, from which the word telegraph has come, to be the first electrical system for reading graphic signs or writing at a distance.[1] Indeed we can go a step farther and say that our everyday vision is an electrical system of communication. The messages we receive are those set up all around us in a spacially disposed world by the reflection of waves in a certain portion of the electromagnetic-wave spectrum.

[1] In its original meaning the word telegraph applied to the writing or displaying of graphs or signs at the *sending* station which was viewed from the receiving end by a telescope. In its electrical adaptation it has been skewed around to apply to the writing down of the information at the receiving end. This has occasioned some ambiguity. In accordance with this interpretation if the receiving operator does not record the message but merely receives by ear and passes it along by telephone, there has been no writing at a distance! Actually in the sense of taking the original graph, a surface-disposed pattern, and reproducing it for reading at the receiving end, the facsimile or television type of transmission is more truly in accordance with the original meaning of the word than is one which operates on the basis of the time-disposed Morse code and which permits of being read by ear rather than the eye.

The form of energy which man himself generates in expressing his thoughts through muscular exertions of one kind and another is essentially one of mechanical motion. These motions are taken off and fed to our senses either

(1) Through the medium of light, wherein we note that they become time modifications of the space-pattern kind of intelligence, or
(2) As continued outward mechanical impulses, as in punching keys or in beating the air with our vocal cords, wherein the intelligence pattern is developed along the time axis.

The latter way of continuing the man-made information beyond the body, through mechanical waves, propagates relatively slowly and with a very high attenuation and is not ordinarily useful for distances beyond a few tens of feet. This is in fact the very information in which we are most interested to transmit to a distance for it conveys so much of man's thought. To do the conveying to considerable distances we must ride this mechanical-action kind of time pattern upon a more fleet and far-reaching carrier—that of electricity, and we do it electromechanically.

We recognize that what gave birth to the age of electric communications was the discovery of electricity in a form which could be reasonably conserved in being propagated to a distance, namely, the galvanic or voltaic electricity of the low-impedance battery as distinguished from the preceding electrostatic machines of impossibly high impedance. With this kind of electricity came the discovery of electromagnetism whereby there became available the electromechanical instrument with which the silent messenger is linked with man's mechanical actions and senses.

In the absence of other instrumentalities the art continued for years on the basis of using as the carrying energy a direct current taken from a battery. And even today, after having learned how to develop high-frequency carriers and implant the intelligence on them, we seem always to start out with the primary source of electricity one of direct current or very low frequency. Once having started our technique with electricity generated at the low-frequency end of the scale, it seems we cannot do other than start there anew even though we wish to land high up in the frequency spectrum. In our engines for getting this low-frequency energy in the first place we start with heat, which itself represents high-frequency oscillations—of the molecules. How long is it going to take us to make a short cut from the energy of heat, directly to the frequencies that we want for transmission? Certainly everyone knows today that the power pack, etc., the primary supply for vacuum tubes, is the bane of one's existence and one of the reasons why the vacuum-tube amplifier is still rather expensive and clumsy in use.

The idea of transmitting intelligence in relays, through a chain of stations, must be as early as the marathon race itself. Before the advent of electric telegraphy we find it carried out in the visual or semaphore type of systems, of necessity because the span of transmission was limited to line-of-sight distances, a reminder of ultra-high-frequency radio today. The conception of electrical relay stations appears to have arisen with the very birth of the electromagnetic telegraph. We find it in one of Morse's early patents.[2] It appears that on the route which Morse's original telegraph line took between Baltimore and Washington, the first stagecoach station out of Baltimore where the horses were changed and fresh horsepower was taken on was named "Relay House," and that this had to do with the origin of the term relay as applied to the electromagnetic telegraph.[3]

But whatever the origin of the term, the action itself, of infusing new energy along the route, is basic and vital to the whole art. It is something to which wire transmission lends itself naturally. In radio transmission one tends to depend more largely upon the original sending-end energy in order to jump as far as possible before coming down to earth again. To ultra-high-frequency radio systems the relay idea lends itself for the reason that such systems begin to approach the guided-wave method of transmission by giving access to the new energy along the line of travel.

It appears that the greater the finesse which attaches to the sending-end information, the weaker is the energy taken off the original intelligence pattern and the greater the need for a sending-end amplifier. In the earliest form of electric communication, namely telegraphy, the muscular motions involved in manipulating the telegraph key represented the availability of considerable energy with which to modulate the source of power. When it came to the telephone, the available signaling power, that of sound waves, was very much less and there was difficulty in getting enough energy into the line and on to the receiving end to be discernible to the ear. The telephone transmitter-microphone represented the early supplying of a transmitting amplifier, which really insured the success of the early telephone. More recently in electric communications we have come to the picking-up of spacial patterns by light waves and here the energy available originally from the pattern is so extremely weak as to require the immediate insertion of a high degree of amplification before anything can be done.

We well recognize that the greatest advance in electric communications in our time is the appearance of the vacuum-tube amplifier. It puts a new face on the whole art in several respects. In the first place it is capable of supplying the carrying power remarkably flexibly over a very wide range of frequencies, and of enabling the information to be impressed thereon. It does this by virtue of being an amplifier, and its more fundamental significance is as a straight amplifier where we can look at it in two senses. One is the fact that it can be used at the sending end to pick up extremely weak original patterns, to maintain them along the route, and to deliver them at the receiving end to our senses at the power required for convenient consumption. The other is the fact that it can do this amplifying at extremely high frequencies of transmission. In this latter respect the vacuum-tube amplifier has in effect rolled back from the frequency spectrum the curtain of attenuation which has hung across it, and has revealed a great new dimension into which the art lately has been expending. Frequency being a subdivision of time, this matter of the frequency spectrum and its significance to the art is discussed in the section to follow.

What makes the vacuum tube such a remarkable instrument is the fact that the minute electrical carriers, the electrons, are obtained by themselves in free space, free from the encumbrance of the solid matter conductors in which they had been contained heretofore in electrical devices. Electricity is obtained in pure and unadulterated form, as it were, out in a bit of vacuous space where it can be operated upon directly by electrical influences such as those of the received signal waves, without the intervention of mechanical action with its inertia limitations. No wonder then that we find ourselves in the midst of a great new era in electric communications, as we shall see in terms of the frequency-spectrum picture given subsequently, and one which is much broader than electric communications itself.

The Time-Dimension—Frequency Spectrum

The first thing that occurs to one in thinking of the element of time in electric communications is the speed with which communication may be accomplished. The electrical mode of signaling is unique in this respect. But this matter of the speed of transmittal contains two aspects which are distinct and should not be confused. First there is the velocity with which the carrying wave stream is projected over the intervening medium, which is essentially that of light. This high speed of travel is especially important in long-distance telephony with its requirement of immediacy of receipt and response in two-way talking. The interval between *go* and *come* is a matter of split seconds, and there are occasions of simultaneous talking in which one speaker *breaks* the other. In the case of wire telephony the velocity at voice frequencies is materially less than that of light, especially over loaded cable circuits, and the higher velocity which characterizes the higher frequencies on nonloaded wires is one of the reasons why carrier transmission is becoming so

[2] U. S. Patent No. 1647, June 20, 1840.

[3] There exists an interesting account of this and the change which occurred in the meaning of the word relay, from a station to a particular device for working into a local circuit, and of the use of the word repeater to designate the working from one line section to another, written by a pioneer, Thomas D. Lockwood, in an article which appeared back in the 90's in three sections, entitled "Telephone repeaters or relays, and repeating systems," *Elec. World*, vol. 20, pp. 593–597; November 14, 1896; pp. 626–630; November 21, 1896; and pp. 660–661; November 28, 1896.

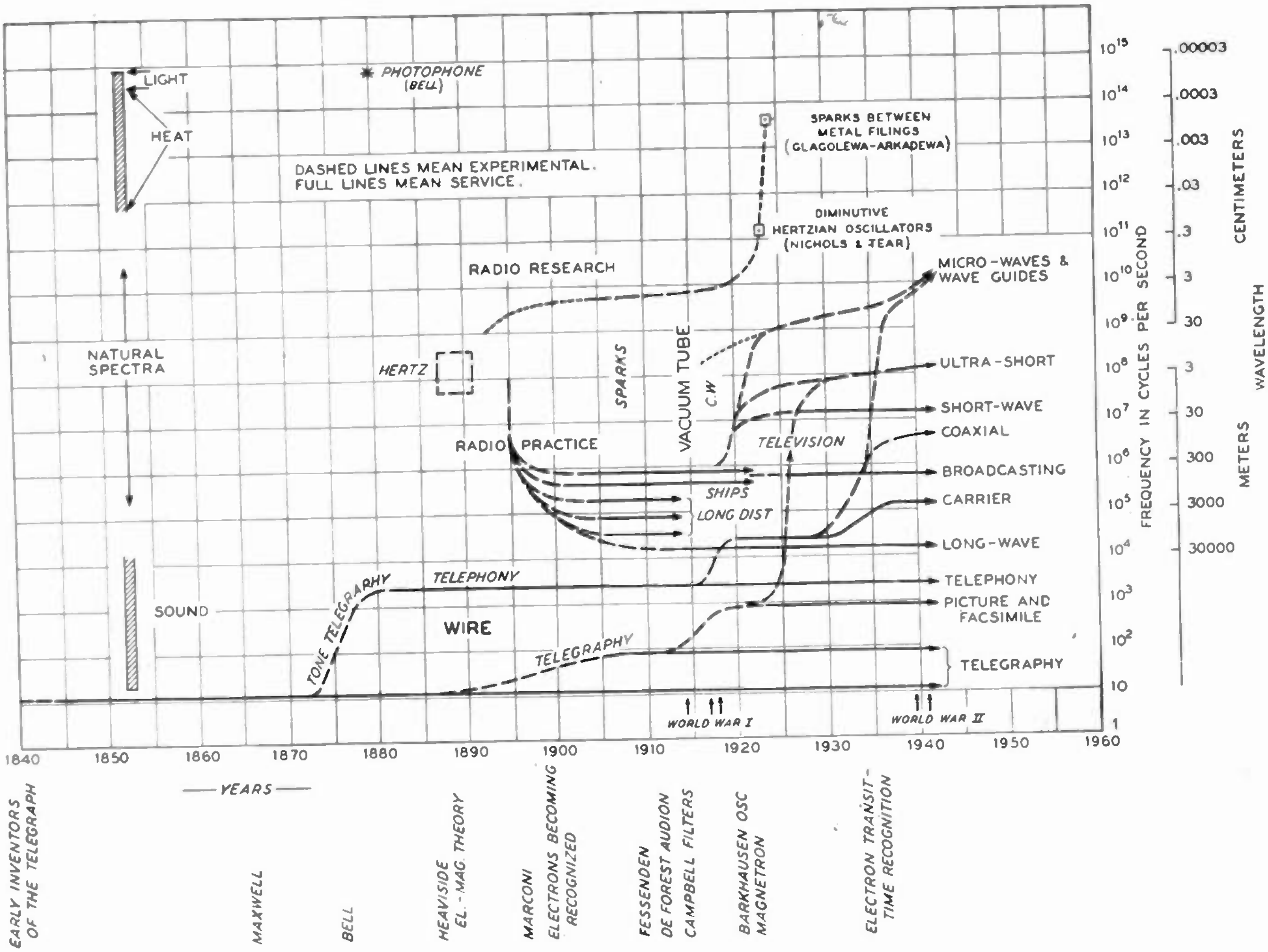

Fig. 1

extensively employed for the longer distances.

The other factor concerned with the speed at which communication may be conducted electrically pertains not to the time of travel between the two terminals but to the rate at which the signaling pattern is gotten off at the sending end and gathered together again at the receiver. This rate of modulation determines the extent to which the sinusoidal carrier wave is made to depart from its original smooth form by the superposed signal wave, to occupy a *band* of signaling frequencies. The faster the rate of signaling the wider the band or frequency aperture represented by the transmission. This bandwidth is substantially constant for telephony, fixed by the nature of our vocal system and hearing. In telegraphy the bandwidth is pared down to a minimum in the first place by the condensing of the code, but is directly proportional to the speed of keying. In picture transmission and television there is no economizing in the number of signal elements by coding, all the dot points are taken as they exist, and the bandwidth is on the whole much greater, again being proportional to the rate at which the dot elements are gotten off, to the rate of scanning over the surface. Here we have a surface-disposed pattern which by scanning is converted to a time-disposed one and then back again. For purposes of television the rate of dispatch of the picture frames is so high as to require a very wide signaling band.

It is a part of the technique of carrier and radio to take such signaling bands and place them anywhere desired in the broad frequency spectrum by the act of modulation. The modulation process is an interesting one in itself, the more so since the signaling wave can be impressed upon the carrier in any of several ways, amplitude modulation, frequency modulation, etc. Also the wedding of the two, the signaling and the carrier wave, can be done at low or high power levels, etc. The end result is the placing of the signaling bands in the broad frequency spectrum of electromagnetic waves. We can, then, look at the art from this standpoint of the placing of any and all kinds of signaling bands on top of each other in the broad frequency spectrum of the common medium.

Fig. 1 presents a historiograph of electric communications in these terms, of the broad frequency spectrum. It is a cross section of the whole art of electric communications, from its beginning about a century ago down to the present time, in terms of this one dimension of the rates of vibration of the terminal devices and of the carrying waves.

The relationship is between the years of the art plotted horizontally and the frequency spectrum vertically. At the left, vertically are the portions of the spectrum with which our senses are concerned, the slits into which we need to feed the final received signals to discern them. It is to be understood that beneath the audio frequencies and extending down to zero frequency are the impulses concerned with our tactile sense and our muscular exertions. In the curves the interrupted lines represent the research and experimental phase of developments and the full lines the undertaking of practice and the giving of service.

Horizontally along the bottom of the diagram is the base line of the art, representing hand-speed telegraphy of about 10 cycles per second, which ushered in electric communications about a century ago. The rate of signaling has been speeded up somewhat over the years until we might say that telegraphy today represents frequencies of the order of 10 to 100 cycles fundamental-dot frequency. Technically much higher speeds of signaling have been attained but the use of such higher-speed systems has been limited. Rather than a few very high-speed channels the main requirement has been for many low-speed ones because of the necessity of transcribing by hand from written texts into the telegraph and from the telegraph back to written texts. With the attainment of more automatic means for effecting this translation, as by electrophoto methods, this trend may be changed. More recently telegraphy has grown into picture transmission and in turn into television and has moved on up to the higher frequencies of radio and the coaxial cable.

Referring again to the chart, telephony will be seen to have been the first major step beyond telegraphy. Here two branches of knowledge merged, that of acoustics, diaphragm devices, and the like, and that of speeded-up electromechanical telegraph instruments, whereby sound-wave and electrical-wave paths were joined together. Telephony represented an enormous speeding-up of the rates of response of the terminal instruments of telegraphy at the time, much as had the realization of the electromagnetic telegraph instruments themselves represented increased agility over the ponderous movement of the signal arms of the old visual telegraph systems. The higher line frequencies of telephony meant a corresponding increase in attenuation, whereby it became necessary to build more carefully constructed lines for conserving the energy and to deal with much weaker received power. The attention to wave propagation over wires that was called forth in telephony, and previously in the submarine-telegraph-cable problem, led gradually to the establishment of the great technique which for some years now has underlain the whole electric communications art, that of electrical networks. It starts with an evaluation of wave propagation over wires in terms of attenuation and phase and goes into the whole line of technology represented by artificial lines and balancing networks, by the recognition of the band character of the signaling wave, by equalization, impedance matching, and by the invention of the electric wave filter—all forming a great technical base much as has the land-line network itself constituted the service base of the art.

Not long after the invention of the telephone there occurred the physical beginning of radio in the experiments of Hertz. He dealt with frequencies of the order of hundreds of megacycles, generated discontinuously by means of sparks, frequencies which were extremely high at the time for electrical generation, and which only recently have been much exceeded on a continuous wave basis. Stemming from Hertz's work will be seen two lines of development. The upper interrupted line leading to still higher frequencies represents the pure-research pursuit of electromagnetic waves which succeeded finally in joining the radio spectrum with that of radiant heat, still by means of electromagnetic waves generated by sparks.

The downwardly sloping line out of Hertz represents the development of practical radio. The frequency trend was initially rapidly downward, to the longer waves, for the reason that the effort was to stretch the transmission over longer distances. This involved bending the waves around the curvature of the earth, larger antennas, greater sending powers meaning larger condensers, and naturally lower rates of vibration. The lower frequencies meant in turn greater persistence of oscillation and led to new methods of generating the waves, such as the arc and the machine alternator, whereby there began the continuous-wave phase of radio.

It was at this juncture, just before the incidence of World War I, that the vacuum-tube amplifier came into the picture, more or less simultaneously in America and Europe through de Forest and von Lieben and Reisz, through both the radio and the wire arts. The attainment of the high-vacuum form of tube in the Bell Telephone Company's Laboratories, and also independently in those of the General Electric Company, marked the great beginning in the practical phase of the electronic amplifier. This occurred just prior to the start in Europe of World War I, high-vacuum amplifier tubes going into service for the first time in 1913. In radio, the vacuum tube quickly spelled the beginning of practical radiotelephony, and equally quickly in wire communication the beginning of high-frequency or carrier-current telephony, both starting about 1914 and reaching considerable technical attainment before the entry of the United States into the war. In time the vacuum tube converted radio completely to continuous-wave transmission and led long-distance wire telephony rather generally to the higher frequencies.

The up-surges in frequency that are shown in the chart starting around 1920 followed after World War I but were by no means due to the war. The genesis lay in the great technical base that was established before the war, starting years before in electron physics and the vacuum tube and in the great underlying technique of wave propagation and circuit networks. It is true that the war did promote certain developments. We are not conscious of what it may have held back, but do know that it gave early experience in manufacturing vacuum tubes in some quantity and in building little radiotelephone sets which marked the beginning of small-boat telephony and of aircraft radiotelephony. Another important bearing of the war was the quick indoctrinating of thousands of bright young men with knowledge of the new vacuum-tube art whereby they became enthusiastic workers in the developments to follow, notably in the launching and carrying out of broadcasting.

The curves show rather dramatically the great upward surge in frequency which has occurred between the two great wars. The art has entered the present war with an enormously expanded front as to frequencies and types of services, in the technical phenomena and tools that were becoming available. With this greater physical front to choose from and the greater effort and duration of the present war, the technology may be expected to come out of it more greatly altered from its normal course than was the case previously, some parts delayed but many greatly advanced. War may be likened to a huge amplifier, a highly selective or distorting one, which greatly enhances certain components and suppresses others; an amplifier which by bare overloading, as it were, generates some interesting by-product harmonics, and in this way contributes to a certain widening of the front! We may, then, expect to enter the next peace period with greater opportunities than ever both in the physical means at our disposal and in the brains and youthful enthusiasm that will have been attracted to this interesting field.

Having dealt with the frequency dimension as a subdivision of time, we should recall before leaving the element of time that we are interested also in bridging *long* time intervals by the storage of the intelligence. In telegraphy and picture transmission we start out with a stored intelligence and we reproduce it as such at the receiving end. The period of preservation can be anything from short duration, as per the fleeting figure on a cathode-ray tube, to impressions made upon a very long-life holding medium such as paper and the phonograph record. The facility with which we can now produce signals of ample energy and place them more or less anywhere in the frequency spectrum enlarges the opportunity for record-making both as to the kind of information and the manner of recording it. The invention of multiple-type printing about the year 1450 constituted one of the greatest advances in intelligence conveyance in general and also in the broadcasting aspect of it up to the time of radio broadcasting. Perhaps now, with all our electrical facilities, we are on the verge of another great new era as regards the record, both for individual and mass use and for local as well as distant recordation. Certainly those of us who struggle with a pen or the pounding of a typewriter would relish a neater way of extending our thoughts in time for later eye consumption.

The Space Dimension

The fact that certain kinds of information, such as that of vision, are of a space-pattern character already has been noted. This kind of information, that which has cross section and streams out from a surface as it were, it not limited to visible rays as we well know today from the use of reflected radio waves for the detecting of objects. This streaming-out kind of information can be used also in another sense which we see exemplified in the ultra-high-frequency marker of an airport. Here the beam extending upward sets up a vertical demarcation having a cross-sectional dimension, which the airplane intercepts. Obviously the principle could be extended to include additional information by more complex patterns. In fact with the growing occupancy of the air we may well expect a sort of compartmentalizing of the space over the earth's surface for purposes of controlling navigation by sharply directive radio. We shall be fortunate if the advertising signs of the future planted on the earth's surface, to be seen from the aerial highways, are not of this character to insure that they can be seen by the passing air-borne throng irrespective of weather conditions!

This brings us to a consideration of space on a grander scale, viewed as the great gap across which we wish to transmit intelligence. Let us cast our eyes at the heavens and think of electromagnetic waves that are being propagated out in interstellar space. There being no guides there is only one way the propagation can take place and that is radially from the source, along the straight lines we call rays. In the stars we see myriads of such radiators each emitting spherically, the whole galaxy setting up a crisscross of countless rays, a network or exchange system of waves in free space, on a colossal scale. It is from this system that we gain our knowledge of the greater universe, of the orientation of the heavens, of the speed of light itself, of the similarity of the remote bodies to our own as told by the light spectra, and of more subtle goings-on not yet understood represented by the spectral shift and the fact that it increases with distance. The distances of this interstellar system are so enormous that electromagnetic-wave propagation seems slow. It depends solely upon sending-end energies, and these are so great that although our own earth intercepts but a minute part of that of our sun we get enough to "keep us going." Indeed this interstellar system is the grandest and most lavish kind of intercommunicating system as it were; yet for us human beings it lacks the one element of the time-disposed pattern of information that ties up with life.

Imagine now that we come down to earth and take up man-made radiating systems of the radio variety. We find immediately that our little radiating oscillator cannot continue its propagation in all directions, for the radiation hits an impenetrable barrier in the earth's surface. A measure of shielding and reflecting immediately arises. In fact we can provide reflecting surfaces at our will, set up compartmentalizing walls as it were. By disposing a reflecting wall around the radiating source, energy dissipation in undesired directions can be stopped off and the transmission in desired directions reinforced. The desired transmission can be throttled down to a narrow cone. We can go farther and extend the shielding surface on down the line of the cone in the form of a pipe whereby the radiation is further restricted to a very narrow cross section *a la* wire transmission.

Thus in our imagination we have gone gradually from a pure radio system radiating in three-dimensional space down to almost a single-ray path, to a tube of transmission to which the wave energy is confined and along which it is reflected to a particular destination.

In this approach we see radio as something which starts out in keeping with its name as a radiant space-occupier but which at the higher frequencies can be narrowed down to a certain general bundle of rays that reach a limited area. We see the guided-wave method of transmission as a closed-in ray as it were which frees space for other occupancies and which enables the path of transmission to be guided around corners and obstacles and to be projected through solid material by boring a hole through it as it were, through walls and along under the surface of the earth itself as in laying a cable. In both cases the conveyance of energy is through space, the difference being the extent to which the spacial dimension is closed in, physically how this is done, and the sharpness of the lines of demarcation. It would seem that altogether between the two extremes of the wide openness of the original form of radio and the complete closed-in nature of the original form of a shielded pair of wires, quite a wide variety of possibilities exists.

Now we come in our consideration of the spacial dimension to what is perhaps the controlling physical fact of our existence on the earth. It is that we are surface creatures who live in that transition plane between the solid matter of the earth and the things around us, and the great open space above. We have two general realms with which to deal, first that of the solid matter all about us, to penetrate which with our spacial wave we must in effect bore a hole. This is what the tubular wave-guide or wire-line kind of thing does. It involves work and expense but provides a medium which is removed from the great outside world and in this sense is private; the price that one pays for this threadlike definition in the form of increased attenuation with increased frequency can be offset by an increased use of amplifiers. The other realm with which we are concerned is that of the open space itself above the earth's surface, now becoming so very important for aerial navigation, for which the radio method is vital and unique. Then there is an intermediate condition which because of its practical importance amounts really to a third realm. It is that of the horizontal dimension skimming along the earth's surface. One can traverse distance immediately outside the earth's surface either by the wave-contained method, by cable or open-wire line suspended above the surface, or by the wave-projected method of radio, or conceivably by a combination of or intermediate condition between the two. Which of the two general methods to employ obviously depends upon the particular purpose in view. For fixed point-to-point communication the closed-in private medium is obviously preferred, if for no other reason than that it prevents the cluttering-up of space generally for the purpose of particular communications. On the other hand for mobile services within the area of the movement, obviously radio is the answer. For effecting a general coverage as in broadcasting, radio likewise is indicated, although not uniquely, for wires are readily spun out into a network.

Conclusion

In recent years the electric-communications technique has been acquiring a greatly expanded physical base, which is characterized by:

Electronics, representing a new command of electricity in the form of carriers free in vacuous space capable of energizing communication systems at enormously high rates;

A greatly extended frequency dimension, representing enlarged intelligence-carrying capacity and a new technique common to the radio and guided method of transmission;

Means in the form of wave guides and directed radio for realizing transmissions of very high frequency characterized by a high degree of spacial definition, whereby the number of communications may be multiplied on a space-segregation basis;

A trend toward undertaking the transmission of additional forms of intelligence, particularly of the space-pattern type.

The new phenomena and instruments of today are the harbingers of the new services of tomorrow, from which we realize that our existing plants and services are the steppingstones to the attaining gradually of a more complete conveyance of intelligence between man and his surroundings. The picture of the future appears to be that of great networks of transmissions covering in general each area of the earth's surface and indeed the globe, with innumerable island systems representing local interests, and myriads of sprouts feeding out into open space in keeping with man's increased mobility and conquest of the air; a world nervous system so comprehensive in terms of time and space and intelligence-carrying capacity as eventually to enable one anywhere to keep in touch, by more or less all his senses and to the extent desired, with his environment and fellow man.

Direct-Reading Wattmeters for Use at Radio Frequencies*

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Summary—The principle upon which the operation of direct-reading wattmeters is based has been known for many years. The contribution of this paper lies in the application of this principle to a practical operating instrument for the measurement of high power at radio frequencies.

Two instruments are described. The first is useful in the range of frequencies from 500 to 2000 kilocycles. This instrument contains circuits which permit operation at any frequency in this band with no tuning or other change in the instrument.

The second instrument operates in the region near 50 megacycles. It is inherently a single-frequency device constructed from sections of transmission lines.

The theory of operation is discussed, as well as calibration methods. Test data taken with the instruments with loads having a wide range of power factors are compared with power measurements made on water-cooled loads.

I. Introduction

In the past, the output power of a radio broadcast transmitter has been determined by either one of two methods, the direct and the indirect. In the indirect method, the plate input power of the last stage of the transmitter is multiplied by a stated factor to obtain the output power. This method is not extremely accurate for a number of reasons. The direct method consists in measuring the antenna resistance or the resistance at some point in the feed system, and inserting an ammeter at this point. Then the power is determined from the I^2R formula. This method of power determination assumes that the antenna resistance or the feed-point resistance does not change. If such a change occurs due to, say, climatic conditions, in the first case, or to circuit changes made either intentionally or inadvertently, the mere fact that the current at the measuring point remains constant does not assure a constant power.

At ultra-high frequencies, the measurement of power into an antenna has always presented a difficult problem. As in the low-frequency case, the constancy of an ammeter reading need not at all mean that the power is constant.

It is the purpose of this paper to describe two instruments which have been developed. The first is a wattmeter for use in the band of frequencies between 500 and 2000 kilocycles. The second wattmeter, operating on the same principle, is useful in the neighborhood of 50 megacycles.

II. The Principle of Paired Thermocouples

The operation of these wattmeters centers upon two identical thermocouples. These thermocouples are of the vacuum type, with the heater insulated from the junction. Let us suppose that in some manner we induce currents in the heaters of each couple as shown in Fig. 1(a). These currents are equal in magnitude and

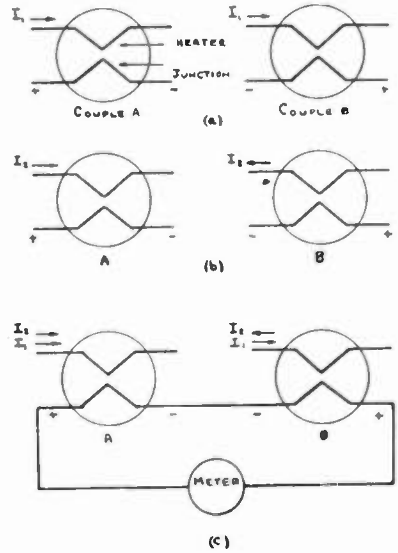

Fig. 1—Fundamental circuits involved in the principle of paired thermocouples.

phase. We will designate these currents by $\bar{I}_1$. Then (Fig. 1(b)) we induce another current $\bar{I}_2$ in each heater but make the current in heater B flow in the opposite direction to the current in heater A. It is assumed that the method of obtaining $\bar{I}_1$ and $\bar{I}_2$ is such that these currents are entirely independent of each other. Then the current flow is as shown in Fig. 1(c), where heater A carries the sum of the two currents, while heater B carries the difference of the currents. The meter is connected so that the junction voltages oppose each other.

If the couples follow the normal square law, the deflection of the meter due to one couple is

$$D = KI^2 \tag{1}$$

where I is the absolute value of the current flowing in the heater and K is a constant determined by means of the circuit shown in Fig. 2, where the current is passed through heater A, while no current is passed through heater B.

Returning now to Fig. 1(c), let us assume that I_2 leads I_1 by an angle θ. Then the current in heater A is

$$\bar{I}_A = \bar{I}_1 + \bar{I}_2 = I_1 + I_2 \cos\theta + jI_2 \sin\theta. \tag{2}$$

The current in heater B is

$$\bar{I}_B = \bar{I}_1 - \bar{I}_2 = I_1 - I_2 \cos\theta - jI_2 \sin\theta. \tag{3}$$

* Decimal classification: 621.374.6×R251. Original manuscript received by the Institute, January 11, 1943. Presented, Winter Conference, New York, N. Y., January 28, 1943.

† RCA Laboratories, Princeton, New Jersey.

The square of the absolute value of each of these currents is

$$I_A^2 = I_1^2 + I_2^2 + 2I_1I_2 \cos\theta \qquad (4)$$

and

$$I_B^2 = I_1^2 + I_2^2 - 2I_1I_2 \cos\theta. \qquad (5)$$

Because of the manner in which the meter is connected

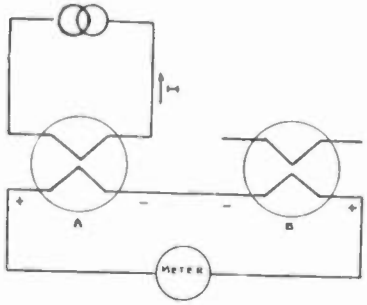

Fig. 2—Circuit for determining the thermocouple constant.

in Fig. 1(c), the deflection will be, from (1), (4), and (5),

$$D = D_A - D_B = K(I_A^2 - I_B^2) = 4KI_1I_2 \cos\theta. \quad (6)$$

If now we feed power to a load and fulfill the following conditions,

1. I_1 is always proportional to the load voltage,
2. I_2 is always proportional to the load current, and
3. the phase angle θ between I_1 and I_2 is equal to the phase angle between the load voltage and the load current,

we see that the deflection as indicated by (6) will always be proportional to the power into the load.

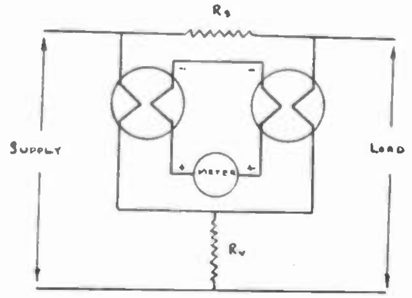

Fig. 3

The principle of the paired couples shown in (6) was disclosed by Bauch[1] and his treatment was later expanded by others.[2–5] The elementary circuit as shown by Bauch is repeated here in Fig. 3.

Applications of this same principle where the pair of thermocouples is replaced by a pair of vacuum tubes used as square-law detectors have been described in the literature.[6–8]

III. The Circuit of the Low-Frequency Wattmeter

The circuit used in the low-frequency wattmeter is shown in Fig. 4. The current I_1 which is set up by the load voltage passes through the resistor R_1 and through each of the thermocouple heaters. The current I_2 is obtained by coupling to the feed line by means of identical but oppositely polarized current transformers. R_1 is

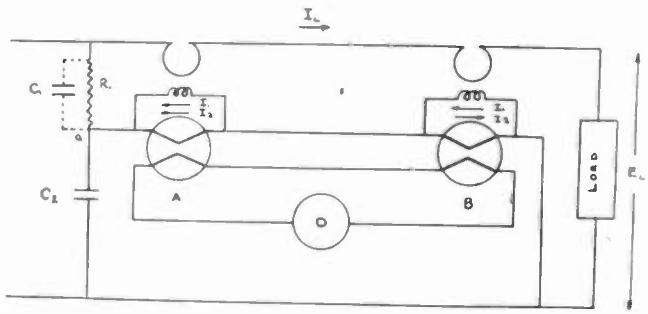

Fig. 4—Circuit of the low-frequency wattmeter.

a rather high resistance, formed by a metallized surface on a ceramic tube. C_1 is the small stray capacitance across the resistor R_1 to the point shown. Any stray capacitance from the top of the resistor to ground plays no part in the functioning of the wattmeter, since it acts simply as a shunt on the load. C_2 is a compensating capacitor which will be considered later.

Since the current transformers are connected in opposition, these circuits will set up no voltage between the point a and ground so that I_2 is not affected in the least by the choice of R_1 and C_2.

The action of the current transformers will be better

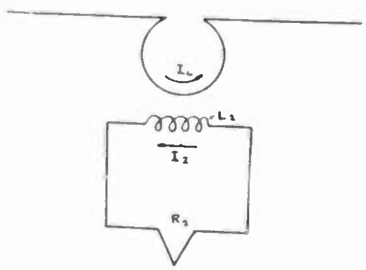

Fig. 5—The current-transformer relations.

understood by an examination of Fig. 5. The current I_2 is

$$\bar{I}_2 = \frac{j\omega M\bar{I}_L}{R_2 + j\omega L_2} \qquad (7)$$

[1] R. Bauch, "Hot-wire wattmeter," *Elek. Tech. Zeit.*, vol. 24, pp. 530–536; July 9, 1903.
[2] E. Kuhnel, "Thermowattmeter," *Zeit. für Instrumentenk*, vol. 48, pp. 127–130; March, 1928.
[3] G. Zickner and G. Pfestorf, "Loss measurements on large condensers," *Elek. Tech. Zeit.*, vol. 51, pp. 1681–1684; December 4, 1930.
[4] B. A. Sharpe, "Thermal wattmeter," *Jour. Sci. Instr.*, vol. 10; pp. 318–321; October, 1933.
[5] F. A. Laws, "Electrical Measurements," second edition, 1938, McGraw-Hill Book Co., New York, N. Y., p. 325.

[6] E. Peterson, U. S. Patent 1,586,533, June, 1926.
[7] H. M. Turner and F. T. McNamara, "An electron tube-wattmeter and voltmeter and a phase shifting bridge," Proc. I.R.E., vol. 18, pp. 1743–1748; October, 1930.
[8] E. Mallett, "A valve wattmeter," *Jour. I.E.E.* (London), vol 72, pp. 295–302; September, 1933.

where M = the mutual inductance of the current transformer
L_2 = the self inductance of the secondary circuit
R_2 = the resistance of the secondary circuit, with practically all the resistance residing in the heater
$\omega = 2\pi f$
f = the frequency of the source.

Equation (7) may be written

$$\bar{I}_2 = \frac{j\omega M \bar{I}_L}{j\omega L_2\left(1 + \frac{R_2}{j\omega L_2}\right)} = \frac{M}{L_2} \frac{\bar{I}_L \angle \beta}{\sqrt{1 + \left(\frac{R_2}{\omega L_2}\right)^2}} \qquad (8)$$

where

$$\tan \beta = \frac{R_2}{\omega L_2}.$$

If we keep $R_2/\omega L_2$ very small compared to unity, we see that $\bar{I}_2$ is proportional to the load current and substantially independent of frequency. However, we find that $\bar{I}_2$ leads the load current by a small angle β, which is a function of frequency. If $R_2/\omega L_2$ should be as large as 0.1, the radical in (8) would differ from unity by only 0.5 of 1 per cent, but the angle β would be equal to 5.75 degrees. However, it should be easy to keep the ratio well below this figure.

Fig. 6 shows the circuits which determine the current I_1. Since R_1 is a large resistance and R_2 so very small in comparison, the current I_0 drawn from the source is essentially independent of R_2, L_2, and C_2. Then

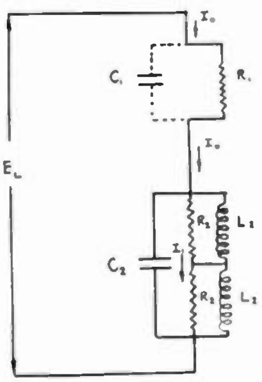

Fig. 6—The equivalent "proportional-to-voltage" circuit.

$$\bar{I}_0 = \bar{E}_L\left[\frac{1}{R_1} + j\omega C_1\right] = \frac{\bar{E}_L}{R_1}\,[1 + j\omega C_1 R_1]. \qquad (9)$$

Also,

$$\bar{I}_0 = 2R_2\bar{I}_1\left[\frac{1}{2R_2} + \frac{1}{j2\omega L} + j\omega C_2\right]$$

$$= \bar{I}_1\left[1 + \frac{R_2}{j\omega L_2} + j2\omega C_2 R_2\right]. \qquad (10)$$

Equating (9) and (10),

$$\bar{I}_1 = \frac{\bar{E}_L}{R_1} \cdot \frac{[1 + j\omega C_1 R_1]}{\left[1 + \frac{R_2}{j\omega L_2} + j2\omega C_2 R_2\right]}. \qquad (11)$$

For the moment, let us consider the expression

$$\left[1 + \frac{R_2}{j\omega L_2}\right][1 + j2\omega C_2 R_2]$$

$$= 1 + \frac{R_2}{j\omega L_2} + j2\omega C_2 R_2 + \frac{2R_2^2 C_2}{L_2}. \qquad (12)$$

Since the second and third terms on the right-hand side of (12) are smaller in comparison to unity, and since the fourth term is the product of the second and third terms, we may drop the fourth term. With this assumption, we may substitute (12) in (11) and obtain

$$\bar{I}_1 = \frac{\bar{E}_L}{R_1} \frac{[1 + j\omega C_1 R_1]}{\left[1 + \frac{R_2}{j\omega L_2}\right][1 + j2\omega C_2 R_2]}. \qquad (13)$$

We now make a choice of C_2, and make

$$2C_2 R_2 = C_1 R_1 \qquad (14)$$

and (13) becomes

$$\bar{I}_1 = \frac{\bar{E}_L}{R_1} \cdot \frac{1}{\left[1 + \frac{R_2}{j\omega L_2}\right]} = \frac{\bar{E}_L \angle \beta}{R_1\sqrt{1 + \left(\frac{R_2}{\omega L_2}\right)^2}} \qquad (15)$$

where

$$\tan \beta = \frac{R_2}{\omega L_2}.$$

From (15) and (8), we see that I_1 is independent of frequency to the same extent as I_2. Also I_1 leads the load voltage by the same angle β that I_2 leads the load current, so that if θ is the phase angle between load voltage and load current, the phase angle between I_1 and I_2 will also be θ.

IV. Method of Calibrating

The instrument may be set up and calibrated with ordinary laboratory instruments. The power source need not have sufficient power to deliver the full-scale watts to be indicated by the meter. In fact, the source need supply only a few watts. After calibration, there will be a certain value of load resistance with zero reactance at which I_1 and I_2 will be equal and in phase. Under this condition, the current in heater A will be twice I_1 while the current in heater B will be zero. For purposes of illustration, we will assume this load to be 100 ohms. The direct-current meter used across the junctions has a full-scale reading of 200 microamperes. We now propose that this meter shall read 100 microamperes when the power into the load is 1000 watts.

If the angle θ is zero, we may find I_1 and I_2 for this condition from (6) to be

$$I_1 = I_2 = \sqrt{\frac{D}{4K}}. \tag{16}$$

In the pair of matched thermocouples used in this wattmeter, the maximum rated current of each heater is 0.10 ampere, with a heater resistance (R_2) of 2.3 ohms. For these particular couples, where the heater current

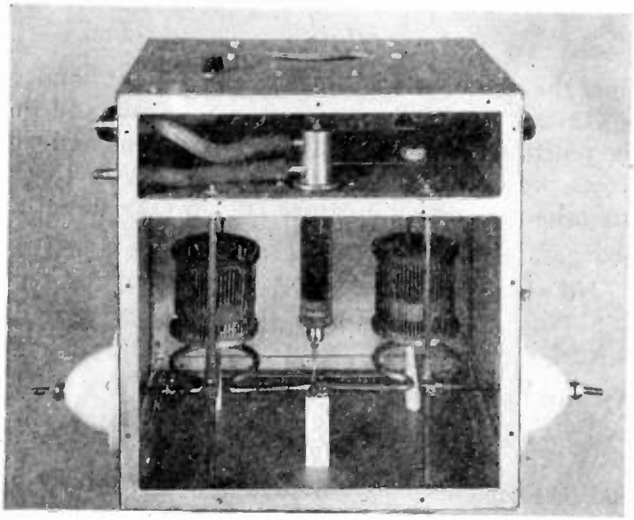

Fig. 7—The broadcast-frequency wattmeter.

is given in amperes and D is expressed in microamperes, the constant $K = 67{,}500$. Then from (16),

$$I_1 = I_2 = \sqrt{\frac{100}{4 \times 67{,}500}} = 0.019 \text{ ampere.}$$

Now, for a power of 1000 watts and a load resistance of 100 ohms and zero reactance, the load voltage will be 316 volts, with a load current of 3.16 amperes. Then from (15)

$$R_1 = \frac{316.}{0.019} = 16{,}600 \text{ ohms.}$$

Before placing this resistor in the circuit, we make the calibration for I_2. A radio-frequency ammeter is used for a load. This ammeter has an essentially zero impedance. A current of 3.16 amperes is now passed through the primary of the two current transformers to the ammeter. Each current transformer is then adjusted so that a current of 0.019 ampere flows in the secondary and the associated heater. This current may be measured by placing the microammeter directly across the individual junctions. When this adjustment has been made, the microammeter is connected across the two junctions in opposition to see that the meter reads exactly zero. If the reading is not zero, further adjustments of the current transformers must be made. This particular adjustment requires a great many trials, until some skill is acquired.

Next, a low-loss capacitor of approximately 100 ohms reactance at the calibrating frequency is used for a load. Sufficient current is supplied to this capacitor until about 1000 volt-amperes are reached. The resistance R_1 is now placed in the circuit. Because of the stray C_1, the meter will probably read some small value. The compensating capacitor C_2 is then adjusted until the meter reads exactly zero.

The wattmeter is now calibrated and ready for use. It is interesting to note that while it was assumed that the load resistance of 100 ohms was fed with a power of 1000 watts, it was not necessary to have such a resistor nor to feed power to it.

Under the calibrating condition, we assumed that I_1 and I_2 were each equal to 0.019 ampere. Then heater A carried 0.038 ampere. The safe current carried by this heater is 0.10 ampere.

V. Test Data of the Low-Frequency Wattmeter

The wattmeter as assembled and calibrated to read 100 microamperes with a load power of 1000 watts is shown in Fig. 7. After a number of laboratory experiments, the wattmeter was tested under full load for a number of load resistances, with varying amounts of power factor. Tests were made in which the power was dissipated in a water-cooled resistor. Measurements of water flow and temperature rise of the water gave an accurate measure of power into the load. It was found that the wattmeter performed satisfactorily under all the conditions imposed. The accuracy of the meter was also checked by measurements of the load resistance with a radio-frequency bridge. A calibrated ammeter in series with the load was then used to measure the power.

Fig. 8 shows wattmeter readings (multiplied by 10)

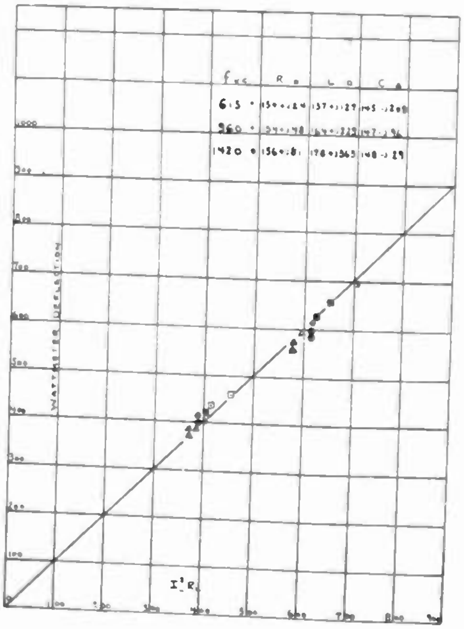

Fig. 8—Response of the broadcast-frequency wattmeter to a variety of load conditions and frequencies.

as a function of load power for three frequencies and for three conditions of power factor.

A wattmeter of the paired thermocouple type was installed at a radio station which operated with a directional antenna system consisting of three towers. The transmitter power was 1000 watts, on a frequency of 920 kilocycles. The antenna system was operated in the directional condition. The wattmeter was first placed in the common-feed point and indicated 1000 watts. Then the meter was placed in each of the three transmission lines with the following results.

TABLE I

Transmission Line	Meter Reading (watts)
Center	640
East	290
West	75
	1005

Here the individual powers added up to agree within 0.5 of 1 per cent with the total power.

Next, the east and west tower were floated, and power was fed to the center tower alone. Under this condition, the center tower had a resistance of 20.6 ohms. The following readings were taken.

TABLE II

Antenna Current (amperes)	Meter Reading (watts)	I^2R (watts)	Error (per cent)
6.97	1000	1000	0
6.50	880	870	1.15
5.47	620	616	0.65
4.97	510	511	−0.196
3.58	270	264	2.27

The transmitter was also modulated with a 1000-cycle tone to various known degrees, and observations made of the increases in the wattmeter reading. It was found that the meter readings increased in exact accord with the theoretical relation between power output and percentage modulation.

VI. Operating Limitations of the Low-Frequency Wattmeter

The wattmeter has been operated successfully at frequencies between 500 and 2000 kilocycles. No attempt has been made to operate outside this band of frequencies, but it is believed that a much wider frequency range is feasible.

Another point which must receive attention in safeguarding the thermocouples is the variation in current in the heaters for a wide range of loads. Under calibrating conditions, we set the currents I_1 and I_2 each equal to 0.019 ampere. Thus with a power of 1000 watts into the 100-ohm load which is free of reactance, these two currents add directly to give a heater current I_A equal to 0.038 ampere. As the load becomes reactive and the load resistance changes, the current I_A increases in magnitude. Care must be taken to see that this heater current does not exceed 0.100 ampere.

Fig. 9 shows the variation of I_A as a function of the load resistance. The solid curves are on the basis of a constant-power-factor angle θ while the dotted curves are for a constant load reactance. The data in Fig. 9 are based on a constant power into the load. Since I_A in the calibrate condition is 0.038 ampere, and the safe current is 0.100 ampere, the safe operating region is below the horizontal line which has an ordinate value

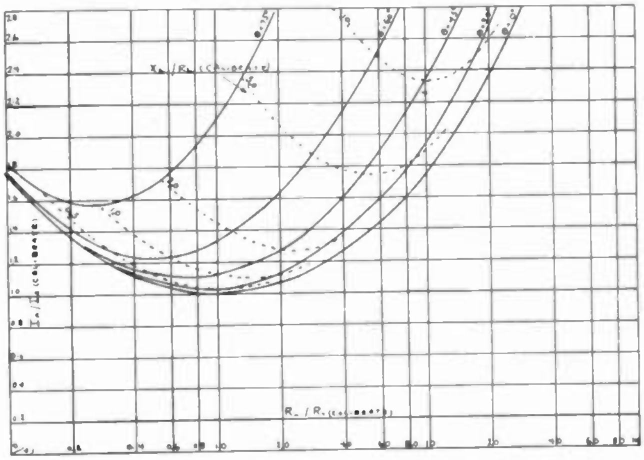

Fig. 9—Variation of heater current with load impedance.

of 2.63. To safeguard the heater properly, it is best to switch the meter so that it is across the junction of thermocouple A alone as a measure of the current in heater A.

VII. Theory of the Ultra-High-Frequency Wattmeter

In the ultra-high-frequency wattmeter, the principle of paired thermocouples is used exactly as in the low-frequency wattmeter described above. The difference in the two instruments lies in the method of coupling the thermocouples to the transmission line. The thermocouples used in this instrument had a maximum allowable heater current of 1 ampere. This allowed the use of a heater resistance so low that the thermocouple placed across a concentric transmission line became a very effective short circuit. The thermocouples were mounted in a specially bored brass block in which the necessary transmission lines terminated so that as much lead inductance as possible was eliminated. This mounting, as well as a shielded case around the meter, was used to eliminate undesired radio-frequency currents which might endanger the junctions of the couples.

The currents I_1 and I_2 were obtained with concentric lines in the manner illustrated in Fig. 10. The power from the transmitter was fed in at H and passed out to the antenna at H'. The transmission-line passing point L was enlarged to facilitate mechanical construction. A characteristic impedance equal to the main transmission-line characteristic impedance should be maintained through the enlarged section. A half-wave section of $\frac{7}{8}$-inch line U was connected in series with the inner conductor of line H–H'. A short-circuiting plug at S_1 a half wave from line H–H' made the series impedance very low, zero if there were no losses in the line. The outer conductor of line U is the inner conductor of line C. This line C is short-circuited at S_3 a quarter wave from L. Thus line C places a very high impedance across line H–H'. For all other lines, a $\frac{3}{8}$-inch transmission line is suitable.

The velocity of propagation v was taken into account

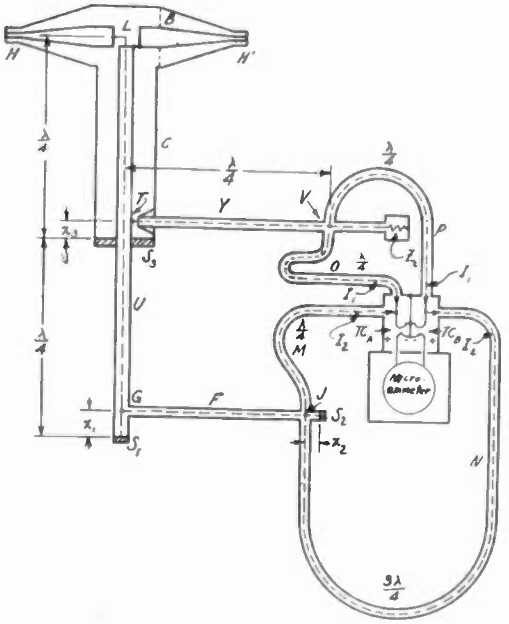

Fig. 10—Circuit diagram of the ultra-high-frequency wattmeter.

in calculating line lengths. v_0 is the velocity and λ_0 the wavelength in free space. The physical length to which the lines are cut is determined by $\lambda = (v/v_0)\lambda_0$.

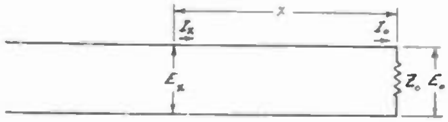

Fig. 11

The current $\overline{I}_x$ and the voltage $\overline{E}_x$ along a transmission line with negligible losses are (Fig. 11)

$$\overline{I}_x = \overline{I}_0 \cos\frac{2\pi x}{\lambda} + \frac{\overline{E}_0}{Z_c} j \sin\frac{2\pi x}{\lambda} \tag{17}$$

$$\overline{E}_x = \overline{E}_0 \cos\frac{2\pi x}{\lambda} + j\overline{I}_0 Z_c \sin\frac{2\pi x}{\lambda} \tag{18}$$

Z_c is the characteristic impedance of the line, $2\pi x/\lambda$ is in radians, and λ is a wavelength of the line measured in the same units as x. If Z_0 the line-terminating resistance is made zero, $\overline{E}_0$ is zero and for lines C, U, F, M, N, O, and P

$$\overline{I}_x = \overline{I}_0 \cos\frac{2\pi x}{\lambda} \tag{19}$$

$$\overline{E}_x = j\overline{I}_0 Z_c \sin\frac{2\pi x}{\lambda}. \tag{20}$$

If $Z_0 = Z_c$ as for line Y

$$\overline{I}_x = \overline{I}_0\left(\cos\frac{2\pi x}{\lambda} + j\sin\frac{2\pi x}{\lambda}\right) \tag{21}$$

$$\overline{E}_x = \overline{E}_0\left(\cos\frac{2\pi x}{\lambda} + j\sin\frac{2\pi x}{\lambda}\right). \tag{22}$$

In examining phase relations and voltage and current magnitude it should be observed that lines F, M, N, O, and P place very high impedances across lines at points G, J, and V because F, M, N, O, and P are a quarter or odd multiples of a quarter wavelength and terminated in short circuits.

The following relations show how the desired thermocouple currents are obtained. It is assumed that the lines all have the same characteristic impedance Z_c. This condition is not necessary but is assumed here for purposes of simplification. From (20),

$$\overline{E}_V = j\overline{I}_1 Z_c \sin\frac{\pi}{2} = j\overline{I}_1 Z_c. \tag{23}$$

From (22) and (23)

$$\overline{E}_T = \overline{E}_V\left(\cos\frac{\pi}{2} + j\sin\frac{\pi}{2}\right) = j\overline{E}_V = -\overline{I}_1 Z_c. \tag{24}$$

From (20) and (24)

$$\overline{E}_T = j\overline{I}_{S_3} Z_c \sin\frac{2\pi x_3}{\lambda} = -\overline{I}_1 Z_c \tag{25}$$

and

$$\overline{E}_L = j\overline{I}_{S_3} Z_c \sin\frac{\pi}{2} = j\overline{I}_{S_3} Z_c. \tag{26}$$

Substituting from (26) in (25)

$$\overline{E}_L \sin\frac{2\pi x_3}{\lambda} = -\overline{I}_1 Z_c$$

or

$$\overline{I}_1 = -\frac{\overline{E}_L}{Z_c}\sin\frac{2\pi x_3}{\lambda}. \tag{27}$$

So we have found that each thermocouple heater has a current I_1 flowing through to ground which is proportional to line voltage and in phase with line voltage E_L. From (20)

$$\overline{E}_J = j\overline{I}_2 Z_c \sin\frac{\pi}{2} = j\overline{I}_2 Z_c \tag{28}$$

where I_2 is the current through $TC-A$. Also

$$\overline{E}_J = j\overline{I}_{S_2} Z_c \sin\frac{2\pi x_2}{\lambda}. \tag{29}$$

Also from (20)

$$\overline{E}_G = j\overline{I}_{S_1} Z_c \sin\frac{\pi}{2} = j\overline{I}_{S_1} Z_c. \tag{30}$$

Eliminating between (28), (29), and (30)

$$\overline{E}_J = \overline{E}_G \sin\frac{2\pi x_2}{\lambda} = j\overline{I}_2 Z_c \tag{31}$$

or

$$\overline{I}_2 = -\frac{j\overline{E}_G}{Z_c}\sin\frac{2\pi x_2}{\lambda}. \tag{32}$$

From (20)

$$\overline{E}_G = j\overline{I}_{S_1} Z_c \sin\frac{2\pi x_1}{\lambda}. \tag{33}$$

From (19)

$$\overline{I}_L = \overline{I}_{S_1}\cos\pi = -\overline{I}_{S_1} \tag{34}$$

so

$$\overline{E}_G = -j\overline{I}_L Z_c \sin\frac{2\pi x_1}{\lambda}. \tag{35}$$

Substituting (35) in (32), we find

$$\overline{I}_2(TC-A) = -\overline{I}_L \sin\frac{2\pi x_1}{\lambda}\sin\frac{2\pi x_2}{\lambda}. \tag{36}$$

The $\overline{I}_2$ through $TC-B$ results from changing (28) to

$$\overline{E}_J = j\overline{I}_2 Z_c \sin\frac{3\pi}{2} = -j\overline{I}_2 Z_c$$

so that

$$\overline{I}_2(TC-B) = \overline{I}_L \sin\frac{2\pi x_1}{\lambda}\sin\frac{2\pi x_2}{\lambda}. \tag{37}$$

It has thus been shown that the current in one thermocouple is the sum of $\overline{I}_1$ and $\overline{I}_2$ while the current in the other couple is the difference of these two currents. It has been shown further that these two currents are proportional to and in phase with line voltage and line current, respectively. This is the relation sought for the wattmeter action.

The wattmeter may be calibrated for higher power by making x_1, x_2, and x_3 smaller. When the power becomes so high that these dimensions can be made no smaller, additional quarter-wave lines may be used at points T and J in the manner that line F was used at point G. For lower power, x_1, x_2, and x_3 may be made larger. It should be borne in mind that increasing x_3 increases the loading on line H–H'.

VIII. Calibration of the Ultra-High-Frequency Wattmeter

A preliminary calibration of the wattmeter may be made with a low-power oscillator of 50 watts. Suppose a full-scale deflection of 2000 watts is desired. The wattmeter will then indicate 1000 watts as half scale on the meter, a reading of 100 microamperes. If the main feed line, with a 70-ohm characteristic impedance, feeds 1000 watts to a 70-ohm load, the line voltage will be $E_L = \sqrt{PR} = \sqrt{1000\times 70} = 264$ volts and the line current will be $I_L = E_L/R = 264/70 = 3.78$ amperes. If the current I_1 and I_2 are made equal for a 70-ohm load, from (16),

$$I_1 = I_2 = \sqrt{\frac{D}{4K}} = \sqrt{\frac{100}{4\times 1225}} = 0.143 \text{ ampere}.$$

Now the tapping positions should be adjusted so a current of 0.143 ampere is produced through each couple by either an E_L of 264 volts or an I_L of 3.78 amperes acting alone.

After adjusting x_1, x_2, and x_3 to calculated values, we make I_2 zero by disconnecting line F at point G. Voltage equal to E_L may be measured a half wave back along the main transmission line from point L. Then the tapping point T is adjusted until 0.143 ampere appears in each thermocouple.

Next we reconnect the line F, and disconnect lines O and P at point V. With a short-circuiting plug at B, we measure the voltage E a quarter wave back along the main transmission line from point L. By making this voltage E equal to 264 volts, we make the current through the short-circuiting plug equal to 3.78 amperes. We then adjust tapping point G or J to adjust each thermocouple current to 0.143 ampere. When this last calibration was made, it was found necessary to make a slight adjustment in the length of M or N to make the couples give equal deflections.

Now the wattmeter should show full-scale deflection for 2000 watts into the load, when the wattmeter is connected normally.

IX. Performance Tests

After preliminary adjustments, the wattmeter was tested with a 1000-watt frequency-modulation transmitter operating on a frequency of 44.9 megacycles.

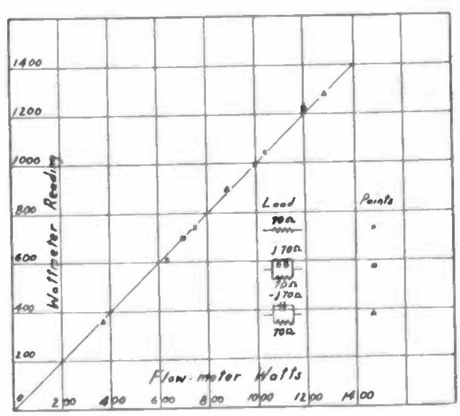

Fig. 12—Test data of the ultra-high-frequency wattmeter.

The calibration was checked against a flowmeter operating on a water-cooled resistor. There was a small error with reactive loads which was corrected by adjusting the length of the line Y between the points T and V. Phase shift in this line should be 90 degrees.

The terminating impedance was made very nearly equal to the characteristic impedance of line Y so that phase shift would be close to linear with length.

The wattmeter deflections as a function of flowmeter measurements are shown in Fig. 12. The reactive loads used caused approximately a 2.5/1 standing wave on the transmission line. Neither 70 ohms of inductive reactance nor 70 ohms of capacitive reactance shunted across the load caused a measurable change in power readings. A 1000-volt-ampere reactance load caused a deflection of only −2 watts. Thus, it is well established that the wattmeter properly takes care of power factor.

The length of transmission line between the load and the wattmeter was changed with no apparent effect.

During these tests the frequency-modulation transmitter was operated fully modulated and entirely unmodulated with no difference in readings.

Since the operation of the wattmeter depends upon the properties of quarter- and half-wave transmission lines, the instrument is inherently a single-frequency device. Tests made with frequencies between 40 and 50 megacycles show that the instrument develops errors of 5 per cent when the carrier frequency is more than 0.5 megacycle away from the calibrating frequency.

X. Conclusion

The writers believe that the instruments described in this paper are practical, have been conclusively proven to be accurate, and should find many valuable applications. Not only will they be of value for transmitter-power-output measurements in field service, but they should also facilitate transmitter development and other laboratory work.

A Wide-Band Oscilloscope*

ELLSWORTH D. COOK†, FELLOW, I. R. E.

Summary—This paper deals with the development of a wide-band oscilloscope intended for use as a precision-measurement tool. The means by which the performance specifications were determined are outlined, and a discussion is given of the methods by which these objectives were obtained.

The various technical problems, such as compensations for both low and high frequencies, the design of attenuator and inverting networks, and test methods, as well as the theory of operation of certain of the components are discussed.

One of the essential differences between power and radio engineering lies in the wave shape of the voltages and currents normally used in each field. The objectives of radio engineering require more detailed knowledge of such wave shapes in general. While considerable information can be derived from indirect measurements, the need for direct visualization is at once evident from an examination of the present-day problems found in television and other fields requiring similar technique.

The work of Crookes and Thomson suggested to Hess a means for the accomplishment of this as early as 1894. Braun produced such a device in 1897. Devices of a similar nature were produced by Dufour, Thomson, and Wood, and in 1905 Wehnelt made a substantial improvement by employing a hot cathode. Subsequent workers have improved both the cathode-ray tube and oscilloscope until television developments have made it an indispensable tool. It was with the thought of producing a cathode-ray oscilloscope especially suited for the more accurate measurement problems of the times that the development of this device was undertaken sometime ago. The performance specifications, which were to be in the nature of a performance guarantee, were quite rigid. They were determined from the average requirements of a number of representative oscilloscopic problems.

Since the size, weight, and cost of such a device are functions of the area of the screen image and the frequency pass band of the deflection circuits, it was apparent that some concessions would have to be made to mobility. The specifications finally chosen were decided upon from the following considerations. In the first place, the amount of detail desired in the complete screen picture together with ease of measurement indicated that a faithful, flat-screen image of about 10 by 10 centimeters should be possible. This, together with the upper limit of sweep frequency (to be discussed later) decided that the No. 914 cathode-ray tube was to be used, since the technique for the design of magnetic-deflection systems for such sweep frequencies is not yet available.

The deflection sensitivity in any given cathode-ray tube is a function of the accelerating voltage. It is, therefore, possible to improve the deflection sensitivity by reduction of the acceleration voltage. This, however, likewise decreases both brilliance and resolution in the screen image since the spot size is increased at the lower accelerating voltage by the greater mutual repulsion between the electrons within the beam. The final acceleration voltage chosen must be a compromise between the normal screen brilliance, deflection sensitivity, and resolution.

The minimum resolution desired corresponded to

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approximately 25 lines per centimeter. It was found that for writing speeds ordinarily encountered on repetitive phenomena up to sweep frequencies of more than 100 kilocycles per second, this could be obtained with sufficient screen brilliance to permit use in an ordinary room without reduction of general room illumination.

In any practical cathode-ray tube, these conditions required a deflection amplifier in order to permit a reasonable picture size to be obtained at the lower signal levels at which it was found necessary to work so much of the time. The gain of these amplifiers was arbitrarily limited to that suitable for the majority of measurement work in order to avoid difficulties due to mechanical vibration. It was found that a practical design would provide a usable screen image for input signals between 25 and 50 millivolts on a root-mean-square, sine-wave basis.

Before the amplifier tubes could be chosen, the bandwidth had to be determined. At the time, the maximum available bandwidth for reasonably flat response in any commercial oscilloscope was approximately 1 megacycle but since television practice was being revised to employ a bandwidth of several times that figure, the need for greater bandwidth was apparent. Since test results had shown that phase displacement at the upper limit of the television frequency band was not too important, the choice of this limit was not made on this basis but rather on the basis of the number of harmonics required to delineate satisfactorily the highest frequency rectangular wave of primary importance in average use. From such considerations, it was decided that if the response were made flat throughout the band to better than 1 megacycle per second and any peak due to compensation were never allowed to exceed 110 per cent, while at 5 megacycles per second, the response was not less than 90 per cent, further extension of the high-frequency response would be of questionable value and not worth the cost to the general user. Such further extension should logically be accompanied by an increase in maximum sweep frequency.

The low-frequency response does not exercise the same degree of control over the choice of tubes since the required gain is not difficult to obtain in this range. The response in this region was chosen so as to permit less than 5 degrees of phase shift at 20 cycles per second because of a wide variety of measurement problems at the lower frequencies and the relatively greater importance of phase shift in this region. The actual test measurements show that the phase shift is less than this at 10 cycles per second. It may be of interest to note a fact employed by the author since 1925, that as the phase-shift characteristic of amplifiers continuously approaches zero over ever wider frequency ranges, the amplitude-response characteristic itself approaches greater independence of frequency throughout the same band. In fact, small departures from uniform response are more sensitively determined from the phase-shift characteristic.

With the gain, bandwidth, and output-voltage characteristics established, the tube complement was chosen as follows: push-pull 807 tubes for the power amplifier, push-pull 807 tubes for the final stage of the voltage amplifiers preceded by two stages of 6AC7/1852 tubes. Degeneration was used as far as necessary to improve stability and wave shape. The plate resistors of each amplifier stage were chosen to provide the maximum signal possible from that stage consistent with the principle of not requiring some other stage to over-compensate for serious deficiencies in any other stage.

The perfection of a deflection system is, of course, of little usefulness unless the sweep system provides a time axis capable of showing the fine structure of a complex high-frequency wave. The survey had shown a sufficient number of measurement problems requiring 100-kilocycle-per-second sweep systems to make this an essential specification. Experience showed that sweep frequencies of this order demanded vacuum tubes rather than gas tubes. The range chosen extended from 10 cycles per second to 100 kilocycles per second, while that obtained in actual practice is somewhat broader. This would correspond to approximately fifty complete cycles at the specified upper frequency limit of amplifier response. The linearity of the sweep circuit on a horizontal velocity basis was specified to be within 15 per cent departure from ideal as a maximum; i.e., no wavelength in a multiple-cycle screen image should depart more than 15 per cent from the average.

The general measurement problem fixed the amplifier deflection linearity at a maximum deviation of about $7\frac{1}{2}$ per cent. The amplifiers were, therefore, designed to provide much better linearity at the level corresponding to specified picture size. Curvature of the face of the cathode-ray tube makes utilization of the full screen impossible for measurements requiring greater linearity in a plane screen image.

In order to measure the applied signal voltages at various parts of a screen image, a deflection calibrating circuit was indicated. It was also specified that this should be accomplished without change of amplifier attenuator setting; i.e., over-all calibration should be provided from the input to the attenuator. The desire to make rapid relative measurements between various portions of a complex wave indicated the use of a transparent co-ordinate system in front of the face of the cathode-ray tube.

Among the other features also desired were complete absence of power circuit "pickup" in the screen image either in visible deflection or brilliancy, provision for selection of any section of a wave synchronized with the 60-cycle power-supply system or one of its lower harmonics, choice of retrace blanking by means of a switch on the front panel, and a rectangular wave of output voltage for test work.

In the mechanical design, it was desired to provide for ease of operation and observation from either a standing or sitting position, a rotatable co-ordinate scale system to simplify the measurement of phase shift by the ellipse method, simple and convenient

Fig. 1.—Wide-band oscilloscope.

service conditions, logical location of controls, and the design and placement of controls whose distinct character would make them readily recognizable without

Fig. 2.—Wide-band oscilloscope showing disposition of components.

the necessity of shifting the attention from the image on the screen.

Since it was at once evident that such an oscilloscope would be larger and heavier than could be conveniently carried, it was necessary to consider carefully the cabinet design from the standpoint of weight distribution for ease of handling, passage through doors, etc. Fig. 1 shows the final exterior design of the oscilloscope.

A number of these units have been in daily operation for at least four years without a serious complaint to date. The fields in which they have been used range from television, lamp manufacturing and the associated test problems, broadcast transmitter and receiver development, test, and installation, frequency-modulation transmitter and receiver development, test, and installation, power testing, to many of the present-day secret war activities. The input terminal design of the oscilloscope has lately been improved to reduce further the possibility of spurious response when used in the radio-frequency field of powerful transmitters in the many-megacycle range.

The arrangement of the various components including the horizontal sweep generator, some of the shielding, the BX power wiring, and general construction may be seen from the side view shown in Fig. 2.

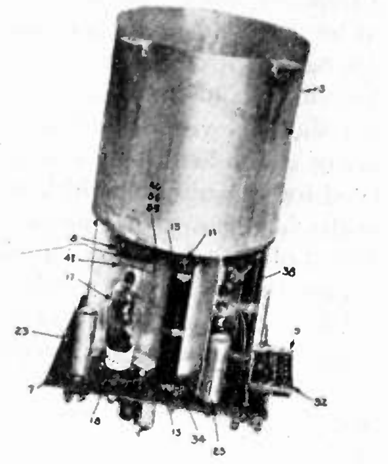


Fig. 3.—Power amplifier and cathode-ray-tube shield assembly.

A summary of performance specifications is given below:

Specified Picture Size: 100 by 100 millimeters with a specified deflection linearity of at least $92\frac{1}{2}$ per cent on symmetrical wave shapes.

Deflection Sensivitity: 140 millimeters peak to peak or more per volt, peak to peak.

Deflection Amplifier Response: Within ± 10 per cent from 10 cycles per second to 5 megacycles; less than 5 degrees of phase shift at 20 cycles per second.

Attenuator Accuracy: Ratio $\pm 2\frac{1}{2}$ per cent and constant within frequency band of amplifiers.

Maximum Input Signal: 600 volts root-mean-square.

Horizontal Sweep Linearity: Less than 15 per cent departure from uniform spot velocity.

Resolution: Approximately 25 lines per centimeter on lowest accelerating voltage.

Approximately 45 lines per centimeter on highest accelerating voltage.

Special Features: Terminals available on front panel for either double or single-ended external synchronizing voltage, also for rectangular output wave. Choice of

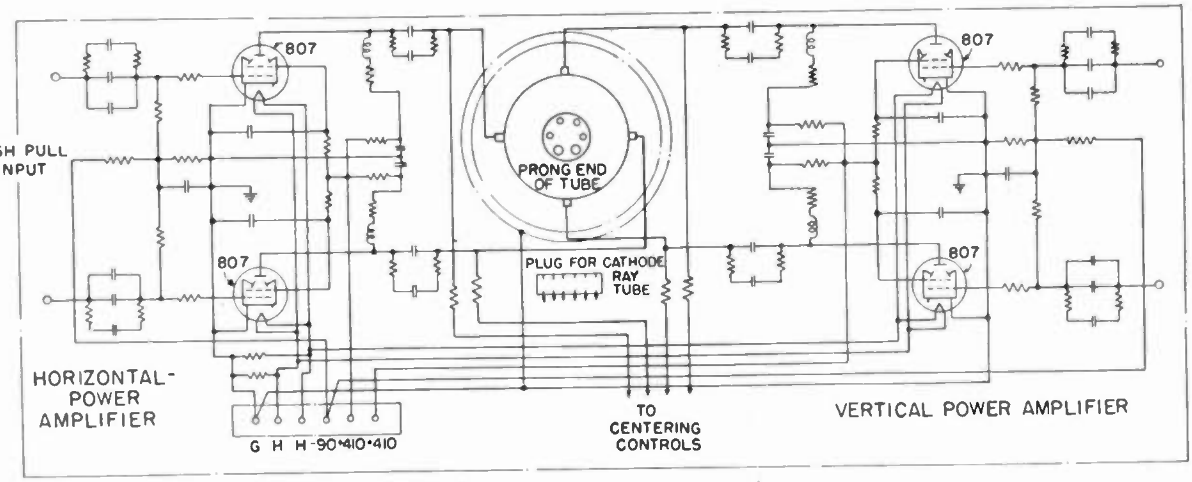


Fig. 4.—Schematic connection diagram of power amplifiers and cathode-ray tube.

retrace blanking by means of front-panel switch.

Power Supply: Commercial 60-cycle source; 105 to 125 volts.

Weight: 750 pounds.

Dimensions: Depth—29 inches. Width—30 inches. Height—51 inches.

Design

The two push-pull power amplifiers for the horizontal and vertical deflection systems have been located on a panel mounted around the socket end of the cathode-ray tube and secured to the heavy, annealed magnetic shield of this tube. The mechanical design will be apparent from Fig. 3. Each tube of the power amplifier has been individually shielded from every other. The panel can be withdrawn from the rear of the cabinet if necessary for servicing. As can be seen from the circuit diagram shown in Fig. 4, it is compensated for both low- and high-frequency response up to and including the cathode-ray tube itself.

The two voltage amplifiers have been mounted in metal shielding trays on either side of the cathode-ray tube. They had been designed to permit removal through the front panel for servicing. These units, together with a spare signal and calibration attenuator unit, are shown in Fig. 5. The change from single-ended input to push-pull output is accomplished at high

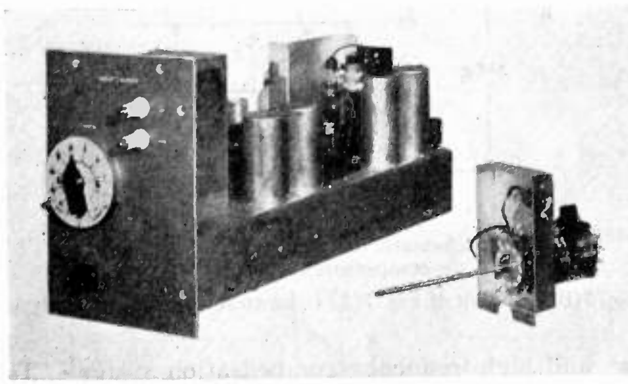


Fig. 5.—Horizontal voltage amplifier with calibration and signal attenuator assembly.

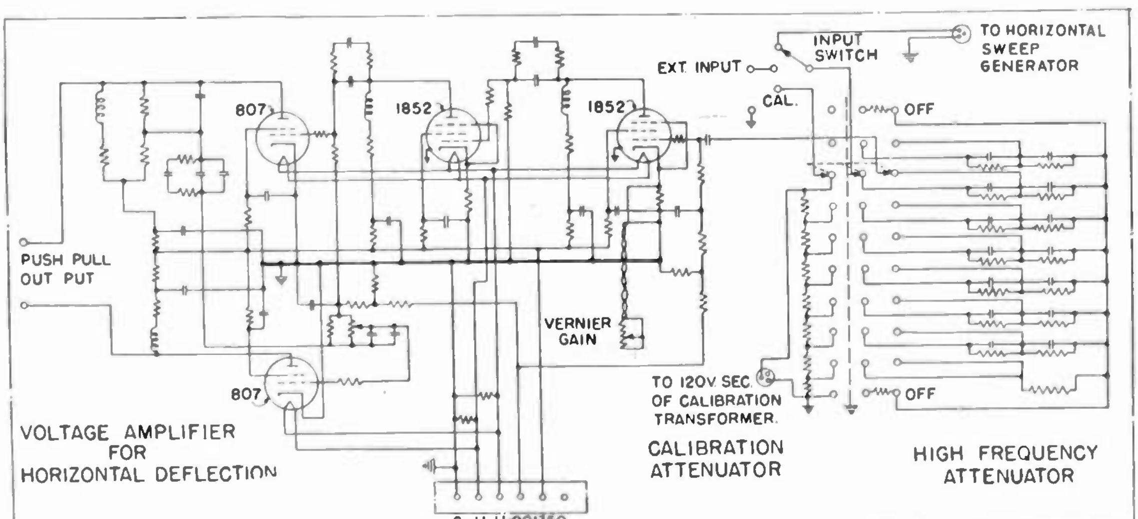


Fig. 6.—Schematic connection diagram of vertical voltage amplifier and attenuators.

serious trouble but further compensation is unwise. In fact, where sharp wave fronts are involved, any response in excess of flat response will cause a damped oscillatory condition in the neighborhood of the sharp wave front which may be undesirable in certain work. The deflection frequency response of the amplifier system is shown in Fig. 11 and the deflection linearity in Fig. 12.

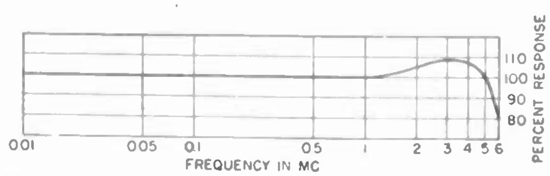


Fig. 11—Average over-all deflection response characteristics for wide-band oscilloscope.

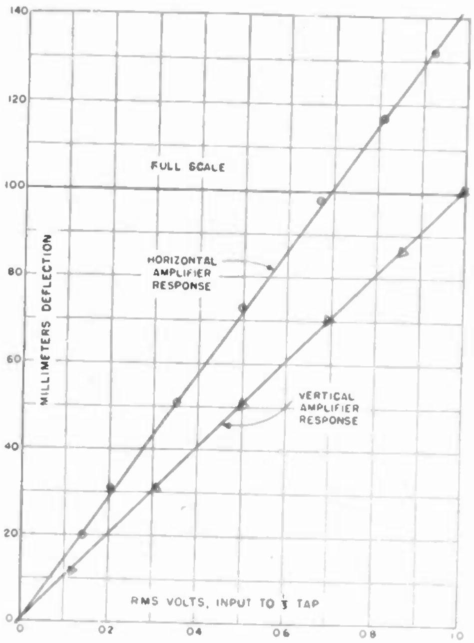


Fig. 12—Deflection linearity diagram for amplifiers of wide-band oscilloscope.

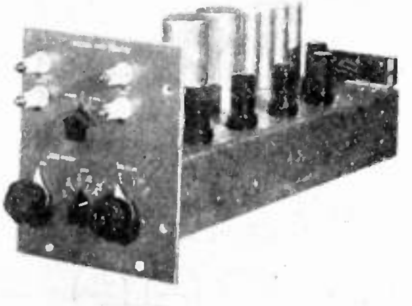


Fig. 13—Horizontal sweep generator for wide-band oscillosope.

Amplifier systems have been developed which extend the frequency range with the same percentage response tolerances to 10 megacycles per second; i.e., the maximum departure from uniform amplifier response does not exceed plus or minus 10 per cent within a high-frequency range which extends to 10 megacycles per second. In this case, the low-frequency response limit was raised to 30 cycles per second from the 10 cycles per second limit used in the standard design. The reason for this slight sacrifice in low-frequency response is to be found in the lack of space for large decoupling condensers when each power amplifier and its associated voltage amplifier were now combined in a single chassis which fitted the same space ordinarily occupied by the voltage amplifier alone. These units were combined to avoid the loss of possible amplification in the lower impedance circuit which would be required by the added burden in this extended frequency range due to the inevitable shunt capacitance of the necessary leads otherwise required to connect to separate chasses. Such amplifiers have not been incorporated in standard oscilloscopes because the majority of users do not seem

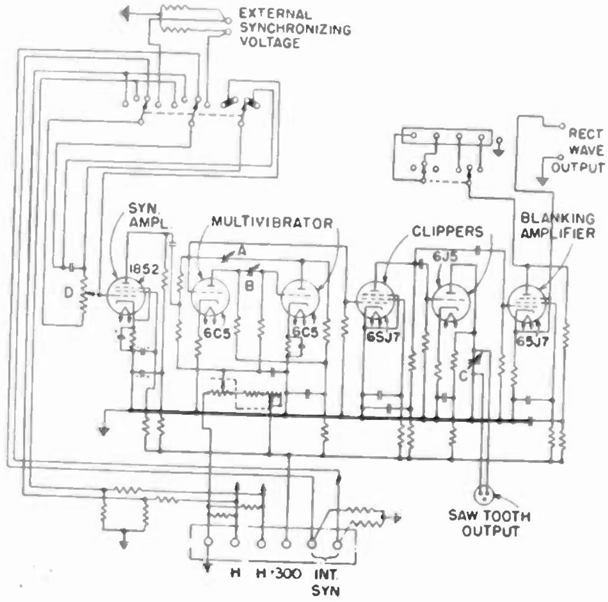
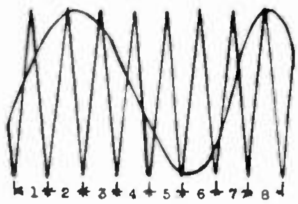


Fig. 14—Schematic connection diagram for horizontal sweep generator.

to require such response characteristics at this time. In fact, although available, they are not used in our own laboratory oscilloscopes at this time.

The horizontal sweep generator which is shown in Fig. 13 employs a conventional multivibrator using positive-bias variation to obtain the vernier adjustment of sweep frequency. The coarse selection of sweep frequency is made by a stepwise change in the coupling condensers of the multivibrators. The fact that steep wave fronts exist in most of the connections to the multivibrator was reason enough for incorporating a synchronizing tube or blocking stage ahead of the multivibrator. The input circuit to the synchronizing tube is interesting in that, by virtue of being double-ended, it provides a choice of polarity as well as a level of synchronizing voltage from one control. This is true of the internal as well as the external synchronizing-voltage position. Since, in the case of synchronization at power-supply frequency, the voltage for synchronizing purposes is obtained from the constant-voltage heater supply, it is unnecessary to provide a voltage control and hence, in this position, the synchronizing control is made to act as a phase-shifting device and thereby permit selection of the phase of the wave depicted on the screen. The circuit diagram is shown in Fig. 14.

The multivibrator is followed by two "clipping" stages to establish the maximum squareness of the voltage wave before conversion to a saw-tooth wave. This conversion takes place in the plate circuit of the second "clipper" stage which also acts as a discharge tube. The fact that a truly square wave has harmonics whose amplitudes are functions of $(1/n)$ where (n) is the order of the harmonic, while the saw-tooth wave has harmonics whose amplitudes are a function of



Sweep Frequency	Wavelength in Millimeters of Each Wave on Screen								Max. Per Cent Non-linearity	Retrace Cycles
kc	1	2	3	4	5	6	7	8		
100	10	10	11	11	11	11	11	10	9.1	1.4
20	11	12	12	12	12	12	12	12	8.3	1.4
20	10	10	10	10	10	10	10	10	0	1.3
4.5	10	10	10	10	10	10	10	10	0	1.2
4500	11	11	11	11	11.5	11.5	11	11	4.5	1.2
900	10	10	10	10	10	10	10	10	0	1.0
900	11	11	11	11	11	11	11	11	0	1.25
150	10	10	10	10	10	10	10	10	0	0.9
150	10	10	10	10	10	10	10	10	0	1.25
35	10	10	10	10	10	10	10	10	0	1.2
35	10	10	10	10	10	10	10	10	0	1.25
10	10.5	10.5	11.5	11.5	11.5	11	11	10	8.7	0.75

Fig. 15—Chart showing over-all deflection linearity of horizontal sweep circuits.

$(1/n^2)$, will show that the conversion of a rectangular wave to a saw-tooth is in reality occasioned by an amplifier whose frequency-response characteristic falls off inversely with frequency. Practically, this also shows why it is necessary to generate about 200 volts of rectangular-wave voltage in order to obtain the required fraction of a volt of saw-tooth sweep voltage having the desired perfection of wave shape throughout its entire duration. The table showed in Fig. 15 is a copy of a test record showing the measurements made to determine the linearity and retrace time for the sweep generator. Since the wavelengths were measured only to the nearest 0.5 millimeter, zeros will be found in the column labeled "maximum percentage nonlinearity."

In order to insure the isolation of the blanking circuit from the saw-tooth circuit, a separate blanking tube has been employed as was shown in Fig. 14. The control of the retrace blanking function, which is produced by a rectangular voltage wave of sweep frequency applied to the grid of the cathode-ray tube, is permitted by a front panel switch. A rectangular output wave, available at the terminals on the front panel, is derived from the cathode circuit of this blanking tube and, if desired, may be used to excite a test circuit whose out-

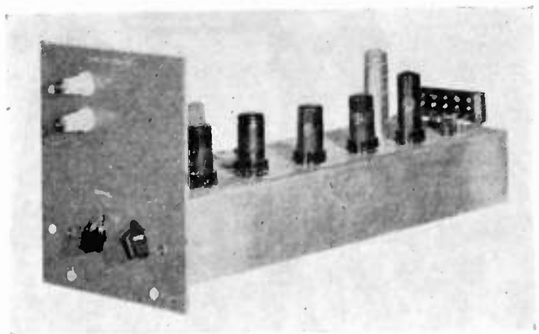


Fig. 16—Servo sweep generator for wide-band oscilloscope.

put is fed back into the vertical deflection system of the oscilloscope. After a little practice, one soon learns to interpret the rectangular diagram obtained.

It has been found feasible to extend the frequency range of the horizontal sweep generator to 1 megacycle per second but this development has not been incorporated in the standard oscilloscope because the majority of users do not seem to have need for such performance at this time.

An auxiliary, called the servo sweep generator, has been developed. This unit is shown in Fig. 16 while the circuit diagram is shown in Fig. 17. As its name is in-

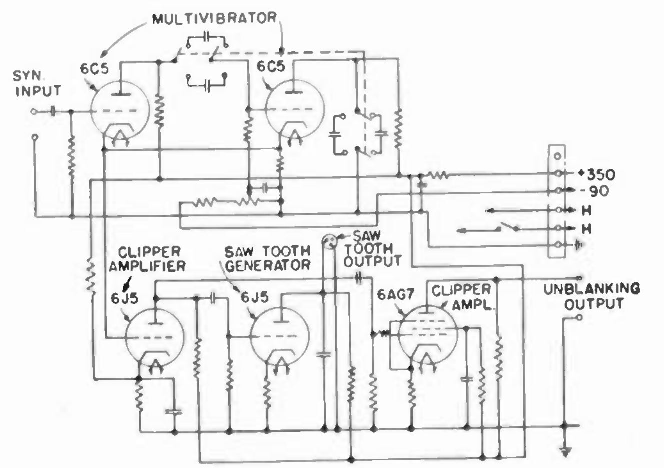


Fig. 17—Schematic connection diagram for servo sweep generator.

tended to imply, it is a sweep generator which provides one trace for each synchronizing pulse. The screen image appears only during the sweep process and has a constant sweep time regardless of repetition rate of the synchronizing pulse. Two sweep times have been provided, namely 5 and 25 microseconds, and the repetition rate of the synchronizing signal may be arbitrary up to 5000 cycles per second. Since the action is serious trouble but further compensation is unwise. In fact, where sharp wave fronts are involved, any response in excess of flat response will cause a damped oscillatory condition in the neighborhood of the sharp wave front which may be undesirable in certain work. The deflection frequency response of the amplifier system is shown in Fig. 11 and the deflection linearity in Fig. 12.

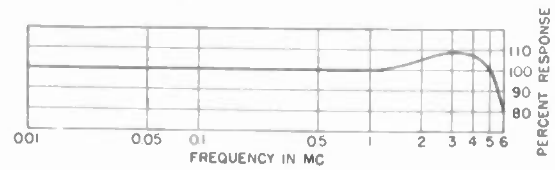


Fig. 11—Average over-all deflection response characteristics for wide-band oscilloscope.

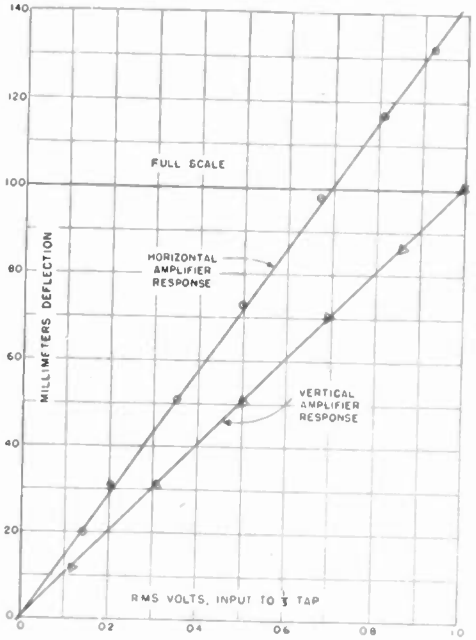
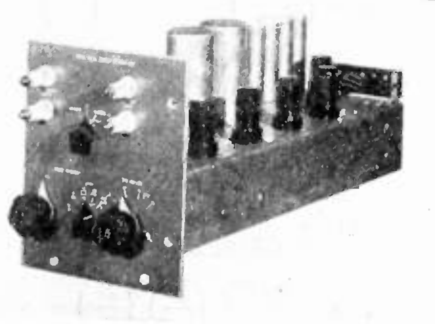


Fig. 12—Deflection linearity diagram for amplifiers of wide-band oscilloscope.



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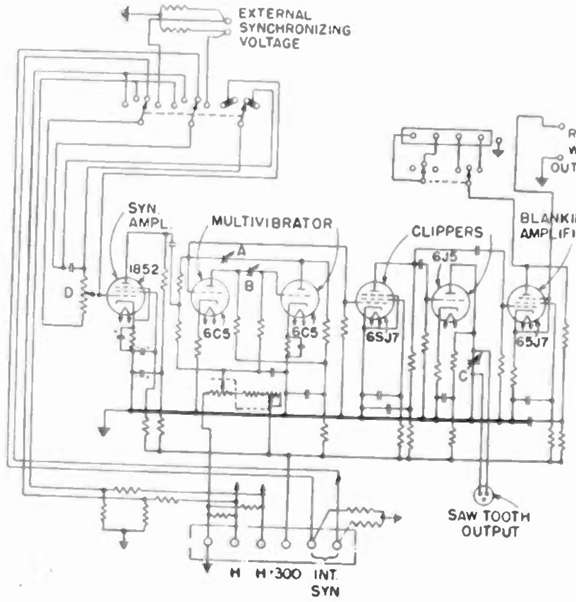

Fig. 14—Schematic connection diagram for horizontal sweep generator.

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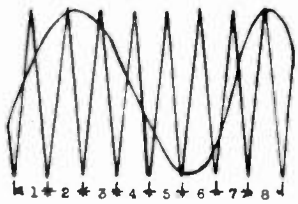

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20	11	12	12	12	12	12	12	12	8.3	1.4
20	10	10	10	10	10	10	10	10	0	1.3
4.5	10	10	10	10	10	10	10	10	0	1.2
4500	11	11	11	11	11.5	11.5	11	11	4.5	1.2
900	10	10	10	10	10	10	10	10	0	1.0
900	11	11	11	11	11	11	11	11	0	1.25
150	10	10	10	10	10	10	10	10	0	0.9
150	10	10	10	10	10	10	10	10	0	1.25
35	10	10	10	10	10	10	10	10	0	1.2
35	10	10	10	10	10	10	10	10	0	1.25
10	10.5	10.5	11.5	11.5	11.5	11	11	10	8.7	0.75

Fig. 15—Chart showing over-all deflection linearity of horizontal sweep circuits.

$(1/n^2)$, will show that the conversion of a rectangular wave to a saw-tooth is in reality occasioned by an amplifier whose frequency-response characteristic falls off inversely with frequency. Practically, this also shows why it is necessary to generate about 200 volts of rectangular-wave voltage in order to obtain the required fraction of a volt of saw-tooth sweep voltage having the desired perfection of wave shape throughout its entire duration. The table showed in Fig. 15 is a copy of a test record showing the measurements made to determine the linearity and retrace time for the sweep generator. Since the wavelengths were measured only to the nearest 0.5 millimeter, zeros will be found in the column labeled "maximum percentage nonlinearity."

In order to insure the isolation of the blanking circuit from the saw-tooth circuit, a separate blanking tube has been employed as was shown in Fig. 14. The control of the retrace blanking function, which is produced by a rectangular voltage wave of sweep frequency applied to the grid of the cathode-ray tube, is permitted by a front panel switch. A rectangular output wave, available at the terminals on the front panel, is derived from the cathode circuit of this blanking tube and, if desired, may be used to excite a test circuit whose out-

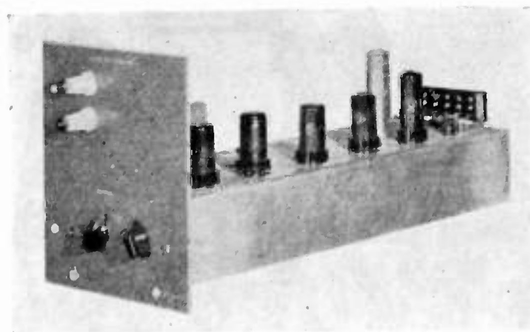

Fig. 16—Servo sweep generator for wide-band oscilloscope.

put is fed back into the vertical deflection system of the oscilloscope. After a little practice, one soon learns to interpret the rectangular diagram obtained.

It has been found feasible to extend the frequency range of the horizontal sweep generator to 1 megacycle per second but this development has not been incorporated in the standard oscilloscope because the majority of users do not seem to have need for such performance at this time.

An auxiliary, called the servo sweep generator, has been developed. This unit is shown in Fig. 16 while the circuit diagram is shown in Fig. 17. As its name is in-

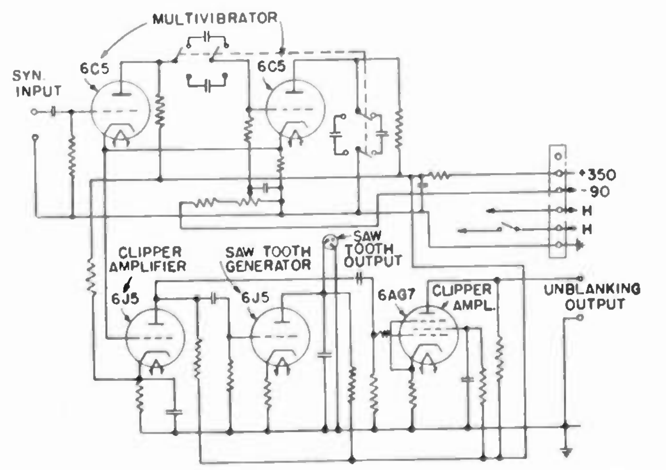

Fig. 17—Schematic connection diagram for servo sweep generator.

tended to imply, it is a sweep generator which provides one trace for each synchronizing pulse. The screen image appears only during the sweep process and has a constant sweep time regardless of repetition rate of the synchronizing pulse. Two sweep times have been provided, namely 5 and 25 microseconds, and the repetition rate of the synchronizing signal may be arbitrary up to 5000 cycles per second. Since the action is

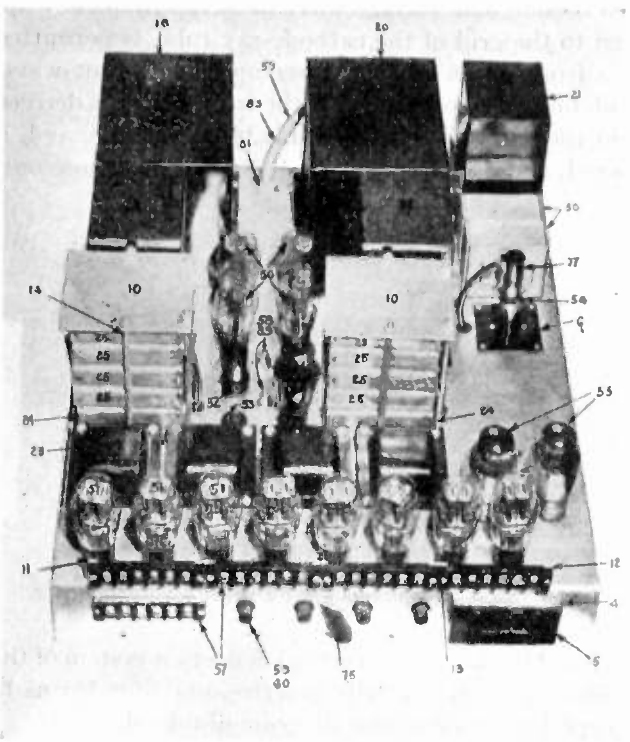

Fig. 18—Power-supply unit for wide-band oscilloscope.

obtained from a multivibrator biased beyond cutoff of plate current, the duration of the synchronizing pulse should not exceed that of the sweep pulse. The measurement of sweep linearity as well as sweep time and its accurate adjustment is accomplished with the aid of a special crystal controlled device called a Microlineometer.

The power-supply unit for the oscilloscope has been specially designed to have an exceptionally low external field. To aid further in this regard, the unit is mounted at the bottom of the cabinet with the transformers and chokes located at the maximum possible distance from the "gun" of the cathode-ray tube. The components operate at exceedingly low flux densities and are individually shielded. Additional compartment shielding is also provided to reduce further possible interference troubles. That these expedients are efficacious is evident from the fact that no trouble is experienced from this cause. The power-supply unit is shown in Fig. 18. It is connected to a metal outlet box by means of a twist lock connector and BX wiring.

Separate rectifiers have been employed for the power-amplifier and voltage-amplifier supply systems. In addition, each of these individual supplies are further divided by separate filters into separate supplies for the vertical and horizontal systems. The unit for the horizontal voltage amplifier together with the horizontal sweep generator and that for the vertical voltage amplifier are each individually electronically regulated. The ripple and internal impedance of these supplies are lower than is generally the case with such regulated supplies and a considerable overload capacitance has been provided. The supplies for the power-amplifier plate voltage are likewise divided but are not electronically regulated since this has been found unnecessary. Individual terminal boards have been provided to supply each component separately. The four individual low-voltage B circuits have been individually fused for their protection.

A separate high-voltage, direct-current supply is

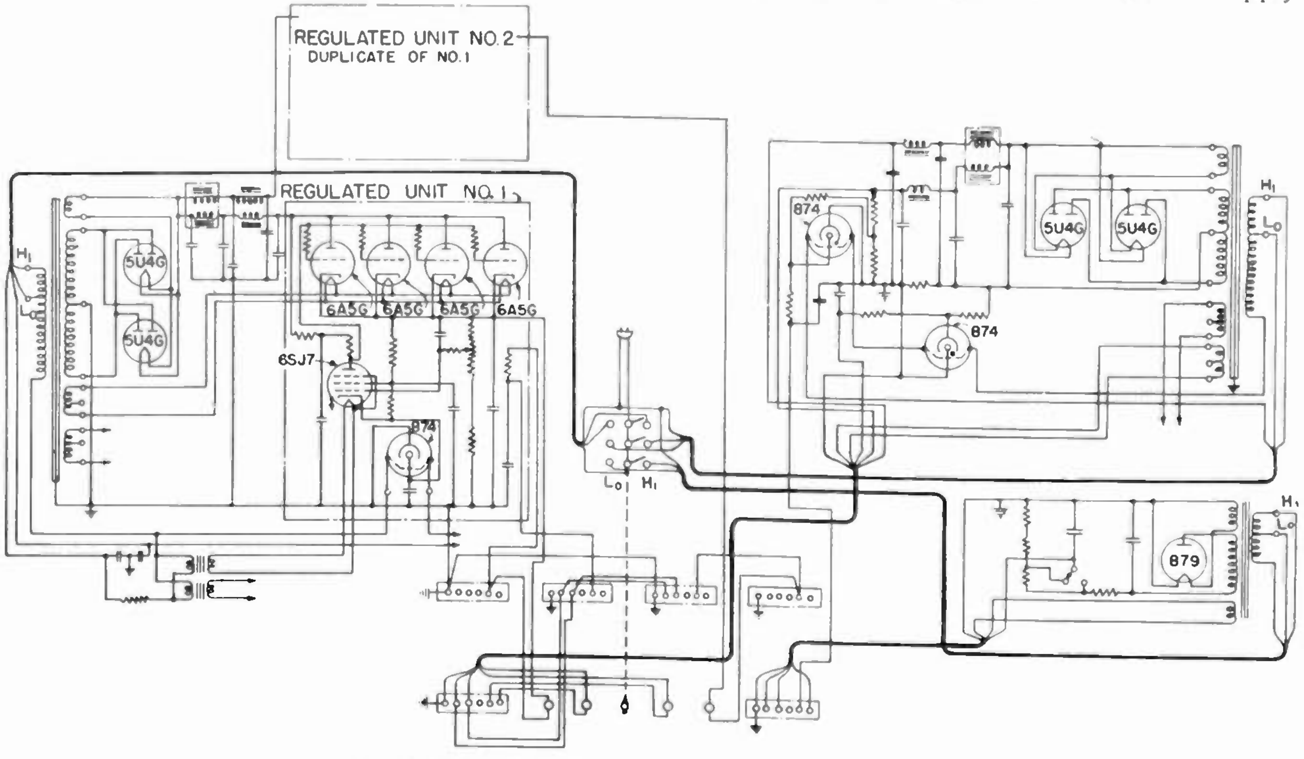

Fig. 19—Schematic connection diagram of power-supply unit.

employed for the cathode-ray tube. By shifting a connection in the filter circuit, it is possible approximately to double the accelerating voltage supplied to the cathode-ray tube. The result is a picture of greater brilliance and resolution but of reduced size. The maximum picture size having the specified linearity is now about 75 by 75 millimeters for symmetrical waves. The circuit diagram of the complete power-supply unit is shown in Fig. 19.

An interesting feature of the power-supply unit is that the line voltage can be read by means of the calibrating circuit voltmeter, and by means of a tap-changing switch, the various power transformers can be restored to their normal output voltage. This aids in keeping the heaters of all tubes at their correct operating temperature over a wide range of supply-line voltages.

Controls for the cathode-ray tube are provided to adjust vertical and horizontal centering, focus, and brightness. These are located on a panel behind the front central panel and are connected by long insulating shafts to the control knobs. These control knobs are conveniently grouped on the front panel around the calibration voltmeter and control and beneath the cathode-ray tube itself. It should be mentioned that it is not necessary to readjust these controls when the accelerating voltage is increased as described above.

The face of the cathode-ray tube is covered by a safety glass for the protection of both the operator and tube, and the bezel of the cathode-ray tube is provided with a transparent co-ordinate system which is rotatable. It has been designed especially to aid in the rapid determination of relative amplitudes for various wave pictures as well as phase shift by the ellipse diagram method. A light shield is provided for protection against general overhead room illumination.

The front of the oscilloscope cabinet has been found a convenient place to mount the input power plug, telltale pilot lamp, and magnetic-overload circuit-breaking switch. The cabinet has been provided with hinged doors on either side and at the back. Since access through the side doors is necessary only for the removal of a chassis, or the replacement of a tube, they are normally closed by means of screws. The hinged rear door, which is normally held closed by spring latches and key-operated locks, has been interlocked with the power-supply circuit to permit access without danger to the operator.

The use of cathode-ray oscilloscopes is now so general that it is unnecessary to discuss this phase of the subject. It is a fact, however, that new fields of application are being continually found for this device.

Acknowledgment

This opportunity is taken to acknowledge the valued assistance of Messrs. J. L. Theisen and E. F. Travis during both the development and production work.

Use of Subcarrier Frequency Modulation in Communication Systems*

WARREN H. BLISS†, ASSOCIATE, I.R.E.

Summary—When subcarrier frequency modulation having a frequency range of 1600 to 2000 cycles was used for transoceanic facsimile transmission, pictures were obtained with finer detail and better half-tone quality than those transmitted by previous systems; the speed of transmission could also be increased. An extension of the system to a two-way multiple-channel radio-relay circuit providing teletype service between New York and Philadelphia gave improved stability of operation when variations in signal strength occurred.

General

As THE art of frequency modulation has developed, it has naturally produced many by-products and has found applications and uses not at first apparent to radio engineers. One of these is in the field of point-to-point communication where an application known as subcarrier frequency modulation has been investigated and found to have important advantages. When a subcarrier wave is frequency-modulated by signal material and is then used to modulate a primary carrier, the system is referred to as subcarrier frequency modulation.

Two applications of subcarrier frequency modulation which are highly successful or show definite promise are (a) the transmission of facsimile material over long radio circuits and (b) the transmission of telegraph characters (printer-type signals in particular) over short-wave radio-relay circuits. The former has been used by R.C.A. Communications, Inc., for over two years with results far surpassing those of any other method.[1] Experimental tests recently conducted with the latter prove there is a definite advantage in stabilizing teletype signal levels when the different tones from multiple-channel operation are combined

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[1] R. E. Mathes and J. N. Whitaker, "Radio facsimile by subcarrier frequency modulation," *RCA Rev.*, vol. 4, pp. 131–154; October, 1939.

together to frequency modulate a common subcarrier wave. In both of these cases the major advantage is due to the great reduction in signal-amplitude variation rather than to any increase in signal-to-noise ratio resulting from the frequency-modulation method.

Radiophoto Practice

All facsimile systems use a scanning device at the sending terminal for converting graphic material into a variable amplitude current or voltage. In the subcarrier-frequency-modulation process this variable

Fig. 1—One of the first test radiophotos sent from Moscow to New York, via R.C.A. Communications, Inc.

quantity representing the point-by-point density of the copy is used to swing or shift a subcarrier tone over a definite predetermined range. This constant-amplitude, frequency-modulated tone is then sent out by means of a standard amplitude-modulation radiotelephone transmitter. At the distant terminal all amplitude variations resulting from fading and multipath effects are removed by means of a limiter, and a simple frequency-discriminator circuit converts the received signals back to the conventional varying amplitude wave. The copy is then recorded photographically, or by other means.

Since May, 1939, when subcarrier frequency modulation first went into public service between London and New York, additional terminals of this type have been put into commercial use. Berlin did and Buenos Aires does provide this service now and just recently transmitting facilities have been put into operation in Moscow. Fig. 1 shows one of the first photographs sent by radio from Moscow over a 5000-mile circuit.

The two outstanding advantages of subcarrier frequency modulation over dot-pattern and other systems are better quality of received copy and increased speed. Much finer picture detail can be reproduced and contrast in the darker half tones is better. Eight- and ten-point type are now transmitted successfully whereas fourteen-point type formerly was the lower limit in size. Streaks and irregularities due to noise are not so noticeable in the radiophotos. Because of this better quality and freedom from a screen, the radio pictures can be handled the same as conventional photographs by publishers. An eight- by ten-inch picture can be transmitted in twenty minutes under radio-circuit conditions which three years ago would require a period of one hour. Improved operating technique and advancements in radio communication indicate that the speed may be increased still more.

The primary reason for this great improvement is due to the fact that subcarrier frequency modulation offers the equivalent of a linear amplitude-variation system which functions independently of signal-level fluctuations caused by fading and multipath phenomena which always occur on long-distance radio circuits. Since the picture intelligence is conveyed by frequency modulation of the subcarrier tone, variations in amplitude can be completely removed by means of a limiting device. Since changes in amplitude as great as 70 decibels can be eliminated, signal fading can have very little influence on the reproduced copy. Only rarely does the signal drop below the usable level.

Although frequency modulation is capable of giving a definite gain in signal-to-noise ratio, this advantage increases with increasing deviation ratio. In the present picture-transmission system this advantage is not fully used because a comparatively low deviation is utilized; actually, the bandwidth is no greater than that for amplitude modulation. Some time ago a series of tests was made in one of the research laboratories of R.C.A. Communications, Inc., to determine the bandwidth requirement for picture transmission by subcarrier frequency modulation. Various depths of modulation and bandwidths were tried and oscillograms and transmitted copy were compared and closely examined. The comparative importance of various sideband components was also investigated.

The results proved that if the index of modulation m, which is the ratio of frequency deviation of the carrier to the modulating frequency, did not exceed 0.47 for the highest important frequency f_m from the picture material, then the total bandwidth needed is equal to $2f_m$. This range (f_c-f_m) to (f_c+f_m), where f_c is the carrier frequency, is the same as that for amplitude modulation and reproduces good copy. The highest important frequency f_m is determined by the speed of scanning and by the "definition" of the graphic material to be reproduced according to the I.R.E. standard method. With this bandwidth of $2f_m$ only the first-order sidebands of f_m are transmitted, the higher-order sidebands being sufficiently unimportant to be neglected.

Under this condition of bandwidth and $M=0.47$ for the top important frequency, it can be shown that for any other lower signal frequency no sideband components having amplitudes greater than 10 per cent of that of the unmodulated carrier are lost although the frequency swing or deviation is held constant. For example, with $f_c = 1800$ cycles per second, $f_m = 600$ cycles per second, and $m = 0.47$, the first-order sidebands have amplitudes of 15 per cent of that of the unmodulated carrier and the second-order sidebands have amplitudes of 2 per cent. The former will lie just within the $2f_m$ band width while the latter will be lost. For a lower signal frequency of 300 cycles per second, m would be 0.94 for the same deviation and the first-order sidebands having amplitudes of 42 per cent lie well within the transmission band. The second-order sidebands in this case have amplitudes of 10 per cent but they are also inside the $2f_m$ band. Hence no components having amplitudes greater than 10 per cent are lost. Picture quality is just as good as that obtained with subcarrier amplitude modulation employing the same bandwidth.

Although a subcarrier frequency range 1600 to 2000 cycles has been used for most of the transoceanic picture traffic, other ranges would be equally satisfactory. It is good practice to have the mid- or carrier-frequency several times the highest signal frequency but good results can be obtained with a wide variety of values of frequency swing. One communications system uses the frequency deviation range 2600 to 3400 cycles. In a higher-speed, experimental, wire-line test the range 7.5 to 10.5 kilocycles gave satisfactory results. From the standpoint of successful radio transmission the phenomena of selective fading would indicate that a comparatively low value of the subcarrier would be desirable since this would keep the radio-frequency bandwidth relatively narrow. However, there was no noticeable difference between results with $f_c = 1800$ cycles per second and $f_c = 3000$ cycles per second.

Another advantage of subcarrier frequency modulation over the old constant-frequency, variable-dot-length method is a more simplified, less critical operating technique. When a picture is to be sent, a band on the transmitting drum containing alternate black and white sections is first scanned. The scanner and modulator are then adjusted to produce 2000 cycles and 1600 cycles for black and white, respectively. At the radio transmitter the percentage amplitude modulation is set at the desired value and this remains constant since there is no amplitude variation in the subcarrier frequency-modulated picture wave. At the receiving terminal the recording galvanometer light valve or a similar device is adjusted to the proper settings for the incoming black-and-white controlled signals. Several pictures may then be transmitted without readjustment at either terminal.

The value and success of the present system of transmitting pictures over long distances by radio is indicated by the large number of such pictures which appear in our newspapers. The quality of the transmitted photos is frequently so good that they are indistinguishable from the regular news pictures.

A means for partially overcoming some of the effects of selective fading has been outlined.[1] This consisted in using the second harmonic of the subcarrier frequency which was developed in the receivers when the radio-frequency carrier completely dropped out. The harmonic wave which also contains the picture modulation was switched into use whenever the level of the normal frequencies fell below the usable point.

Telegraph Channeling with Subcarrier Frequency Modulation

For several years there has been in operation between New York and Philadelphia a two-way, multiple-channel, telegraph radio-relay circuit consisting of three links.[2] There are eight narrow-band, voice-frequency, teletype channels and also some wide-band, higher-frequency channels for facsimile. Each teletype channel is 100 cycles wide and the channels are spaced 170 cycles apart beginning with 425 cycles for the lowest frequency. The signal of each channel consists of its frequency keyed on and off according to the message material and the outputs of all eight channels are combined directly together through filters. The composite signal is then used to amplitude modulate the radio-frequency carrier, but in order to prevent overmodulation from peaks due to random combination the signal levels in the individual channels must be kept comparatively low.

Although this system has been successful in operation, difficulty is experienced from time to time because in this system of operation the teletype printers in narrow-band channels are sensitive to variations in incoming signal level. This requires a very careful adjustment of levels at both terminals and at the relay points. If for any reason there is a signal-level variation of a few decibels, then printer failures occur.

Because of the great success of subcarrier frequency modulation in picture transmission, it seemed reasonable to expect an improvement if it were applied to subcarrier telegraph operation. Tests were accordingly made and a marked improvement was noted. Fig. 2 is a block diagram showing how a subcarrier frequency-modulation unit was used between the group of outgoing channels and the radio-frequency transmitter and a combined limiter and frequency discriminator were included between the receiver and incoming channeling filters. By direct comparison it was found that where a 4-decibel drop in equivalent signal level caused failure in the amplitude-modulation method, a 22-decibel drop was required to cause failure when subcarrier frequency modulation was used. This shows an appreciable advantage.

In both cases the signal-to-noise ratio was of secondary importance. The advantageous gain in allowable

[2] J. Ernest Smith, Fred H. Kroger, and R. W. George, "Practical application of an ultra-high-frequency radio-relay circuit," Proc. I.R.E., vol. 26, pp. 1311–1327; November, 1938.

signal-level fluctuation before printer dropouts occur is due primarily to elimination of most of the effect of this fluctuation before it gets to the actual printer machine. As in the case of pictures, amplitude variations encountered in transmission are reduced to nil by the limiter device included at the receiving terminal.

Another suggested means for overcoming the variation in signal level in the present telegraph channeling system is the use of automatic volume control in each channel. However, the use of subcarrier frequency modulation is preferable because of its inherent nature of conveying intelligence irrespective of amplitude fluctuations.

Still another means of accomplishing the same desired result is to use subcarrier frequency modulation in each of the individual channels. This, in reality, would be a case of frequency-shift keying. Although it provides for satisfactory operation, frequency-shift keyers and limiters and discriminators are needed for each channel.

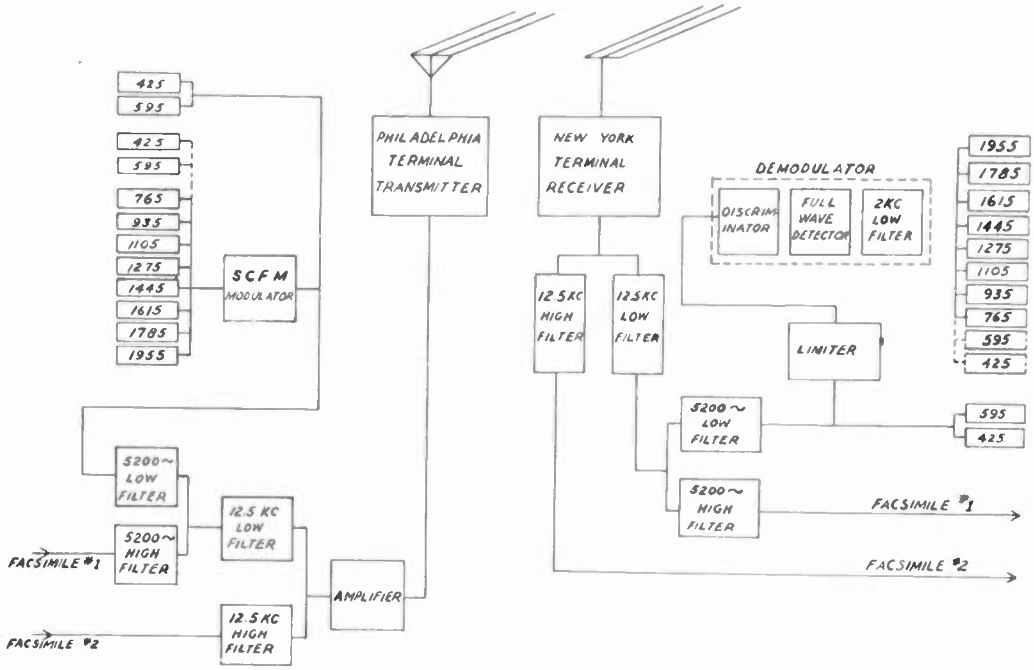

Fig. 2—Diagram of test circuit showing subcarrier frequency modulation applied to multichannel teletype operation.

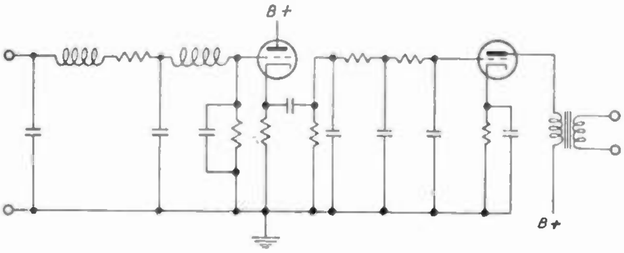

Fig. 3—Wide-range frequency discriminator.

If either amplitude or frequency modulation of a subcarrier is used, the matter of a low ratio of subcarrier frequency to maximum signal frequency is of importance. If this subcarrier is to modulate a higher-frequency primary carrier it is desirable to use as low a value of subcarrier as possible so as to conserve bandwidth or to use a given bandwidth most efficiently. Results for the subcarrier amplitude-modulation case

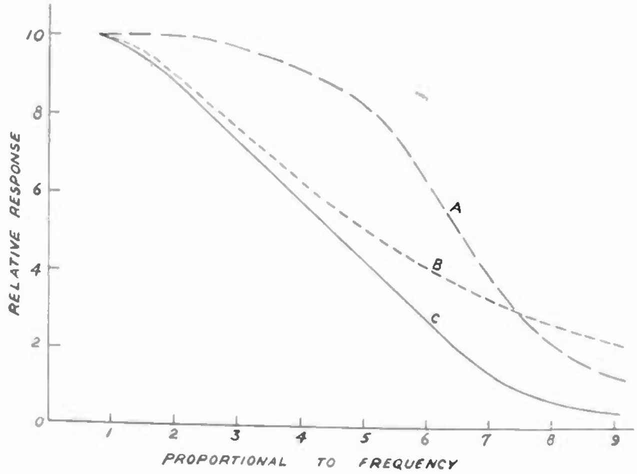

Fig. 4—Frequency-response curves of wide-range frequency discriminator.

with a facsimile type of signal have been published.[3] With too low a ratio of carrier to signal, poor results were obtained but a value of 2.5 to 1 was considered "entirely adequate for commercial service." However,

[3] R. E. Mathes and J. E. Smith, "Propagation requirements for facsimile," in book "Radio Facsimile," R.C.A. Institutes Technical Press, New York, N. Y., 1938.

in the teletype channeling tests with subcarrier frequency modulation, satisfactory results were obtained with a maximum signal frequency of 2 kilocycles modulating a 3-kilocycle carrier, a ratio of 1.5 to 1. This was with a low index of modulation and is unusual because the lower first-order sideband has a frequency value less than the modulating frequency.

The reader who is familiar with various types of frequency discriminators will probably be interested in the unit used in the above mentioned tests. It was made up with two parts as illustrated in Fig. 3. The first part consisted of two sections of constant-K, low-pass filter, with coils having a comparatively low value of Q. The frequency response is given by curve A of Fig. 4 and shows discrimination such as that used in the radiophoto terminal equipment. The range of linearity of this filter unit was not as great as desired for another application so two sections of a resistance-capacitance-type low-pass filter, having the characteristic of curve B, were added. Isolating vacuum tubes are used for coupling and impedance matching. Curve C shows the over-all frequency response of the complete unit which has an exceedingly wide range of linearity.

Such a characteristic is necessary when the ratio of subcarrier to signal frequency is low, but in some of the applications of subcarrier frequency modulation, this mode of operation is highly advantageous in the conservation of frequency bandwidth. An actual saving is made because if a high value of subcarrier is used in the modulation of a higher primary carrier then there will be a "waste" space of frequencies between the primary carrier and its nearest sidebands.

Conclusions

In summing up the results and observations with subcarrier frequency modulation, the outstanding advantage of this system in point-to-point communications is in overcoming the ill effects of unwanted amplitude variations which are always present to some extent in all transmission circuits. In any system in which the intelligence is conveyed by amplitude variation, any noise wave introduced into the path of transmission will add directly to the signal and produce distortion, masking, or other undesirable effects. In frequency modulation, all amplitude irregularities are removed by limiting or clipping. The inherent gain in signal-to-noise ratio obtained with this type of modulation is also utilized to an extent although the index of modulation is usually less than unity. In general the same advantages of frequency modulation are obtained in subcarrier frequency modulation as in the more common application of straight frequency modulation of a radio-frequency carrier.

It has been planned to present in the Proceedings of the I.R.E. instructional material of timely interest. This procedure was instituted some time ago, and here continues by the publication, in successive issues of the Proceedings, of a series of co-ordinated parts, together entitled "Some Aspects of Radio Reception at Ultra-High Frequencies" by Messrs. E. W. Herold and L. Malter. Part I of the five parts is here presented. Forthcoming issues of the Proceedings will contain succeeding Parts of this series. Each Part will be preceded by its own related summary. *The Editor*

Some Aspects of Radio Reception at Ultra-High Frequency*

E. W. HEROLD†, MEMBER, I.R.E., AND L. MALTER‡, ASSOCIATE, I.R.E.

Part I. The Antenna and the Receiver Input Circuits

E. W. Herold

Summary—This paper is in five parts, of which this is the first, and includes material prepared by the authors for a lecture course given during 1941–1942.

At ultra-high frequencies the fluctuation noise of tubes and circuits in the receiver is sufficiently greater than antenna noise and other forms of interference so as to limit the reception of weak signals. Signal-to-noise ratio is often one of the chief problems in reception at ultra-high frequencies. The bandwidth of the receiver is also of great importance, both for the determination of the maximum speed at which intelligence can be received and for the determination of the total noise which will be encountered. Circuit and noise bandwidth are not always the same and are distinctly separated in the analyses. Finally, selectivity is a third important aspect of ultra-high-frequency reception.

It is shown that the receiving antenna "captures" an amount of the transmitted power which, at a given wavelength, depends only on the directivity. Thus, receiving-antenna design is chiefly concerned with directivity, which determines the maximum signal-to-noise ratio and with bandwidth, or Q, which determines the useful frequency range. The Q of the half-wave dipole is determined by its surge impedance which in turn depends on the ratio of

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diameter to length of the conductors. The antenna Q is low even for small diameters and is lowered even further by the receiver load. Coupling the antenna to the receiver requires low-loss transmission lines and proper impedance matching. The amount of reflection at line-coupling elements or at insulating beads can easily be calculated and formulas are given for cases in which the nonuniform sections are short compared with the wavelength. The losses due to such reflections rise with frequency and may be appreciable even for rather small beads, if isolated.

The selectivity of superheterodyne receivers is largely obtained in the intermediate-frequency stages. However, the receiver input circuits, prior to the mixer stage, determine the image response. It is found that the image-to-signal ratio depends on the ratio of the input-circuit bandwidth to intermediate frequency. Thus, high intermediate frequencies are desirable. The design of resonant-line input circuits between the antenna and the tube to obtain any desired bandwidths can be carried out by use of lumped-circuit equivalents. This procedure is simplified since distributed losses in the line circuits can often be neglected in comparison with the lumped loss introduced by connection of a tube. If a bandwidth is too narrow, the Q can be lowered by coupling the antenna tighter. If the bandwidth is too wide, the Q can be raised by coupling the tube more loosely to the circuit.

An Appendix gives useful data for transmission-line circuits including their equivalents in the form of lumped L and C circuits.

I. Introduction

During the winter of 1941–1942, a lecture course entitled "Ultra-High-Frequency Practice" was conducted in the northern New Jersey area by the Newark College of Engineering and Rutgers University under the auspices of the Engineering, Science and Management War Training Program. Most of the following material was prepared for the portion of the course dealing with reception and was presented to seven sections of the course by the authors. The material was also used by E. W. Herold in a portion of an ESMWT course given at the Polytechnic Institute of Brooklyn during the summer of 1942.

The work is intended to be useful to those who, although familiar with radio technique at lower frequencies, have only recently encountered the problems peculiar to the very high frequencies. It is also intended that this series survey the field in a broad way so as to acquaint the specialist in other ultra-high-frequency fields with the basic problems of reception. Many of the aspects covered have been previously published in scattered papers and frequent reference is made to the more important of these. They must be consulted for detailed information since this presentation is necessarily condensed.

The subject has been divided into five parts, chiefly corresponding to individual lectures of the aforementioned course, and each part is essentially complete in itself. These five parts are:

Part I. The Antenna and the Receiver Input Circuits—E. W. Herold

Part II. Admittances and Fluctuation Noise of Tubes and Circuits—L. Malter

Part III. The Signal-to-Noise Ratio of Radio Receivers—E. W. Herold

Part IV. General Superheterodyne Considerations—L. Malter

Part V. Frequency Mixing in Diodes—E. W. Herold

II. The Chief Problems in Ultra-High-Frequency Reception

1. *Signal-to-Noise Ratio*

One of the chief differences between the ultra-high-frequency problem and reception at lower frequencies lies in the fact that there is frequently only one receiver for each transmitter. At lower frequencies, one of the major fields is broadcasting whereas at ultra-high frequencies (above 100 megacycles) it is not usual to use a single transmitter covering a very large number of receivers. Instead, we often have what is called point-to-point communication. This is a factor which greatly influences receiver and receiving-antenna design. At the antenna, it leads to directive arrays; these are all the more useful at ultra-high frequency because of the high directivity which may be obtained in a small space. In the receiver it means better design and more justification for good performance irrespective of cost. For example, suppose a distance d must be covered by a transmitter of power P_t. The received power will vary as some rapidly diminishing function of the distance $f(d)$, and is

$$P_r = kf(d)P_t.$$

However, in any receiving system, we have a certain amount of unavoidable interference which we may broadly class as "noise":

1. The noise received by the antenna along with the signal.

2. The noise generated in a receiver during the process of reception. We may call the total noise power at the receiver P_N and, for the signal to be heard readily, $P_r > P_N$. Thus the transmitter power

$$P_t > \frac{P_N}{kf(d)}$$

and it is important that the total noise power be small if d is large since then $f(d)$ is small. Now at frequencies below 20 megacycles, the receiver noise may be made small compared with that received by an efficient antenna. Jansky[1] found that the antenna noise was always at least four times that of his receiver at 20 megacycles. Antenna noise at lower frequencies is chiefly a result of local atmospherics but at 20 megacycles much of it appears to come from interstellar radiation. From 20 megacycles up to about 70 megacycles the noise received by the antenna diminishes although ignition noise and man-made interferences are present. Above 70 megacycles the antenna noise level becomes so small compared with receiver noise that there is no

[1] K. G. Jansky, "Minimum noise levels obtained on short-wave radio receiving systems," Proc. I.R.E., vol. 25, pp. 1517–1530; December, 1937.

clear evidence as to its magnitude. At ultra-high frequencies, therefore, receiver noise is the chief component of P_N and one is faced with quite a different problem from that encountered at lower frequencies.

In a good all-wave radio receiver, with a tuned radio-frequency stage properly coupled to an antenna, we would not, as a rule, find any contribution to P_N by receiver noise. The same radio-frequency stage, which might use an ordinary broadcast receiving tube such as the type 6SK7, could be made operative at 100 megacycles. At this frequency, however, laboratory measurements would show that the receiver contributed something like 31 times as much noise as the dummy antenna and P_N would be about 32 times the antenna noise, a high value. By choice of the proper ultra-high-frequency tube, a receiver could be designed to reduce P_N to around four times the dummy antenna noise. We could then reduce our transmitter power to $\frac{1}{8}$, a saving of $\frac{7}{8}$ of its original power. This simple example shows that a reduction in receiver noise, which is the same as an improved signal-to-noise ratio, is worth many dollars in terms of transmitter power.

As the frequency gets higher, the antenna noise level remains small and the receiving tubes are less efficient so that the contribution of the receiver to P_N is larger and larger in comparison with our ultimate noise level. Every improvement in the receiver, therefore, cuts the transmitter power.

These points are emphasized because the most important fundamental problem of receiver design at ultra-high frequency is the signal-to-noise ratio.

2. *Bandwidth*

The next aspect of ultra-high-frequency receiver design which must be emphasized is bandwidth. As we know, a steady, unvarying direct current can transmit no intelligence. If this current varies slowly, it transmits information slowly, i.e., at a low rate. The more rapid the variations, the more intelligence can be transmitted in a given time. However, radio transmissions operate by modulating the carrier by the intelligence, so the carrier must be transmitted. That is, the carrier frequency must be at least as high as the highest modulating frequency. If many channels are required, a very high carrier frequency must be used. In very many ultra-high-frequency applications, a wide band is necessary; bands of 1 or 2 megacycles are common and for television purposes bands of 10 megacycles or more are desirable. It may be pointed out that, even if one wishes to send only voice transmission on an ultra-high-frequency carrier one must consider the frequency stability of the oscillator. Thus the receiver band must be considerably wider than 10 kilocycles to take care of possible frequency drift.

Now, to return to bandwidth, as the frequency becomes higher, circuit Q's become higher but often tubes load down the circuits to some extent and the net Q is not far from that at low frequencies. Thus, since the bandwidth of a circuit tuned to a frequency f is

$$\Delta f' \text{ (3 decibels down)} = \frac{f}{Q}$$

it is found that a loaded circuit of over-all $Q=100$ still has considerable bandwidth. In general, the circuit Q's in the ultra-high-frequency field are very high and the low over-all Q is a result of connecting a tube with

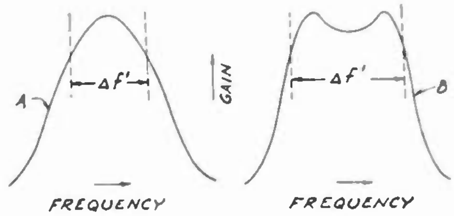


Fig. 1—Frequency responses of typical amplifiers or receivers. The circuit bandwidth $\Delta f'$ is defined roughly by points 3 decibels down from the maxima.

its associated lead and electronic losses. However, by tapping down on the circuit, it is possible to drive a tube and yet keep the over-all Q high, with a slight sacrifice in signal.

While on the subject of bandwidth, there are two quantities, between which we must distinguish, both of which are called bandwidth. If we have a circuit, or an amplifier, or a receiver, it will have an output-versus-frequency curve such as A or B of Fig 1. If

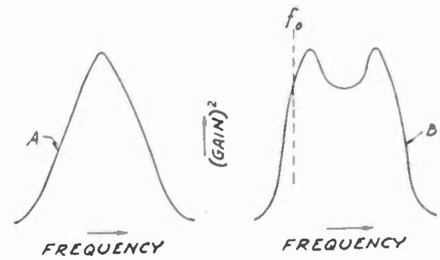


Fig. 2—Square of the frequency response of the typical amplifiers or receivers of Fig. 1. Frequency f_0 denotes position of a possible signal for signal-to-noise comparison.

all the frequencies in a transmitted signal are to be received with a specified fidelity we design our receiver response to be reasonably smooth and then define the bandwidth arbitrarily in terms of this fidelity. In Fig. 1, for example, a bandwidth $\Delta f'$ is indicated on both curves and is defined by the 3-decibel down points. For noise purposes, however, this bandwidth has no direct significance. Suppose we consider a noise which is distributed uniformly over all frequencies so that the noise output spectrum has the same amplitude shape as the selectivity curves of Fig. 1. The measurement of the noise must be made in terms of the over-all effect. Since the noise may be considered as similar to a large number of different frequencies (closely spaced) we must add them in root-mean-square fashion or better yet, add their power. The noise power is proportional to the *square* of the curves of Fig. 1; such squared curves are shown in Fig. 2. The total area under the curves in Fig. 2 gives the total noise power output

$$\text{noise power} = k_n \int_0^\infty G^2(f)df$$

where k_n is a measure of the noise input power per unit bandwidth and $G(f)$ is the gain at a frequency f. Now if the signal was also *uniformly* distributed over the band we could compare the noise power to the signal power by using k_n only, without regard to the integral, which would be the same for both signal and noise. This would be true, for example, if we used a calibrated noise source as the source of signal. Actually, in most measurements and tests, a single-frequency signal is used. Suppose this signal is tuned in at a frequency f_0. Then our signal output power is

$$\text{signal power} = k_s G^2(f_0)$$

where k_s is a measure of the signal input power. Then

$$\frac{\text{noise power}}{\text{signal power}} = \frac{k_n}{k_s} \frac{\int_0^\infty G^2(f)df}{G^2(f_0)} = \frac{k_n}{k_s} \Delta f$$

and we define Δf as the noise bandwidth. What it represents is simply the bandwidth of a rectangle of height $G^2(f_0)$ having the same noise as $\int G^2(f)df$. It is clear that Δf is altered if the frequency f_0 of the test signal is altered. It differs markedly, therefore, from the circuit bandwidth $\Delta f'$, which was shown in Fig. 1.

In this discussion, Δf will stand for the noise bandwidth and $\Delta f'$ for the circuit or receiver bandwidth for signal purposes, as exemplified in Fig. 1.

3. Selectivity

The final item which is needed in a receiver is *selectivity*. We must be able to distinguish between two signals of different frequency so as to select one of them. The superheterodyne principle is universally employed to improve selectivity since, with it, the same selectivity is obtained for all stations and the intermediate-frequency amplifier is easily designed to meet any requirements. With the superheterodyne there is, however, one factor which gives trouble and that is the response to the so-called image frequency. A signal whose frequency is *higher* than that of the local oscillator will give the same intermediate frequency as one which is *lower* by the same amount and both can be heard. If we use a high intermediate frequency the two responses are far apart and our radio-frequency signal circuit will separate them. This is the cure usually adopted at ultra-high frequency. In a broadcast receiver, the intermediate frequency is of the order of 450 kilocycles. In the ultra-high-frequency receiver, it is commonly above 10 megacycles and is sometimes as high as 100 megacycles when really excellent image reduction is desired.

III. The Receiving Antenna

1. Induced Signal Voltage, Directivity, and the Equivalent Circuit

The purpose of the receiving antenna is to "capture" some of the transmitted radiation so as to enable one to utilize the transmitted intelligence. It should be made clear at the outset that the effectiveness of the receiving antenna in picking up the transmission imposes a fundamental limitation on the signal-to-noise ratio. We cannot think about, or compute, the absolute sensitivity of a receiving system without first knowing the absolute sensitivity of the antenna. Actually a receiving antenna may be considered as having an induced electromotive force (induced by the transmitted signal) and an internal resistance, the radiation resistance. Thus the available received signal power is

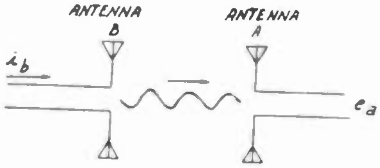


Fig. 3—Transmission of signal from antenna B to antenna A. A current i_b in the transmitting antenna induces an open-circuit electromotive force e_a in the receiving antenna.

a definite quantity which we must know before the signal-to-noise ratio can be discussed in Part III of this series. We shall now demonstrate that this received signal power at a given wavelength is a function of only one quantity which can be controlled at the receiver, namely, the receiving-antenna directivity. As a result, the design of the receiving antenna is concerned only with directivity, which determines the maximum signal-to-noise ratio and with bandwidth or Q which determines the useful frequency range.

For most ultra-high-frequency applications, the so-called reciprocity law holds and it may be used to demonstrate the fundamental property of the receiving antenna. Following the analysis of D. O. North,[2] if antenna B transmits to an antenna A, as in Fig. 3, we know that the open-circuit voltage at antenna A is proportional to the current into B; i.e.,

$$e_a = ki_b.$$

The reciprocity law in one of its forms[3] tells us that, if antenna A is used at the transmitter end, k is the same in both directions; i.e.,

$$e_b = ki_a.$$

We also know that antenna B will have a radiation resistance R_b, and radiates a power $i_b{}^2R_b$, so that the receiving antenna A will have a voltage proportional to the square root of this power,

[2] D. O. North, "The absolute sensitivity of radio receivers," *RCA Rev.*, vol. 6, pp. 332–343; January, 1942.
[3] S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Co., New York, N.Y., 1943.

$$e_a \propto \sqrt{R_b}\, i_b.$$

But if A were the transmitter

$$e_b \propto \sqrt{R_a}\, i_a.$$

So we see that k must be proportional to the square root of the radiation resistances of both transmitting and receiving antennas. Thus

$$e_a = k'\sqrt{R_a R_b}\, i_b.$$

Now we know more about k'; we know that it becomes smaller as our distance increases, and we also know that it depends on the product of directivities of receiving and transmitting antennas. Thus we may write

$$e_a = \sqrt{A}\, f(d) D_a D_b \sqrt{R_a R_b}\, i_b$$

or, squaring

$$e_a^2 = A f^2(d) D_a^2 D_b^2 R_a R_b i_b^2$$

where A is some constant and D_a^2 and D_b^2 are directivities. North[2] defines the directivities to represent the ratio of effectiveness of transmission or reception at a particular direction and polarization angle to the average over all directions and angles. That is, D_a^2 represents the power gain of the receiving antenna over a hypothetical completely nondirectional antenna.[4] We may associate $S = f^2(d) D_b^2 i_b^2 R_b$ with the transmitter, and write

$$e_a^2 = A R_a D_a^2 S$$

where S is not controllable at the receiver. Actually S represents simply the fraction of the total transmitter power which is effective at the receiver. It may be expressed as the density of radiation in watts per square meter and is called the Poynting vector. The relation between S and the field strength E is simple and is

$$S = \text{watts/meter}^2 = E^2/120\pi$$

where E is in volts per meter and 120π has the dimension of ohms. For any received radiation, S has a definite direction and D_a is also a function of direction. Finally, it was shown by North[2] that $A = \lambda^2/2\pi$. Hence

$$e_a^2 = \frac{\lambda^2}{2\pi} R_a D_a^2 S.$$

All this is perfectly reasonable. First, consider the dependence on λ^2. We know a half-wave dipole has a radiation resistance of some 73 ohms and a certain directivity pattern, no matter what the wavelength is. But surely at $\lambda = 1$ meter where the dipole is $\frac{1}{2}$ meter long we would expect less pickup than at 10 meters where the dipole is 5 meters long. Ordinarily we say

$$e_a = Eh$$

where E is the field strength and h the "effective height." For a half-wave dipole, approximately, $h = (2/\pi)l$ where l is the length $\lambda/2$ so that we see in this example that the square of e_a is proportional to λ^2. The same should be true of any antenna.

Next it is seen in the equation that e_a^2 depends on R_a, the radiation resistance. This is again reasonable. The quantity e_a^2/R_a is a power and is proportional to the maximum power we might abstract from the receiving antenna. Surely this depends on the radiation density S from the transmitter and $\lambda^2 D_a^2/2\pi$ is a sort of "capture" area. That is, S is watts per square meter and the effective area of the receiving antenna is $\lambda^2 D_a^2/2\pi$.

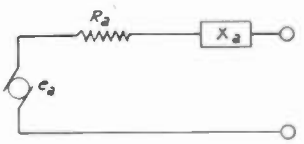


Fig. 4—Equivalent circuit of receiving antenna. At a given wavelength the quantity e_a^2/R_a is a function only of the directivity and the field strength of the transmitter. The antenna reactance is represented by X_a.

The importance of these considerations is that we now know that, no matter what the antenna is like, its open-circuit voltage depends only on the wavelength, the directivity, the radiation resistance, and, of course, on the intensity of received radiation. We may substitute for the real antenna, then, an equivalent circuit consisting of a generator e_a, a resistor R_a, and the antenna reactance X_a, as in Fig. 4. For a given transmitter and a given wavelength, we see that e_a^2/R_a is only affected by a change in the directivity. In other words, the maximum power we can abstract from the receiving antenna can only be increased by increasing the directivity; we cannot, for example, redesign an antenna to give a different radiation resistance and expect improved reception unless, at the same time, the directivity has been increased.

Let us look into this property of directivity and "capture" area a little further by again considering the half-wave dipole as an example. The directivity factor D_a^2 is approximately 3 so that the effective "capture" area is

$$\text{area} = \frac{\lambda^2}{2\pi} D_a^2 \approx \frac{\lambda^2}{2} \approx 2l^2$$

or the area of a square each of whose sides is $\sqrt{2}l$. Thus the half-wave dipole captures the radiation from a square whose diagonal is the wavelength. For a large reflector or some other type of array large compared with the wavelength, we should expect that the maximum "capture" area is fairly close to the actual area; i.e., that e_a^2/R_a is approximately equal to the product of S and the actual area. We can, therefore, get some idea of the directivity of a large reflector or array by its area. The power gain of a large array over a half-wave dipole will be given very approximately by the ratio of the area to $\lambda^2/2$ and is

[4] By this definition the half-wave dipole has a value D_a^2 of about 3.

power gain of reflector or array

$$= 2\,\frac{\text{area of array}}{\lambda^2}\ \text{(very approximate).}$$

This assumes that the area of the reflector is large compared with the wavelength.

2. *Antenna Frequency Range (Bandwidth)*

Although we have now found a first equivalent circuit for the antenna it is still necessary to discuss the nature of the antenna reactance. For practically all ultra-high-frequency applications, the antenna is tuned by itself; that is, our antenna reactance is practically zero somewhere in the frequency range. This means we need not tune out the reactance at that frequency but, on the other hand, we must examine the frequency limits over which our antenna may be used. Excellent papers[5–7] on antennas and their bandwidths are available and should be consulted. We will look into the simple half-wave dipole as an illustrative example, since this antenna is the basis of practically all ultra-high-frequency designs.

Fig. 5—A half-wave dipole formed by two conical conductors of apex angle 2ψ. Such a biconical dipole presents a uniform surge impedance to waves traveling along the length of the conductors.

To begin with, the ordinary half-wave dipole is nothing but an opened-out transmission line. However, if the dipole is uniform in cross section, it is a line of variable surge impedance. A very simple change which makes very little difference physically but which makes all the difference theoretically, is to assume that the conductors are conical.[7] Then the surge impedance is uniform and a wave starts at the center and travels out to the ends as on a uniform line. The surge impedance of such an antenna with the conical conductors having an angle 2ψ at the apex (Fig. 5) is

$$Z_0 = 120 \log_e \cot \psi/2$$

$$\approx 120 \log_e \frac{2r}{\rho}\ \text{(for small } \psi\text{'s)}$$

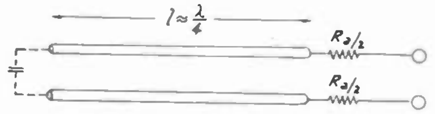


Fig. 6—Parallel-wire line circuit which is equivalent to the biconical dipole of Fig. 5. The radiation resistance is shown in lumped form. A small capacitance at the open end is also due to radiation.

where ρ is the radius of the cone at a distance r from the apex. This approximate expression is identical with that for the surge impedance of a parallel-wire line of spacing $2r$ and radius of wire ρ. Thus we may replace the dipole by a uniform open-ended $\lambda/4$ line, Fig. 6, but there is one difference between this equivalent line and the lines usually used. Usually the line spacings are small compared with the wavelength and radiation may be neglected. In the dipole equivalent, however, radiation is the major source of loss. For all the practical cases of reasonably small ψ, the radiation effects may be lumped as a 73-ohm (approximately) resistance at the input terminals and a reactance term which acts as a small capacitance at the end[7] as shown in Fig. 6. By shortening the dipole slightly, we may forget about the reactance since this restores tuning to the originally contemplated wavelength. We may remember that an open-ended $\lambda/4$ line behaves like a series resonant circuit.[8] The lumped-circuit equivalent of the dipole is

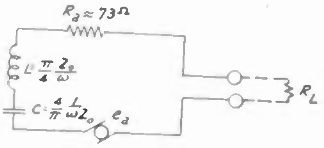


Fig. 7—Equivalent lumped circuit of typical dipole receiving antenna near its half-wave resonance.

therefore as shown in Fig. 7. L and C are the equivalent inductance and capacitance and are determined in terms of Z_0.

Now we may easily find the bandwidth of the antenna, near its $\lambda/2$ resonance. The Q is

$$Q = \frac{\omega L}{R} = \frac{\pi}{4}\,\frac{Z_0}{R_a + R_L}$$

where R_L is the load imposed by the transmission line or receiver. Thus, if Z_0 is high, the Q is high and the bandwidth is narrow. The actual antenna bandwidth for 3 decibels down is just $1/Q$ times the resonance frequency. However, as very clearly shown by Carter,[5] it is not advisable to use an antenna as far out in frequency as its 3-decibel down band edge if one also wishes to use a long low-loss transmission line between antenna and receiver. Near such a band edge the antenna reactance approximately equals the antenna resistance and the impedance at the far end of the transmission line (as seen by the receiver) goes through very extreme variations for a rather small change in frequency. For example, consider a transmission line of length l. Its electric length is

$$\theta = \frac{2\pi l}{\lambda} = \frac{2\pi f l}{v}$$

where v is the velocity of propagation on the line. Thus,

[5] P. S. Carter, "Simple television antennas," *RCA Rev.*, vol. 4, pp. 168–185; October, 1939.

[6] N. E. Lindenblad, "Television transmitting antenna for Empire State Building," *RCA Rev.*, vol. 3, pp. 387–408; April, 1939.

[7] S. A. Schelkunoff, "Theory of antennas of arbitrary size and shape," Proc. I.R.E., vol. 29, pp. 493–521; September, 1941.

[8] See the Appendix to this part for an outline of lumped-circuit equivalents to transmission-line circuits.

the rate of change of the electrical length with the frequency is

$$\frac{d\theta}{df} = \frac{2\pi l}{v}$$

which increases linearly with the length. Thus for a long line a very small change in frequency will result in a $\lambda/4$ change in line length. This causes an extreme impedance nonuniformity at the receiver when a reactive term is present in the antenna impedance, as occurs near the band edge. The curves given by Carter[5] show the results in striking fashion. The antenna should be designed, therefore, to give a much lower Q than that required by the frequency range, at least whenever a transmission line is used.

In most practical cases, dipoles are cylindrical in cross section rather than conical. However, the above results are only slightly affected by this difference in shape. As Schelkunoff[7] has shown, the cylindrical dipole may be considered roughly in terms of an *average* surge impedance along its length which is

$$Z_0 = 120 \log_e \frac{2l}{a} - 120$$

where $2l$ is the total length and a is the radius. Thus the surge impedance is just 120 ohms less than that of the narrow-angle biconical dipole of over-all length $2l$ and end radius a. In other words the Q of the cylindrical dipole is slightly less than that of the small-angle biconical dipole. The wide-angle cone behaves quite differently and will not be treated.

At 300 megacycles a cylindrical dipole of 1-inch diameter at the end will have a Z_0 of approximately 325 ohms and a Q of 4 when short-circuited. The Q is halved when working into a matched load. This means a maximum frequency range (3 decibels down) of from 263 to 337 megacycles when short-circuited or something approaching from 225 to 375 megacycles when matched. However, if our dipole had been a number 36 wire, Z_0 would be 980 ohms and Q would be 11. This would still give a 27-megacycle frequency range. If a long low-loss transmission line is used with the antennas of the above examples, it would be wise to consider the useful frequency range to be only a fraction of the figures given so as to avoid extreme variations in the impedance seen at the receiver end when tuning from one end of the range to the other.

To summarize, a *cylindrical* dipole of total length $2l$ and radius a will have a short-circuited Q of approximately

$$Q \approx 1.4 \left(\log_e \frac{2l}{a} - 1\right)$$

and, of course, a lower Q when operated into a load. The frequency range of the antenna is given roughly by f/Q (where f is the resonant frequency of the antenna) when short transmission lines are used, but is much less with long, low-loss lines.

3. *Random Noise of the Antenna*

An antenna picks up radiation from all points within its directivity pattern. Since, due to temperature, there is a thermal-energy exchange which is random as regards frequency, we might expect one type of noise which is equivalent to thermal-agitation noise. It has been shown[2] that, if an antenna is placed in an enclosure at a uniform temperature T_a, the noise it will pick up is the same as thermal agitation in its radiation resistance

$$\overline{e_n^2} = 4kT_aR_a\Delta f$$

where $\overline{e_n^2}$ is the open-circuit noise voltage, $k = 1.37 \times 10^{-23}$ joule per degree Kelvin, and Δf is the effective noise bandwidth as previously defined.

Actually, an antenna is not in such an enclosure, but, if it receives its entire radiation from *within* the Heaviside layer, it would amount to the same thing. At ultra-high frequency, however, we would expect the waves to pass right through the Heaviside layer so that if we have an antenna directed toward interstellar space, it might be receiving little or no thermal noise; i.e., its effective temperature, as far as noise is concerned, would be zero. It is interesting to realize that, with such a condition, a resistor connected to the antenna would be cooled at a very slow rate by the radiation at antenna frequency. In practice, a directive antenna might be turned in almost any direction and the noise will presumably vary considerably. Because receiver noise is so high, it has not been possible to measure the effective antenna temperature at ultra-high frequency.

Of course, there are other sources of noise received by the antenna, and, at *low* frequencies, these far outweigh the thermal noise we have been talking about. At ultra-high frequency it is no longer certain that other noise sources than actual man-made interference are present and we may usually assume that, under favorable conditions, the antenna noise will be as low or lower than thermal noise corresponding to room temperature.

4. *Dummy Antenna for Laboratory Measurement*

For measurement of a receiver in the laboratory, we may replace an actual antenna by an equivalent circuit. As we already saw (Fig. 4) a known voltage source, a resistance equal to radiation resistance, and a reactance, represent our antenna. If the dummy antenna is to be complete, R_a should be at the temperature T_a corresponding to antenna thermal noise. If other sources of noise are present, they should be added. Until measurements to the contrary are made, it is convenient to assume that a lower noise than T_a = room temperature is not to be expected. It is, therefore, a simple matter to replace the radiation resistance R_a by a lumped resistance at room temperature.

As to the reactance, in almost every ultra-high-frequency application, the antenna is tuned; i.e., its reactance is zero or very small in the frequency band in which we are interested. We may go further and assume that, if the antenna is correctly designed, its bandwidth is so wide (i.e., its Q is so low) so that the reactance is negligible at any point near its resonance. Thus the dummy antenna circuit may omit X_a, the reactance.

It is now possible to outline the chief requirements of a signal generator for receiver measurements since such a generator must supply the known voltage source. It should have a known internal impedance,

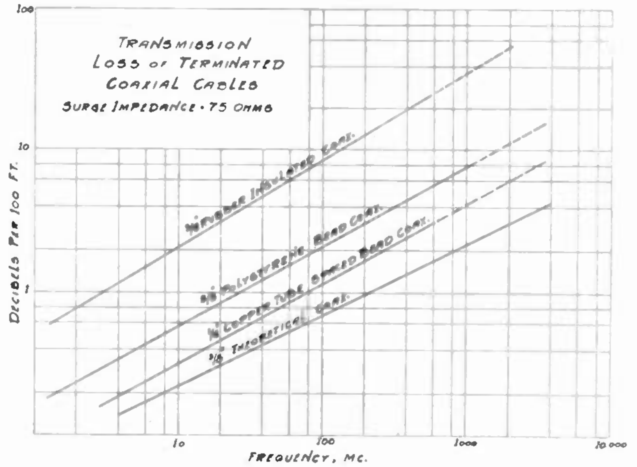


Fig. 8—Attenuation of properly terminated coaxial cable as a function of frequency.

which is a pure resistance, if possible. If it contains internal reactance, the latter must either be tuned out or be very small in magnitude compared with the radiation resistance (i.e., the equivalent dummy-antenna resistance). If the signal generator has a low internal impedance, the dummy antenna may be added externally. Otherwise the internal impedance must be considered as part or all of the dummy antenna.

At the very high frequencies, it is extremely difficult to get a lumped, reactance-free resistance to replace the antenna-radiation resistance for use in the dummy antenna. Ordinary carbon resistors (e.g., of 75 ohms direct-current resistance) are entirely satisfactory over much of the ultra-high-frequency region, however. They have an effective inductance of the order of 10^{-8} henry, which is not too bad; they should be regarded with suspicion at those frequencies at which their reactance exceeds 20 to 30 ohms. One expedient which may be used at the very high frequencies is to connect a long length of transmission line with appreciable loss between the signal source and the receiver. Such a line will look like its surge impedance and, if the loss per wavelength is not too great, it will be almost resistive and independent of frequency. The line loss becomes the equivalent of the radiation resistance in the dummy antenna. However, the signal output voltage must be continuously calibrated by some independent means since the loss of many transmission lines varies markedly with temperature and humidity.

IV. Coupling the Antenna to the Receiver

1. Magnitude of Transmission-Line Losses versus Frequency

Let us now consider how to run a lead from the antenna to the receiver. As a first approximation the antenna may be considered as a resistance R_a with a voltage generator e_a and the simplest connection is made by using a transmission line of impedance R_a. If this is possible, at the receiver end, we shall see an impedance R_a and a voltage generator of reduced magnitude, depending on the line losses. The line losses are most easily expressed in decibels per 100 feet and typical values are given in the curves of Fig. 8. Obviously, every decibel of line loss must be supplied by the transmitter and a 3-decibel loss means doubled transmitter power at the noise-level limit. It is, therefore, highly desirable to place early tubes of a receiver near the antenna.

A word in regard to choice of surge impedance. An air-insulated, solid-outer-conductor, coaxial cable has only copper loss which is a minimum at slightly over 75 ohms impedance.[9] For this reason 75 ohms has been a widely used value. The impedance which will give highest voltage-handling capacity without breakdown is about 60 ohms, and highest power-handling capacity without breakdown about 30 ohms. In cables not air-insulated, the losses are increased by the dielectric and the impedance for minimum attenuation is usually less than 75 ohms but is not the same for all cables. Balanced open-wire lines are much higher in impedance but twisted-pair or balanced shielded lines are usually between 75 and 150 ohms. At the lower frequencies, where radiation loss is not too serious, the high-impedance, balanced, open-wire line is probably the lowest cost line for a given attenuation and has had frequent application. At the higher frequencies, however, it is desirable to use coaxial lines because of their low loss and freedom from radiation; there is a growing tendency to choose an impedance of either 50 or 75 ohms for such coaxial lines.

For further information attention is directed to the early and excellent paper of Sterba and Feldman,[9] much of which is applicable to ultra-high-frequency practice.

2. Matching Circuits for Antenna to Line

When a balanced antenna such as a dipole or dipole array is used, even though the line impedance may match it, in order to preserve the balance, it is desirable to make use of *balanced* lines or two parallel coaxial lines. Because of its shielding and lower loss, a *single* coaxial line is most widely used and we may well

[9] E. J. Sterba and C. B. Feldman, "Transmission lines for short-wave measuring systems," Proc. I.R.E., vol. 20, pp. 1163–1202; July, 1932.

review the special coupling methods which permit its use without detriment to a balanced antenna. The most obvious method is the quarter-wave skirt. To change a coaxial to a balanced line, we simply make a double coaxial for a length equal to a quarter wavelength as in Fig. 9. This causes both inner and outer conductors to have a high impedance to ground. The quarter-wave section, being tuned, acts as an open circuit at its resonance. However, although the arrangement is ideal only at resonance, its bandwidth is fairly wide. The impedance of the outer coaxial section at an angular frequency ω is

$$Z = Z_0' \tan \frac{2\pi l}{\lambda} = Z_0' \tan \frac{\pi}{2} \frac{\lambda_0}{\lambda} = Z_0' \tan \frac{\pi}{2} \frac{\omega}{\omega_0}$$

where Z_0' is the surge impedance of the outer coaxial section and ω_0 is its resonant angular frequency. If the inner coaxial has an impedance Z_0, and Z_0' is made much higher than Z_0, the band extends from $1/2\ \omega_0$ to $3/2\ \omega_0$ with safety. If Z_0' is comparable with Z_0,

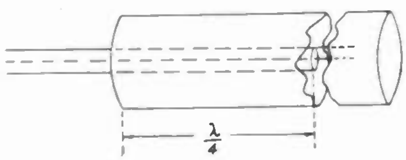


Fig. 9—The quarter-wave skirt for connecting the unbalanced (coaxial) line to a balanced line (or antenna). The coaxial is extended a quarter-wave inside the outer skirt.

safer limits are $0.8\ \omega_0$ to $1.2\ \omega_0$, i.e., a bandwidth of 40 per cent. For example, such a matching skirt at 300 megacycles would cover a band from 240 to 360 megacycles.

A second means of coupling a coaxial line to a balanced antenna is shown in Fig. 10. This is suitable only for vertical antennas. It operates by folding back the outer conductor of the coaxial line to form the lower half of the radiating dipole. There are also other

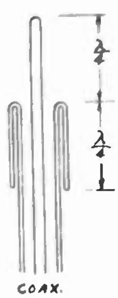


Fig. 10—A method of connecting a coaxial transmission line to a vertical half-wave dipole.

means whereby the coupling from an unbalanced transmission line may be made to a balanced antenna, many of which have been described in the patent literature.

Now suppose a balanced antenna, or balanced line is to be connected to a coaxial line of a different surge impedance. Here one may use the basic impedance transformer, the $\lambda/4$ section of line.[9] In a $\lambda/4$ line terminated in a resistor R the input looks like

$$Z = \frac{Z_0^2}{R}$$

so that we may use this section to step up R to a higher value. Thus, to match two lines of different impedance Z_1 and Z_2, we may join them by a $\lambda/4$ section whose impedance is

$$Z_0^2 = Z_1 Z_2.$$

This is true for any type of line, balanced or coaxial. We may, then, modify the skirted coaxial line, as in Fig. 11, by a reduced inner-conductor diameter. How-

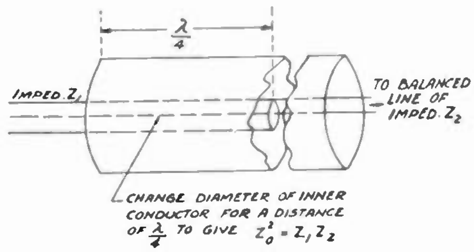


Fig. 11—The use of a combination of quarter-wave skirt with a quarter-wave transformer for matching an unbalanced line of one surge impedance to a balanced line of another surge impedance.

ever, the total bandwidth free from reflection is now greatly reduced. An impedance transformation of more than 2:1 is good only for frequency deviations of a few per cent.[5] By using two $\lambda/4$ sections and a progressive impedance rise, greater bandwidths are obtainable. This leads to the impedance matching by a

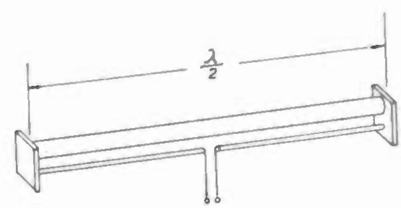


Fig. 12—An impedance-raising transformer for a half-wave dipole antenna. The impedance ratio is determined by the relative diameters of the two parallel dipole conductors.

tapered line. If two lines of different surge impedance are joined by a tapered section several waves long, a good wide-band match is obtained with very little reflection loss.

The best-known impedance-raising transformer for the dipole is the folded-dipole arrangement wherein the antenna itself forms the transformer as in Fig. 12. The paper by Carter[6] gives all details.

3. *The Connection of Transmission Line to Receiver*

In connecting the antenna to the receiver through a transmission line, an impedance mismatch at the receiver end leads to standing waves on the line and thus increases the line losses somewhat. Thus, when the receiver is strictly a power-operated device, it is desirable to maintain an impedance match at the receiver end as well as at the antenna. On the other hand, a receiver is not necessarily power-operated[10] and there is sometimes an improvement in signal-to-noise ratio or in frequency response by a mismatch at the receiver terminals. This aspect of receiver design will be taken up in more detail in Part III of this series. When, in

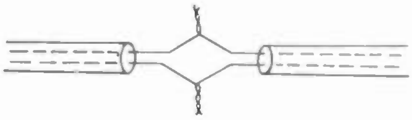


Fig. 13—Two pieces of balanced low-impedance transmission line connected by a joint which may give rise to serious reflection.

such a case, there is a mismatch at the receiver, the the resulting reflection will lead to an increased apparent power loss but, since optimum power transfer is not the criterion of receiver performance, this is not pertinent. However, when the receiver is mismatched to the line it is important to maintain a good match at the antenna end or else there will be a decrease in signal-to-noise ratio due to increased line losses.

Thus, we may lay down the general rule that, when the best receiver performance is obtained with maximum power to it, the transmission line should be matched at both ends. When a mismatch improves the receiver performance, the transmission line should be matched to the antenna and the mismatch should occur between the transmission line and the receiver.

If an antenna is operated at a frequency slightly off from its natural resonance, its impedance presents a

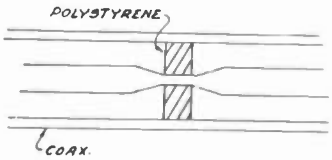


Fig. 14—Method of introducing a polystyrene support in a coaxial line so as to minimize reflection.

small reactance which does not greatly affect the magnitude of the impedance but which nevertheless creates a mismatched condition to a transmission line. The result at the other end of a long, low-loss line is an impedance which may change appreciably with a small change in frequency, as clearly shown in the curves by Carter.[6] With a wide-band receiver, the impedance change may appreciably affect the frequency response. The effects are minimized by using the lowest possible antenna Q and the shortest possible transmission-line length. The effects are also less severe when the line losses are appreciable.

4. *Reflection Loss by Line Couplings*

While on the subject of impedance match the effect of connectors must not be forgotten. When two pieces of line are joined by a connector, care must be taken to see that the reflection at the connection is minimized. For example, two pieces of 75-ohm balanced line might be joined by an open connection as in Fig. 13. When the frequency is high, this connection can easily cause an appreciable loss. By pushing the two pieces of wire at the joint close together and adding some rubber tape to lower the surge impedance, the loss can be greatly reduced. In designing connectors, therefore, one should try to match impedance. If a polystyrene support is used, in a coaxial cable, as in Fig. 14, its dielectric constant is 2.5. The inner conductor should then be restricted as shown so as to maintain $Z_0 = (1/v_0\sqrt{\epsilon}C_0)$ a constant where ϵ is the dielectric constant. In the case of a 75-ohm coaxial line, the ratio of diameters, when air insulation is used, is approximately 3.5 to 1. With polystyrene insulation, this ratio being 7.2 to 1, at a point of support, the inner conductor diameter should be cut roughly in half.

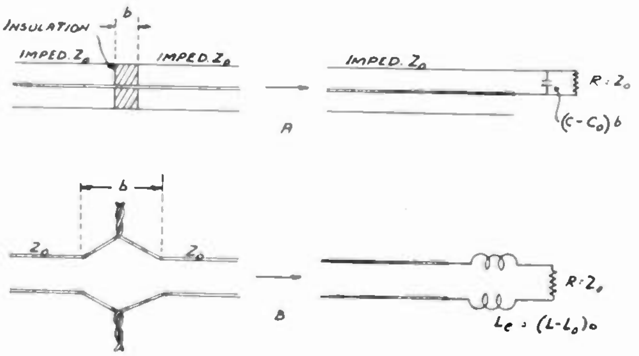


Fig. 15—Equivalent circuits which may be used to calculate the reflection at an insulating bead or a transmission-line connection; A, coaxial line with bead, B, balanced line with joint of higher surge impedance.

It is of interest to calculate the reflection coefficient when no such precautions are taken at a connection or point of support. An idea of the magnitude is readily obtained by assuming that the discontinuity represents added shunt capacitance or added series inductance depending on whether the surge impedance is decreased or increased at the point on the line under consideration. The added capacitance, or added inductance may be computed and the reflection ratio estimated from the usual formula

$$\text{reflection ratio} = \frac{Z - Z_0}{Z + Z_0}$$

where Z_0 is the normal surge impedance and Z is the combination of added capacitance in parallel with Z_0 or added inductance in series with Z_0. Fig. 15 shows the equivalent circuit for the two cases, i.e., a decrease in surge impedance at a point, and an increase in surge impedance at a point. Although the method is easily extended to include discontinuities of any length, the equivalent circuits shown in the figure are exact only

[10] For example, the receiver might conceivably have an infinite, or even a *negative* input resistance. The latter condition need not be an unstable one since the antenna loading may be made sufficient for completely stable operation.

for discontinuities short compared with the wavelength. With this approximation, the reflection ratio for the illustration of Fig. 15A can be shown to be

$$\text{reflection ratio} \approx j\pi(\epsilon - 1)\frac{b}{\lambda} \qquad (b \ll \lambda)$$

where $j=\sqrt{-1}$, ϵ is the dielectric constant of the bead, b is its length, and λ is the free-space wavelength of the transmitted radiation. For the case of Fig. 15B, where the reflection is due to a short length of increased surge impedance, as in the twisted joint of a parallel-wire line, the reflection ratio is

$$\text{reflection ratio} \approx j\pi\left(\frac{L}{L_0} - 1\right)\frac{b}{\lambda} \qquad (b \ll \lambda)$$

where L is the average inductance per unit length at the joint and L_0 is the normal inductance per unit length of the line. It should be noted that the maximum possible power-loss ratio due to reflection alone is given by the square of the reflection ratio in the above cases. The total power loss can be less than this maximum under special conditions but it can also be higher if there is any loss in the discontinuity (which was assumed purely reactive in the above analysis).

We may notice that the reflection ratio becomes worse as the wavelength is decreased; i.e., as we approach higher frequencies. For example, an isolated polystyrene disk only $\frac{1}{8}$ inch long may cause an appreciable loss when the frequency is sufficiently high. In cases where there are many regular reflections, such as in a beaded cable, the losses can be reduced by proper spacing of the beads.

V. The Receiver Input Circuit

1. *Selectivity and Image Response*

The purpose of the receiver input circuit is first to match properly the antenna and its associated transmission line to the first tube of the receiver, and second to provide some selectivity, particularly against the image response. Let us consider the second purpose first.

If we are tuned to a frequency f_s and the intermediate frequency is f_i then we know that our local oscillator frequency f_0 will be given by

$$f_s \pm nf_0 = f_i$$

where n is an integer. Usually $n=1$ and we operate at the fundamental of the local oscillator. The receiver will respond at frequencies

$$f = f_0 - f_i, f_0 + f_i, 2f_0 - f_i, 2f_0 + f_i, 3f_0 - f_i, 3f_0 + f_i$$

and the intermediate-frequency circuit will not distinguish between them. The input circuit must select the desired one and discriminate against the others. Now if one of these is chosen as the signal, the next nearest one differs by the frequency $2f_i$ so the input must discriminate between a frequency f_s and a frequency $f_s \pm 2f_i$. For most cases, f_i is not more than $\frac{1}{5}$ of f_s so that, if we have an input circuit tuned to f_s, and it has a reasonably symmetrical response, the rejection of f_s+2f_i and f_s-2f_i will be the same, and it makes little difference upon which side of the signal the local oscillator frequency is placed.

Many input circuits, though they may be tuned sections of transmission line, have a lumped circuit

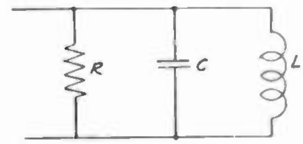


Fig. 16—Equivalent lumped-circuit analog to many resonant-line input circuits.

equivalent which is a simple parallel-tuned circuit such as shown in Fig. 16. Such a circuit has a bandwidth

$$\Delta f' \text{ (3 decibels down)} = \frac{f_s}{Q} = \frac{1}{2\pi CR}$$

and in using such a circuit for selectivity it is necessary to tap the unavoidable load resistances (e.g., the tube input) down on the circuit so as to obtain the desired bandwidth (or, if one likes, the desired Q). At the image frequency (f_s+2f_i) such a circuit behaves as a reactance

$$x_c = \frac{1}{2\pi C}\frac{f_s + 2f_i}{4f_i(f_s + f_i)}.$$

Thus the ratio of its response at image frequency to its response at signal frequency is

$$\frac{\text{image response}}{\text{signal response}} = \frac{1}{R}\frac{1}{2\pi C}\frac{f_s + 2f_i}{4f_i(f_s + f_i)} = \frac{\Delta f'}{4f_i}\frac{f_s + 2f_i}{f_s + f_i}$$

$$\approx \frac{\Delta f'}{4f_i} \text{ when } f_i \ll f_s.$$

The value of a high intermediate frequency is readily apparent.

A pair of coupled circuits has much greater discrimination against image response, for the same bandwidth. In fact, if the intermediate frequency is small compared with the signal frequency, the ratio is approximately

$$\frac{\text{image response}}{\text{signal response}} \approx \left(\frac{\Delta f'}{4f_i}\right)^2$$

which is just the square of the single-tuned circuit ratio, or the equivalent of two single-tuned circuits in cascade.

2. *Signal-to-Noise Ratio*

Although major consideration of the signal-to-noise ratio will be reserved for Part III of this series, it may be well at this point to emphasize the effect which the input circuit may have on this important factor. Although there are some instances where important noise sources are present in a receiver ahead of, or in the input of the first tube, in very many cases the major noise source occurs in the output of the first tube. In the latter event, best signal-to-noise ratio will be obtained by designing an input circuit to deliver

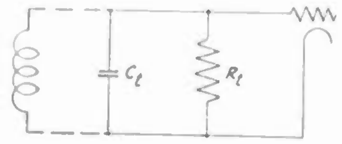


Fig. 17—The connection of a loss-free inductance to a tube of input resistance R_t and capacitance C_t to form an input circuit.

the maximum possible signal voltage on the input electrode. This implies, first of all, a reduction of all losses to a minimum which is usually imposed by the electronically active portion of the tube input and second, it implies impedance matching from the antenna or transmission line to the tube input. The first implication is a valid one in all cases.

It is, however, a mistake to worship the matched-impedance condition to the extent to which it is sometimes done. In an ideal receiver with no losses in the input tube or circuit, for example, impedance matching is also equivalent to zero bandwidth and is not a practicable operating condition. By a mismatch to the antenna, the bandwidth may be increased with a minimum loss in signal and hence a minimum reduction in signal-to-noise ratio. In many receivers, there are sources of noise present in the input (such as thermal agitation in the input circuit and induced noise in the tube input electrodes) which cannot be neglected. Again a mismatch to the antenna is advantageous and leads to an improved signal-to-noise ratio.

The input circuit is, therefore, an important part of receiver design for best signal-to-noise ratio, since it permits an adjustment of the coupling between the antenna and the tube input.

3. *Wide-Band Considerations*

We have already introduced the bandwidth $\Delta f'$ into selectivity considerations. This is a bandwidth which must usually be considered in the initial design. If the image rejection is sufficient, one may, however, let the radio-frequency circuit bandwidth be as wide as desired since the intermediate-frequency circuits will give adequate selectivity. But the radio-frequency bandwidth cannot be permitted to be too narrow. Since the bandwidth of a circuit depends inversely on its Q and for the same resonant impedance depends on the total effective circuit capacitance, we usually desire low capacitances.

In wide-band ultra-high-frequency receivers, it is invariably found that the circuits may be made very low loss and most of the losses are in the tubes, leads to the tubes, etc. In fact we may neglect the circuit losses, as a rule, and assume the circuits to be perfect, but we must be careful to introduce the tube and lead losses properly. Suppose we have a tube and we find that it behaves like a resistance R_t in parallel with a capacitance C_t. If we connect a loss-free inductance (with no distributed capacitance) we get a circuit (Fig. 17) of

$$Q = \omega C_t R_t$$

$$\Delta f' = \frac{f}{Q} = \frac{1}{2\pi C_t R_t}$$

and the smaller the value of C, the wider is the band. If, as always happens, the added inductance has a distributed capacitance, the bandwidth is always less than that expected with an ideal inductance. When the inductance consists of a transmission line its effective capacitance is sometimes as high as half of its total capacitance.[11] Thus wide-band circuits require low-C line circuits, or high surge-impedance line circuits.

It is seen that when R_t is increased, C_t must be decreased to keep the same bandwidth. We shall see later that the bandwidth may be varied by the effect of antenna loading on the input circuit.

4. *Equivalent Circuits of Input Transformers*

We have talked a great deal about line circuits but have not shown anything but lumped-circuit diagrams. This is because the easiest way for one versed in low-frequency technique to understand line circuits is in terms of lumped circuit equivalents. A table of useful data including such equivalent circuits has been prepared and is attached in an Appendix to this part. Only one arrangement will be discussed in detail here.

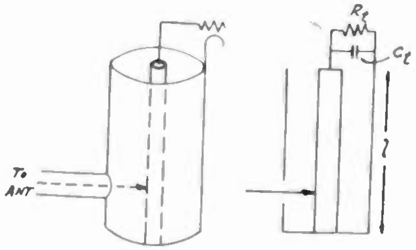


Fig. 18—A typical concentric-line, resonant circuit used as an input transformer with a tube of input resistance R_t and capacitance C_t. The length l determines the resonant frequency.

A typical input transformer to a tube will be a concentric cylinder,[12] quarter-wave line with a tap for the antenna such as shown in Fig. 18. Let us assume a connection to a tube whose parallel resistance is R_t and whose capacitance is C_t. The condition for resonance is then $Z_0 \tan \theta = 1/\omega C_t$ where θ is the electrical

[11] See Appendix for appropriate formulas.

[12] C. W. Hansell and P. S. Carter, "Frequency control by low power factor line circuits," Proc. I.R.E., vol. 24, pp. 597–619; April, 1936.

length of line $2\pi l/\lambda$. To obtain a transformer action, the antenna may be tapped down on the line. From the table in the Appendix, it is found that, with the antenna line disconnected,

$$Q = \frac{1}{2}\left[\omega C_t R_t + \frac{\theta}{\sin^2\theta}\,\frac{R_t}{Z_0}\right].$$

If the antenna tap is moved up until the impedances are matched, the Q will be just half of this. By moving the tap up further, the Q can be lowered as much as desired (until the Q of the antenna itself is approached). The equivalent lumped circuit is as shown in Fig. 19. It is also possible to couple in the antenna by a loop inserted in the line. If this is done it is usually best to couple near the short-circuited end so as to link the strongest magnetic field. The coupling loop will have some reactance but, as a rule, this may be tuned out by a readjustment of tuning.

The tuning-up of the circuit and the adjustment of antenna coupling is often a cut-and-try process, with each adjustment made to get either highest signal into the receiver or highest over-all signal-to-noise ratio. Since the antenna equivalent is a voltage in series with

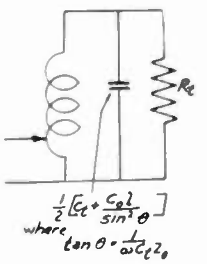


Fig. 19—Input circuit which is equivalent to the resonant-line circuit of Fig. 18. The line length l is included in the angle $\theta = 2\pi l/\lambda$ and C_0 is the capacitance per unit length of the resonant line while Z_0 is its surge impedance.

a resistance, optimum signal voltage on the first tube will occur with impedances matched. In the material which follows in Part III of this series, much more will be learned about antenna-coupling adjustment from the signal-to-noise point of view, and it will be found that coupling for the highest signal-to-noise ratio is not always the same as that which gives optimum signal on the tube.

Suppose in the above example that the antenna circuit has too narrow a bandwidth when it is adjusted for best gain (antenna coupled in). This means that Q is too high. Although there are many ways in which Q can be lowered, and they will all serve equally well to increase the bandwidth, they are not all alike in their effect on the signal-to-noise ratio. For example, if we add a shunt resistor for additional damping, the Q will be lowered but will be accompanied by marked reduction of signal gain and hence a reduction in signal-to-noise ratio. A second remedy is to couple the antenna more tightly than optimum. This will lower the Q with little change in the signal-to-noise ratio (it may even improve in some cases). A third possibility is the use of degenerative feedback in the first tube. If this is correctly done, by feeding back both signal and noise in like amounts, the effective tube input resistance may be reduced with little or no

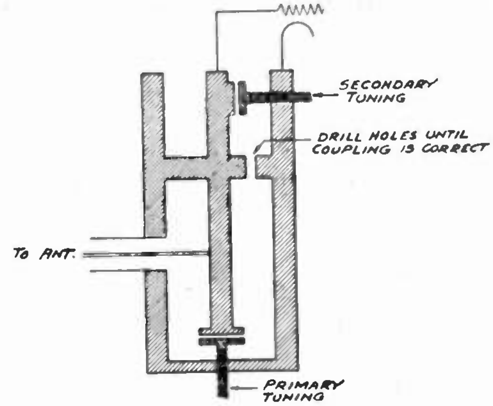


Fig. 20—A double-tuned transformer of resonant coaxial-line sections, as used for coupling antenna to the first tube of a receiver.

change in signal-to-noise ratio.[13] Finally, and best of all, the Q may be reduced by choice of a lower capacitance, higher input-resistance tube and the use of a lower capacitance circuit.

For wide-band input circuits, a considerable advantage in flatness of response and in image rejection is obtained by the use of a double-tuned input circuit. A possible design is shown in Fig. 20. In the figure, two end-to-end resonant coaxial-line circuits are shown, one as a primary which has the antenna coupled to it and the other used as a secondary circuit driving the first tube. The coupling between primary and secondary may be adjusted by drilling the central short-circuiting bar until the response curve has the proper flat-top shape. The use of capacitive tuning, as shown in Fig. 20, is simple and convenient but is disadvantageous in that the total circuit capacitances are increased.

Suppose, on the other hand, we have a single-tuned transformer and the bandwidth is so *wide* that we could well afford to narrow it so as to improve image response. Here we wish to raise the Q. Even with a double-tuned transformer we may find the secondary Q is too low. One method of raising the Q is to add low-loss capacitance preferably by use of a lower surge-impedance line circuit and this is probably a satisfactory remedy. Another expedient makes use of the fact that the Q of an unloaded, resonant-line transformer is usually very high. If we couple the tube more loosely to the line, as by tapping it down on the line, the over-all effective Q can be raised until it approaches the Q of the unloaded line alone. Part 2a of the Appendix gives a formula which is applicable when

[13] Further discussion will be reserved for Part III of this series entitled, "The Signal-to-Noise Ratio of Radio Receivers."

the tube, of capacitance C_t, resistance R_t, is tapped down on the resonant line. We may obtain a physical understanding of the behavior by considering the lumped-circuit analogy shown in Fig. 21. In A of the figure the tube is shown tapped down on a resonant circuit whose inherent capacitance is C_d and whose Q may be considered very high. The equivalent circuit shown in Fig. 21B, discloses that the over-all Q is

$$Q = \omega R_t\left(C_t + \frac{C_d}{m^2}\right)$$

where m is the effective turns ratio of the tapped-down section of the transformer to the total. Thus, if the

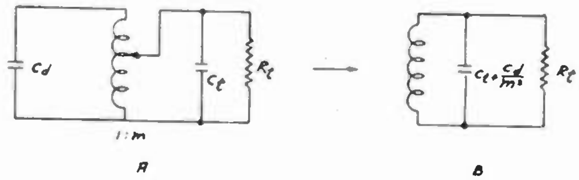


Fig. 21—Equivalent lumped circuits showing the effect of tapping a tube of capacitance C_t, resistance R_t down on a very high Q circuit. The over-all Q is thereby increased over a straight parallel connection.

tap is brought down on the circuit, m is made smaller and Q is made higher. Of course the upper limit to Q is imposed by the circuit losses which, in the case of resonant lines, are often small enough to permit attainment of a Q of 1000 or more. The antenna can be coupled into the circuit in the usual way and no appreciable loss of signal need result if the Q is raised a moderate amount by the tapping-down process.

As an example of a particular design of input transformer, suppose we have the following problem:

$$\Delta f' = 3.5 \text{ megacycles} \qquad \lambda_0 = 86 \text{ centimeters}$$

$$f = 350 \text{ megacycles} \qquad Q = \frac{f}{\Delta f'} = 100.$$

The input tube is to be a type 955 mixer used at oscillator fundamental. We will find that the input conductance of this tube as a mixer[14] is about 435 micromhos corresponding to $R_t = 2300$ ohms and the input capacitance C_t (considering the plate as radio-frequency ground) is about 4 micromicrofarads or so.

Let us use a concentric-cylinder line circuit, in the quarter-wave mode. Such a line, designed for maximum Q, may have an inner conductor of $\frac{1}{4}$-inch diameter copper, and a surge impedance of 77 ohms, corresponding to an inside diameter of 0.875 inch for the outer conductor. The line alone, if made a quarter wave long, would be 21.4 centimeters long and would have a Q as follows: (see Appendix for formula)

$$Q = 5.2 \times 10^3 \frac{1/8 \times 2.54}{\sqrt{86}} \approx 1780.$$

[14] E. W. Herold, "An analysis of the signal-to-noise ratio of ultra-high-frequency receivers," *RCA Rev.*, vol. 6, pp. 302–331; January, 1942.

This Q is obviously high enough to be neglected in comparison with the desired Q of 100, so we may use the data in the Appendix of equivalent circuits (which neglects line losses). We find, if the tube is connected to the open end of the line as in Fig. 18, that the 4-micromicrofarad capacitance will require the line length to be reduced to

$$l = \frac{\lambda_0}{2\pi}\theta = \frac{\lambda_0}{2\pi}\sin^{-1}\sqrt{\frac{1}{1 + (\omega Z_0 C_t)^2}}.$$

Using some of the relations given in the Appendix we find that

$$C_0 = \frac{33}{Z_0} = 0.43 \text{ micromicrofarad per centimeter.}$$

Thus

$$\omega Z_0 C_t = \frac{6.2}{\lambda_0}\frac{C_t}{C_0} = 0.67$$

$$l = \frac{\lambda_0}{2\pi} \times 0.98 = 13.4 \text{ centimeters}$$

and

$$\sin^2\theta = 0.69.$$

The equivalent lumped circuit gives, therefore, an equivalent-lumped-circuit capacitance

$$C = \frac{1}{2}\left[C_t + \frac{lC_0}{\sin^2\theta}\right] = 6.2 \text{ micromicrofarads}$$

and

$$Q = \omega C R_t = 31.$$

This gives a bandwidth of 11 megacycles even before connecting the antenna and is much too broad.

We must consider tapping down the tube on the line and we refer to the last of the cases treated in Part 2a of the Appendix. It must be remembered that we wish a final Q of about 100 but that when the antenna is coupled in, and approximately matched to the circuit, the Q will roughly be halved. The circuit by itself must therefore be designed for a Q of around 200, if the antenna impedance is to be somewhere near matched. This means the effective lumped equivalent capacitance must be about 40 micromicrofarads. Trying out $l_1 = 4.0$ centimeters, $\theta_1 = 0.29$ radian, we find

$$\cos^2\theta_2 = 0.12, \qquad l_2 = 16.6 \text{ centimeters}$$

so that

$$C = 42 \text{ micromicrofarads}$$

which gives a value of Q of 212. As shall be shown later, it is preferable to couple the antenna just a little more tightly than optimum and this will lower the Q to less than half. Thus, we may consider the choice as satisfactory and the antenna coupling can then be adjusted to give an over-all Q of 100. It should be noted that by this simple expedient of tapping the tube down on the line we have narrowed the bandwidth to the desired value without appreciable sacrifice in signal strength or reduction in signal-to-noise ratio.

Let us note that the image response ratio for an intermediate frequency of 25 megacycles is

$$\frac{\text{image response}}{\text{signal response}} = \frac{\Delta f'}{4f_i} = \frac{1}{29} \text{ or about 29 decibels.}$$

This is usually considered adequate. A local oscillator may be coupled in inductively or capacitively and the coupling should be very loose so as not to affect the impedance.

It is hoped that this example will suffice to show the utility of the lumped-circuit-equivalent formulas given in the Appendix. They can, of course, be used equally well for other input-circuit problems.

Appendix

Useful Data for Resonant Transmission-Line Circuits

In most applications to ultra-high-frequency receivers, the radio-frequency ohmic losses in the parts external to the tube are so small in comparison with the loading introduced by the tube, that they may safely be neglected. Thus, when the external circuits are resonant sections of line, we may safely neglect their distributed ohmic losses and make computations on the assumption that they are negligibly small. Under these conditions, a resonant section of transmission line, near one of its resonant modes, behaves in a fashion similar to lumped reactance elements. The loss and capacitance which are introduced by the connection of a tube, for example, are then readily affixed in lumped form to form an equivalent circuit. These

1—Uniform Low-Loss Lines.

a. Symbols

ϵ = dielectric constant of insulating material.
C_0 = capacitance per unit length of air-insulated line, farads per meter.
$C = \epsilon C_0$ = capacitance per unit length of dielectric-insulated line, farads per meter.
L_0 = inductance per unit length, henries per meter.
v_0 = velocity of propagation on air-insulated line $= 3 \times 10^8$ meters per second.
v = velocity of propagation on dielectric-insulated line $= v_0/\sqrt{\epsilon}$.
λ_0 = wavelength, meters on air-insulated line $= v_0/f$.
λ = wavelength, meters on dielectric-insulated line $= v/f$.
Z_0 = surge impedance, ohms.
l = length of line, meters.
f = frequency, cycles per second.
$\omega = 2\pi f$, angular frequency, radians per second.
n = any integer.

b. General Relations

$$v_0 = \frac{1}{\sqrt{L_0 C_0}} \qquad v = \frac{1}{\sqrt{\epsilon L_0 C_0}}$$

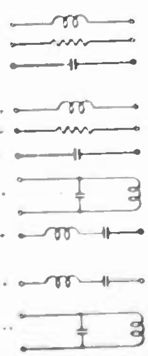
$$Z_0 = \sqrt{\frac{L_0}{C}} = \frac{1}{\sqrt{\epsilon}}\sqrt{\frac{L_0}{C_0}} = v\,L_0 = \frac{v_0}{\sqrt{\epsilon}}\,L_0 = \frac{1}{vC} = \frac{1}{v\epsilon C_0} = \frac{1}{v_0\sqrt{\epsilon}C_0}$$

c. Open and Short-Circuited Lines

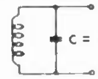
$$\text{Impedance of short-circuited line} = jZ_0 \tan\left(\frac{2\pi l}{\lambda}\right)$$

$$\text{Impedance of open line} = -jZ_0 \cot\left(\frac{2\pi l}{\lambda}\right)$$

d. Characteristic of Particular Lengths of Line (Losses Neglected)

Line Length	Far-End Termination	Input Impedance	Equivalent Lumped Circuit
$l \ll \lambda$	0	$j\omega l L_0$	
$l \ll \lambda$	$R = Z_0$	Z_0	
$l \ll \lambda$	∞	$1/(j\omega l C_0)$	
$l = (2n-1)\lambda/8$	0	jZ_0	
$l = (2n-1)\lambda/8$	$R = Z_0$	Z_0	
$l = (2n-1)\lambda/8$	∞	$-jZ_0$	
$l = (2n-1)\lambda/4$	0	∞	
$l = (2n-1)\lambda/4$	Z_t	Z_0^2/Z_t	
$l = (2n-1)\lambda/4$	∞	0	
$l = n\lambda/2$	0	0	
$l = n\lambda/2$	Z_t	Z_t	
$l = n\lambda/2$	∞	∞	

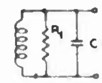
2a—Equivalent Circuits of Quarter-Wave Short-Circuited Lines with Negligible Line Loss.



$$C = \frac{C_0 l}{2}$$

Resonance requires $l = \frac{\lambda_0}{4}$

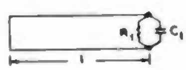
Q = That of line alone* (assumed very high)

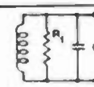


$$C = \frac{C_0 l}{2}$$

Resonance requires $l = \frac{\lambda_0}{4}$

$$Q = \omega \frac{C_0 l}{2} R_t = \frac{\pi}{4} \frac{R_t}{Z_0}$$





$$C = \frac{1}{2}\left[C_t + \frac{C_0 l}{\sin^2\theta}\right]$$

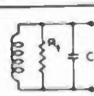
$$\approx C_t + 0.36\, C_0 l \quad (\text{for } \theta < 60 \text{ degrees})$$

Resonance requires

$$l = \frac{\lambda_0}{2\pi}\theta = \frac{\lambda_0}{2\pi}\sin^{-1}\sqrt{\frac{1}{1 + (\omega Z_0 C_t)^2}}$$

$$\sin^2\theta = \frac{1}{1 + (\omega Z_0 C_t)^2}$$

$$Q = \frac{\omega R_t}{2}\left[C_t + \frac{C_0 l}{\sin^2\theta}\right] = \frac{1}{2}\left[\omega C_t R_t + \frac{\theta}{\sin^2\theta}\frac{R_t}{Z_0}\right]$$



Resonance requires

$$l_1 = \frac{\lambda_0}{2\pi}\theta_1, \qquad l_2 = \frac{\lambda_0}{2\pi}\theta_2$$

choosing l_1 or θ_1 then

$$\cos^2\theta_2 = \frac{1}{1 + (\cot\theta_1 - \omega C_t Z_0)^2}$$

$$C = \frac{1}{2}\left[C_t + \frac{C_0 l_1}{\sin^2\theta_1} + \frac{C_0 l_2}{\cos^2\theta_2}\right]$$

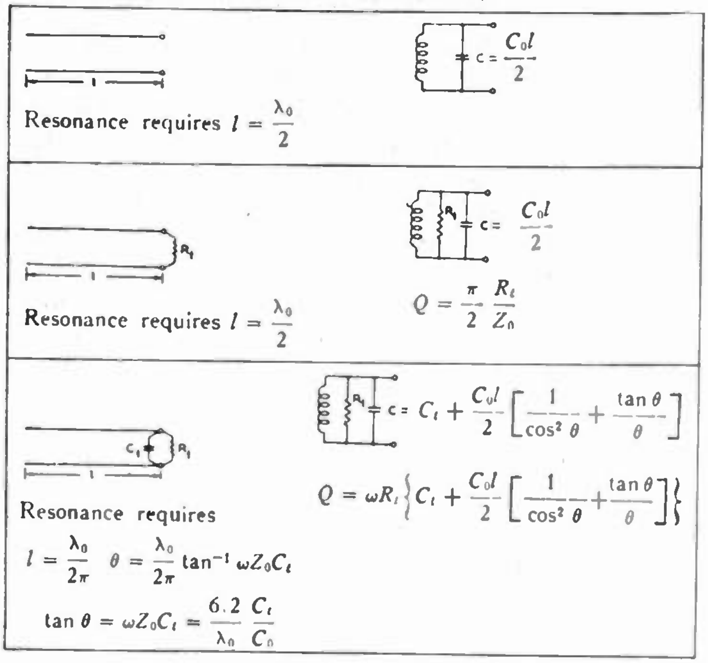
$$Q = \frac{\omega R_t}{2}\left[C_t + \frac{C_0 l_1}{\sin^2\theta_1} + \frac{C_0 l_2}{\cos^2\theta_2}\right]$$

Useful relations: If C_0 is in micromicrofarads per centimeter, C_t in micromicrofarads, λ in centimeter,

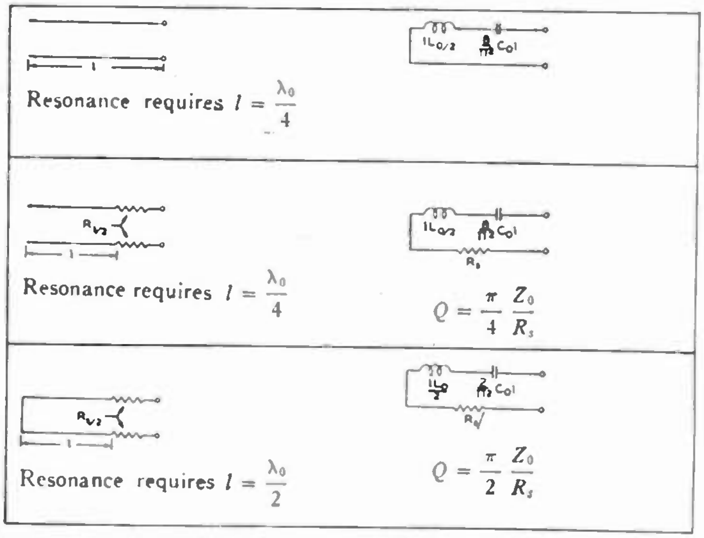
$$C_0 = \frac{33}{Z_0} \qquad Z_0 = \frac{33}{C_0} \qquad \omega Z_0 C_t = \frac{6.2\, C_t}{\lambda_0\, C_0}$$

* The maximum Q of concentric-cylinder, copper, $\lambda/4$ line is given when $Z_0 = 77$ ohms and is $Q_{\text{line}} = 5.2 \times 10^4 (a/\sqrt{\lambda_0})$, where a is radius of inner conductor in centimeters and λ_0 is also in centimeters.

2b—Equivalent Circuits of Open Lines in Half-Wave Resonance.



2c—Lines in Series Resonance.



equivalent lumped circuits are illuminating to the engineer familiar with low-frequency technique, and, in addition, greatly simplify circuit analysis.

The following compilation of transmission-line information is, therefore, based on the assumption that the losses may be lumped at one point. In the figures, although parallel-wire lines are schematically indicated, the results are, of course, applicable to coaxial or to any other uniform line configuration.

The general method which was used to derive the equivalent, *parallel*, lumped circuits of Parts 2a and 2b of the Appendix made use of the voltage distribution $V(x)$ as a function of the distance x along the resonant line. This distribution, which is composed of appropriate portions of a sinusoid, was then used to calculate the total stored energy by integrating the product $\frac{1}{2}C_0V^2$ over the total length, where C_0 is the capacitance per unit length. The stored energy was equated to the stored energy of an equivalent lumped circuit having the same voltage as the terminals of the line circuit. Thus the effective, equivalent, lumped capacitance (excluding the externally added capacitance) was just

$$C = \frac{C_0}{V_t^2}\int_0^l V^2(x)dx$$

where V_t is the voltage at the terminals and l is the length. For example, in the first illustration of Part 2a below, $l=\lambda/4$ and $V(x) = V_t \sin 2\pi x/\lambda$ so that

$$C = C_0\int_0^{\lambda/4}\sin^2\frac{2\pi x}{\lambda}dx$$
$$= C_0\frac{\lambda}{8}$$
$$= C_0\frac{l}{2}.$$

In some of the more complicated cases the result can be expressed in many different forms by trigonometric manipulation.

Equivalent, lumped, *series* circuits, such as those in part 2c of the Appendix, were most easily found by using the *current* distribution on the resonant line and integrating the total stored magnetic energy, again equating the total to the energy of a lumped circuit having the same current as the terminals of the resonant line.

There is here presented to the readers of the PROCEEDINGS of the I.R.E. a chapter of the forthcoming book on "Hyper and Ultra-High Frequency Engineering" by Drs. Robert I. Sarbacher and William A. Edson. Acknowledgment is expressed for the permission to publish this material which has been courteously granted by the authors and by the publishers, John Wiley and Sons Company of New York, N. Y. It is believed that the following material will be found to be of timely interest and instructional value.

The Editor

Tubes Employing Velocity Modulation*

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Summary—The fundamental principles involved in the operation of velocity-modulation tubes are discussed with particular emphasis upon physical concepts. The relation of these devices to familiar low-frequency tubes is examined.

The operation of several devices for the production of velocity modulation is considered, and several methods of converting a velocity-modulated beam into an intensity-modulated beam are explained.

The behavior of the Klystron amplifier is considered on a kinematic basis, and is illustrated by means of the Applegate diagram.

Phase shift in the Klystron is discussed in terms of electron transit time, the mechanism of bunching and catching, and the orientation of the coupling loops.

The Klystron oscillator is considered in terms of general oscillator theory. The values of accelerator voltage required are developed in terms of the feedback coupling and the dimensions of the tube.

The operation of the cavity resonators is explained in a qualitative manner. (A more detailed discussion of cavity resonators appears in an earlier chapter of the book.[1])

The inductive output amplifier is discussed and its operation is compared with that of true velocity-modulation tubes.

I. Introduction

IT HAS been shown[2] how the transit time of the electron in the interelectrode space affects and limits the performance of conventional vacuum-tube structures at high frequencies. In particular, we have found that the power required to drive the control grid of ordinary tubes becomes comparable to the output power and that amplification is, therefore, impossible. We shall now describe several devices in which these difficulties are more or less completely overcome.

Although the transconductance of ordinary vacuum tubes becomes complex, and its magnitude decreases at very high frequencies, this effect seldom proves to be the real limit of operation. At considerably lower frequencies the effective conductance of the grid circuit becomes so high that it is impracticable to operate the tube. The power supplied at radio frequencies by the grid circuit serves to increase the average velocity of the electrons which reach the plate and so is lost as heat at the plate.

In the ordinary vacuum tubes the grid is successful in controlling the plate current only if a considerable space charge exists in the grid-cathode region. The variation of electronic current must, therefore, exist in the entire cathode-anode space. It was shown[2] that the cathode-grid current must properly be considered to be the sum or difference of a grid-cathode current and a grid-anode current. At low frequencies in negative-grid tubes these two currents are identical and the effective grid impedance is infinite. At high frequencies, however, the two currents are out of phase with each other and with the grid voltage. For this reason both a grid conductance and a grid susceptance are developed. The losses that result may be reduced by reduction of the transit time since they vary as the square of this time, but this process has rather definite limitations. A more profitable approach is through utilization of velocity modulation which is a by-product of our efforts to modulate the density or strength of the electron current. It has been found possible to produce satisfactory velocity modulation[3] in practical vacuum tubes, without reducing the effective grid impedance below 50,000 ohms, even at frequencies of 5000 megacycles. Several methods of producing and utilizing this type of modulation will now be described.

II. Velocity Modulation

The operation of velocity-modulated tubes is probably no more complicated than that of some forms of low-frequency tubes. However, the principles involved in velocity-modulation tubes are relatively unfamiliar to most engineers and we shall consider them in some detail. Just as the operation of triodes is greatly clarified by a separate consideration of the direct and alternating components of voltage, current, etc., so is the operation of the velocity-modulated tubes clarified in the same way. The most characteristic property of

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[1] Robert I. Sarbacher and William A. Edson, "Hyper and Ultra-High Frequency Engineering," John Wiley and Sons, New York, N. Y., 1943.

[2] In Chapter 12 of "Hyper and Ultra-High Frequency Engineering" and in numerous other references.

[3] The term "to modulate" as used in communication engineering means *to vary*. Thus velocity modulation means velocity variation.

velocity-modulated tubes is that the current leaving the cathode is constant. Space charge is, therefore, unnecessary to the successful operation of such devices, though it usually exists for reasons of cathode design.

Let us consider for a moment the motion of electrons at some small region in a conventional vacuum tube. Provided that all the electrons in this region have the same velocity at a given instant we can resolve the velocity into two components, one constant, the other alternating. The steady-velocity component is due to some direct accelerating voltage; the alternating component is due to some alternating or modulating voltage. Such a condition exists in the region of the electrode system which produces the velocity modulation.

At other points in the tube the number of electrons passing in a given unit of time may vary and the electrons may have different velocities. Such a condition exists if a velocity-modulated stream of electrons is allowed to "drift" for an appreciable distance. Here it may be more appropriate to consider only the variation in number of electrons with respect to time, neglecting the effect of velocity.

Modulation of the current density as carried out in ordinary tubes is readily expressed in terms of the total current since it is the same at all points in the system. The degree of velocity modulation, however, is not so simply expressed, for, although it may be expressed in percentage of the steady velocity at a given point, a different result obtains if the observation is made at a point where the steady velocity is different. It is commonly expedient to express the velocity modulation directly in terms of volts, a figure which is constant at all points in the tube which follow the modulating electrodes.

III. Production of Velocity Modulation

Two adjacent grid structures form the simplest device for producing velocity modulation. Such a device in conjunction with other electrodes is shown in Fig. 1.

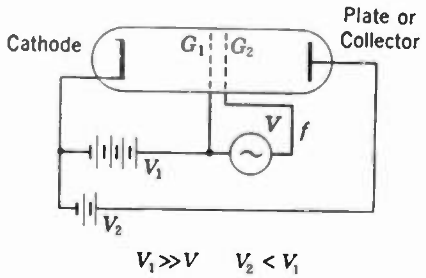


Fig. 1—Structure for producing velocity modulation. The maximum value of the alternating voltage of frequency f is represented by V.

Let us assume that the direct accelerating voltage V_1 is large in comparison to the alternating voltage V. Also let the velocity of the electrons reaching the first grid G_1 be high enough, owing to the action of V_1, so that the interval of time spent in traveling between the grids is short compared to the period of the applied alternating-voltage wave. Under these conditions the electrons which pass through the second grid have a velocity which depends upon the instantaneous sum of V_1 and V. All the electrons which pass through the grid structure continue on to the plate or collector. Since V_2 is lower than V_1 the electrons are decelerated before reaching the plate and dissipate only a relatively small amount of power there.

The electric field in the space between the cathode and G_1 is fixed because V_1 is constant, and accordingly the velocity and number of electrons which reach G_1 per unit time are fixed.[4] Since $V_1 \gg V$ and since the transit time between the grids is short it follows that the number emerging from G_2 per second is also virtually constant. This statement is of great importance since it permits us to estimate the power required of the generator V. That is, we may deduce the effective input grid conductance in this way.

It has been shown[2] that the effective current density due to the motion of the electrons between G_1 and G_2 is equal to ρv, where ρ is the volume density of electronic charge and v is the electron velocity. During the time that V is positive, therefore, the electron velocity is in the same direction as the electric field and energy is drawn from the source V. During the next half cycle the voltage is opposite to the electron velocity and an equal amount of energy is returned to the source. The situation is equivalent to a low-frequency system in which direct current flows through an alternator. The power flow is reversed with each half cycle, and the net power integrated over a cycle is zero.

Actually the input conductance is not identically zero as indicated above. An analysis too lengthy to include here is necessary in order to obtain the exact result. Such an analysis shows that the conductance, although finite, is ordinarily too small to be of practical importance. The resonators or other circuits which supply the excitation V ordinarily have inherent losses large in comparison to this active loss.

Thus it is seen that velocity modulation may be produced by a pair of grids similar to those employed at low frequencies. The success of the grids in producing velocity modulation is independent of space charge near the cathode so that any form of cathode may be used. The power required for the production of velocity modulation is intrinsically low and decreases as the accelerating voltage V_1 is increased and as the spacing between the grids is decreased.

IV. Velocity Modulation Produced in Two Steps

The structure of Fig. 2 may be used for the production of velocity modulation. It is twice as effective as that of Fig. 1 in that the same voltage acts twice upon each electron. Let us first consider the action of this device when the alternating voltage is zero. Electrons which leave the cathode are accelerated to a relatively high velocity as they approach G_1. In the space between

[4] Except for small variations due to the random character of emission.

G_1 and the sleeve electrode they are somewhat decelerated so that they travel the length of the sleeve at the uniform velocity corresponding to V_3. In the region between the sleeve and G_2 they are again accelerated to the velocity corresponding to V_1, and finally they are decelerated toward the plate to a velocity corresponding to V_2.

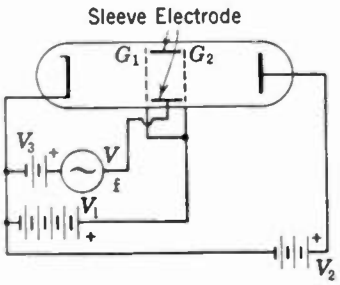


Fig. 2—Alternative structure for producing velocity modulation. The maximum value of the alternating voltage of frequency f is represented by V.

By an appropriate choice of V_1 and V_3 and the dimensions, the time required for an electron to travel from G_1 to G_2 can be made equal to that of a half cycle of a given alternating voltage V. If this voltage is applied as is indicated in Fig. 2, the same electrons which experience a minimum of deceleration in passing from G_1 to the sleeve experience a maximum of re-acceleration between the sleeve and G_2 and thus leave G_2 with a velocity corresponding to the voltage V_1+2V. Electrons which come through the system a half cycle later experience a maximum deceleration and a minimum re-acceleration and leave G_2 with a velocity corresponding to V_1-2V. Electrons arriving at intermediate times experience intermediate values of acceleration or deceleration.

Since $V \ll V_1$ it is still true that the stream of electrons leaving the second grid G_2 is practically uniform. Actually, of course, the electrons which are accelerated at the first grid-sleeve transit tend to come out of the second grid a little ahead of their normal time. Similarly, electrons which are retarded at the first grid-sleeve transit tend to come out of the second grid somewhat behind normal. So far as these effects are small the power required to produce velocity modulation in this device also is negligible.

V. Utilization of Velocity Modulation

The electrode arrangements so far described have been useless in the sense that no output signal is derived from the electron beam even though velocity modulation has been achieved. In order to be of practical use, the velocity-modulated beam must first be converted into an intensity-modulated beam.[6] There are at least three ways to accomplish this:
(1) by deflection method; (2) by the use of a retarding field; and (3) by the drift tube.

[6] Intensity-modulated beams are those in which the convection current ρv at a given point is a function of time.

Conversion by Deflection

If a beam of electrons is deflected by means of a transverse field, either electric or magnetic, the path described depends upon the velocity of the electrons. Accordingly, by the use of an appropriate field it is possible to separate a velocity-modulated beam into two beams which are modulated in intensity. The two beam currents are necessarily out of time phase by 180 degrees so that a natural push-pull system results. One possible arrangement for utilizing this effect is shown in Fig. 3. Electrons which leave the velocity-modulating grids with maximum velocity are least deflected and reach the further anode P_1. Electrons with minimum velocity are most deflected and strike P_2.

Actually the arrangement shown in Fig. 3 is not particularly practicable. The voltages must be controlled rather accurately, and the operation is quite sensitive to stray electric or magnetic fields. Moreover, it is fundamentally nonlinear, since a finite division between the plates P_1 and P_2 must exist. Accordingly, very small alternating voltages at V may be unable to excite any finite output in the tank circuit. At high levels of input the beam is completely separated and no further increase of output is possible.

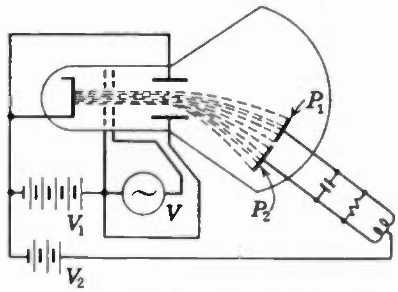


Fig. 3—Tube employing fixed electric field to separate velocity-modulated beam into two intensity-modulated beams.

Conversion by Retarding Field

If in Fig. 1 the auxiliary voltage V_2 is made equal to zero we find that electrons which leave the grids with velocity above normal reach the collector or plate but that electrons which leave the grid with lower velocity are turned back and ultimately return toward the grid or other positive structure. Thus a true conduction current exists in the plate lead of the tube. Fairly efficient operation is possible with such a tube provided that suitable steps are taken to prevent the return of electrons to the exciting grids. This precaution is necessary because the distance traveled to and from the plates permits the faster electrons to overtake the slower ones, producing an intensity-modulated beam. Such a beam causes the input impedance to have a large resistive component which may be either positive or negative. If the resistance is positive a large driving power may be required. If it is negative an undesired mode of oscillation may occur.

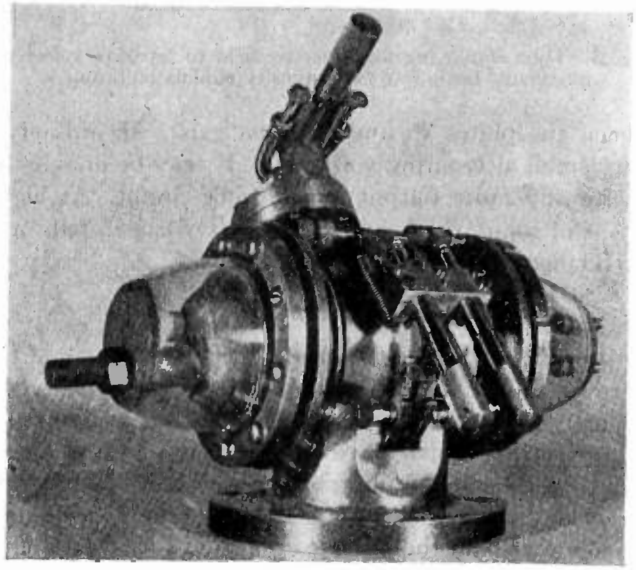
A rather different process takes place if the voltage of V_2 in Fig. 1 is made sufficiently negative so that no electrons reach the plate. Now we find that the faster electrons approach the plate very closely while the slower ones are turned back somewhat farther away. The charge induced in the plate circuit by these moving electrons is thus controlled by the potential of the velocity-modulating electrodes. Again, suitable precautions must be taken if the electrons are not to return to the grid with undesirable consequences.

Successful tubes have been constructed upon this principle, and reasonably satisfactory results are obtained. It appears, however, that certain other designs are fundamentally better, and we shall therefore give this one only brief consideration.

Conversion by Drift

An electron beam which is velocity-modulated will convert itself into one which is intensity-modulated if given sufficient time. That is, a uniform velocity-modulated beam will automatically gather itself into clumps if allowed to drift freely in an equipotential region. Such a region is known as a drift space, and the resulting clumped or bunched beams may be partly or completely intensity modulated.

The most familiar device for utilization of this principle is the Klystron.[6] Although a number of other workers, notably Hahn and Metcalf of the General



(R. H. and S. F. Varian, courtesy of Jour. Appl. Phys.)
Fig. 4—Experimental model of the Klystron.

Electric Company, produced similar devices at about the same time, the brothers S. F. and R. H. Varian are generally credited with its development.[7] Fig. 4 is a photograph of an early experimental model (1939). Fig. 5 shows the internal arrangement of two typical Klystrons suitable for amplification, detection, or oscillation. The cathode is plane and relatively large. The grid indicated is actually a beam-forming electrode which operates at a small positive bias and serves to control the total cathode current. The principal structure is of copper and comprises two cavity resonators coupled mechanically and at the same direct potential. Each cavity resonator includes two gridlike structures close together and operating in much the same way as the double grid of Fig. 1.

In operation the electrons emitted by the cathode are formed into a relatively narrow circular beam by the grid and are accelerated toward the metallic structure. Some of them strike the grid mesh, but a majority pass through the first cavity resonator into the equipotential drift space. They proceed through this space with essentially uniform velocity to the grids of the

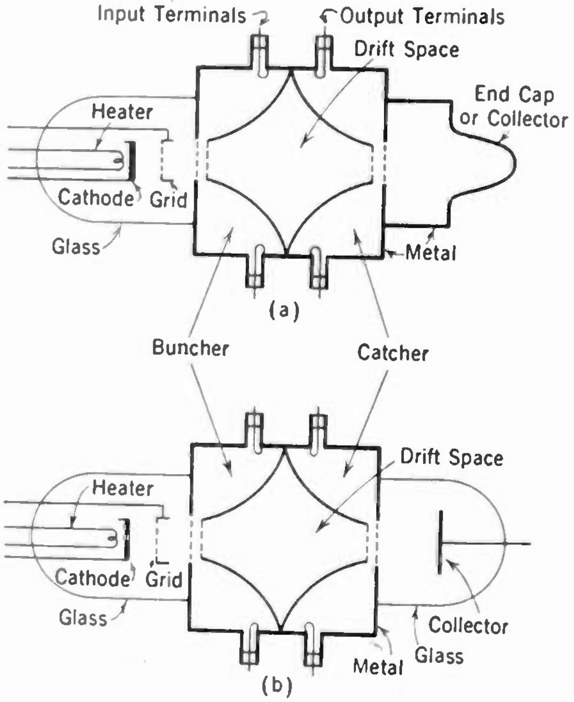


Fig. 5—Schematic representations of two modern Klystrons.

second resonator; again a portion of the electrons strike the mesh but a majority pass through and are decelerated, finally being captured by the plate or collector. By applying a moderate potential to the plate of Fig. 5(b), it is possible to capture most of the electrons without the dissipation of a large amount of power as heat.

The situation is appreciably altered if some alternating voltage exists between the first two grids (the buncher) due to oscillations fed into the resonator. Now the electrons which pass into the equipotential or drift space are velocity-modulated. Electrons of higher velocity may overtake slower electrons which preceded them in time. That is, the electrons which arrive at the second pair of grids (the catcher) are in clumps or bunches. The development of this condition is shown in Fig. 6. To simplify the picture it is assumed

[6] The word Klystron is a trademark owned by the Sperry Gyroscope Company. The term is derived from Greek and means "waves breaking on a beach."

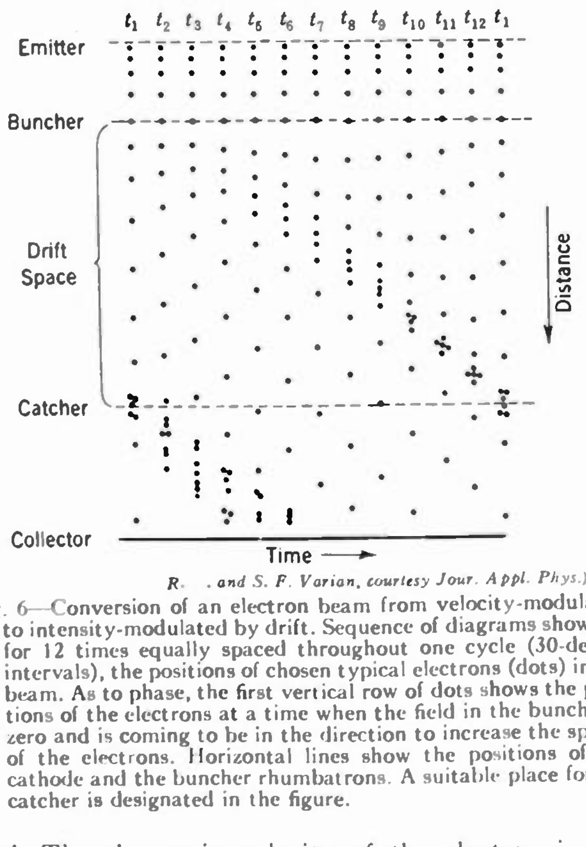
[7] R. H. and S. F. Varian, "A high-frequency oscillator and amplifier," *Jour. Appl. Phys.*, vol. 10, pp. 321–327; May, 1939.

that the space between the two grids of the buncher is negligible. Also, it is assumed that the voltage fed to the buncher is of such a value as to produce maximum intensity modulation of the beam at the catcher, an optimum condition.

VI. The Applegate Diagram

Further clarification of the action of the buncher may be obtained from a study of the Applegate diagram, Fig. 7. The straight lines in this diagram represent the positions of individual electrons after they leave the buncher, as a function of time. The slope of these lines is made proportional to the electron velocities, i.e., proportional to the alternating and direct components of the buncher voltages. In drawing this diagram, reference is made to Fig. 8 and the following assumptions are made:

1. All electrons have the same velocity before passing through the buncher.
2. Electrons pass through buncher at equal intervals.
3. The velocity $v_1 \sin \omega t_1$ due to the alternating voltage component $V_1 \sin \omega t_1$ is small compared to the velocity v_0 due to the voltage V_0.



(R. . and S. F. Varian, courtesy Jour. Appl. Phys.)

Fig. 6—Conversion of an electron beam from velocity-modulated to intensity-modulated by drift. Sequence of diagrams showing, for 12 times equally spaced throughout one cycle (30-degree intervals), the positions of chosen typical electrons (dots) in the beam. As to phase, the first vertical row of dots shows the positions of the electrons at a time when the field in the buncher is zero and is coming to be in the direction to increase the speeds of the electrons. Horizontal lines show the positions of the cathode and the buncher rhumbatrons. A suitable place for the catcher is designated in the figure.

4. The change in velocity of the electron in the buncher is $v_1 \sin \omega t_1$.

All these conditions are met to a fair degree in the practical Klystron. The applied direct voltage V_0 is in all cases sufficiently large in comparison to V_1 so that v_0 is always much greater than v_1. If only a small number of electrons strike the grid mesh of the buncher, it is evident that practically as many electrons leave the buncher as come to it. Further, the currents induced in the buncher by the electrons approaching and leaving it are of opposite sign. Hence the current in the external circuit of the buncher is

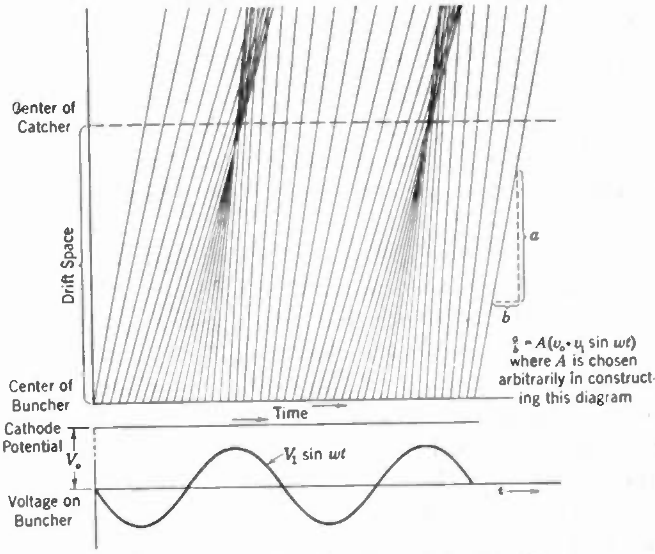
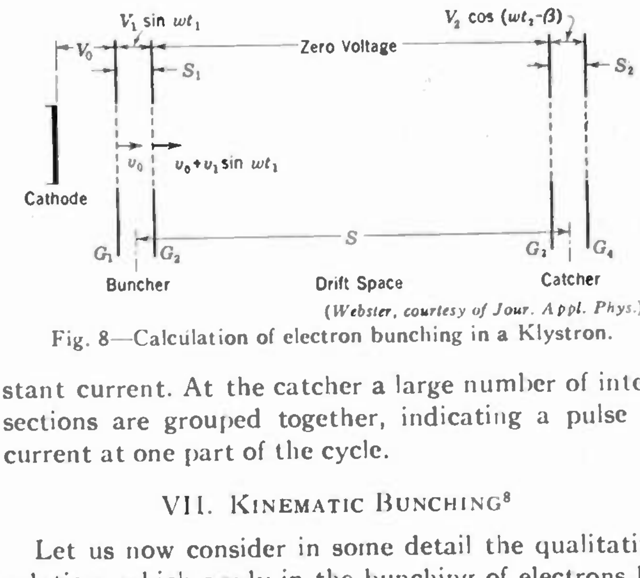


Fig. 7—The Applegate diagram of the Klystron.

quite small, and power losses in the buncher are likewise small.

The variation of current with respect to time at any particular distance from the buncher is represented in Fig. 7 by the intersections of the sloping lines with a horizontal line at that location. At the buncher the intersections are uniform, indicating a uniform or con-



(Webster, courtesy of Jour. Appl. Phys.)

Fig. 8—Calculation of electron bunching in a Klystron.

stant current. At the catcher a large number of intersections are grouped together, indicating a pulse of current at one part of the cycle.

VII. Kinematic Bunching[8]

Let us now consider in some detail the qualitative relations which apply in the bunching of electrons by the first resonator of a Klystron. This treatment is approximate in that the debunching caused by space

[8] This treatment follows closely the work of D. L. Webster, "Cathode ray bunching," *Jour. Appl. Phys.*, vol. 10, pp. 501–508; July, 1939; and 864–872; December, 1939. The authors are indebted to Dr. Webster for a very gracious personal letter in which he clarified certain details of the original paper.

The word kinematic is used to signify the fact that the calculation treats the electrons as small solid bodies, projected into the drift space with various velocities. Kinematics is the study of the motions of bodies.

charge is neglected. It is also assumed that all electrons leave the cathode with zero velocity and proceed parallel to the axis of the structure. The grids of the buncher are taken to define true equipotential planes perpendicular to the axis of the tube, and the loss of electrons by collision with the metal of the grid structure is ignored.

The result of such an analysis may be affected by the focusing action of actual grid structures, by loss of electrons at the grids, and by electrons which stray from the beam. However, it is found that the results of this analysis describe the observed behavior to a good approximation.

The diagram of Fig. 8 is convenient for the calculation of kinematic bunching. Velocities are designated by the small v with suitable subscripts. Applied voltages are designated by the capital V, again with appropriate subscripts. To be explicit we shall assume that no alternating potential difference exists between the first grid G_1 and the cathode. Further, no potential difference of any kind is assumed in the drift space between G_2 and G_3. The alternating voltage $V_1 \sin \omega t_1$ is supplied from some external source to excite the buncher. The useful output is derived from the voltage $V_2 \cos (\omega t_2) - \beta$. The constant β is required to account for the time which electrons spend in traversing the drift space. The separate subscripts on the time variable help to simplify the mathematical procedure.

The velocity v_0 with which electrons leave the first grid is defined by the familiar equation

$$\tfrac{1}{2} m v_0^2 = eV_0 \tag{1}$$

where m is the mass and e the charge of the electron. Under our assumptions that S_1 is small compared to the wavelength of the voltages in question and that the electron velocity is small compared to that of light we may state that the number of electrons which leave the second grid is uniform with respect to time and that their velocity is given by the relation

$$v = v_0 + v_1 \sin \omega t_1 \tag{2}$$

where the magnitude of v_1 is determined by a process analogous to that used in (1). We have

$$\tfrac{1}{2} m (v_0 + v_1)^2 = e(V_0 + V_1). \tag{3}$$

Subtracting (1) from (3) and neglecting the term in v_1^2,

$$m v_0 v_1 = eV_1. \tag{4}$$

This equation is taken to define the velocity v_1 which is the maximum value of the velocity contributed to the electrons by the buncher.

Because the acceleration in the buncher and the deceleration in the catcher are essentially linearly distributed it is appropriate to compute the transit time as if the velocity changes were discontinuous at the centers of these spaces. This approximation is consistent with Fig. 8, where S is drawn from the center of the buncher to the center of the catcher. Accordingly we modify our previous statements slightly so that the velocity at the middle of the buncher space S_1 is given by

$$v = v_0 + v_1 \sin \omega t \tag{5}$$

where v_1 and v_0 are determined by (4) and (1), respectively.

VIII. Current Relations in the Klystron

Let us consider an electron which passes the center of the buncher at a time t_1 and arrives at the center of the catcher at some later time t_2. This time is expressed by the relation

$$t_2 = t_1 + \frac{S}{v_0 + v_1 \sin \omega t_1} \tag{6}$$

where $S/(v_0 + v_1 \sin \omega t_1)$ is thet ime required for the electron to travel the distance S when its velocity is $v_0 + v_1 \sin \omega t_1$. If we introduce the approximation

$$\frac{1}{1+a} \simeq 1 - a \quad \text{if} \quad a \ll 1 \tag{7}$$

we have

$$t_2 = t_1 + \frac{S}{v_0}\left\{\frac{1}{1 + \dfrac{v_1}{v_0}\sin \omega t_1}\right\} \simeq t_1 + \frac{S}{v_0} - \frac{S v_1}{v_0^2} \sin \omega t_1. \tag{8}$$

Differentiating, we have

$$\frac{dt_2}{dt_1} = 1 - \frac{S v_1}{v_0^2}\,\omega \cos \omega t_1. \tag{9}$$

The current represented by the modulated beam as it reaches the catcher may be deduced from the continuity of the electron stream. If we designate i_0, i_1, and i_2 as currents at the input of the buncher, middle of buncher, and middle of catcher, respectively, we have $i_0 \simeq i_1$ at all times. The continuity relation thus becomes

$$i_1(t_1)\,dt_1 = i_2(t_2)\,dt_2 \tag{10}$$

where the first term represents i_1 evaluated at t_1 multiplied by the incremental time dt_1. The second term represents i_2 evaluated at t_2 multiplied by the incremental time dt_2. Replacing i_1 by i_0, which is independent of time, we have, upon substitution of (9)

$$i_0 = i_2(t_2)\frac{dt_2}{dt_1} = i_2(t_2)\left\{1 - \frac{S v_1}{v_0^2}\,\omega \cos \omega t_1\right\} \tag{11}$$

or

$$i_2(t_2) = \frac{i_0}{1 - \dfrac{S v_1 \omega}{v_0^2} \cos \omega t_1}. \tag{12}$$

Equation (11) defines the current at the catcher i_2 as a function of the time electrons pass through the buncher. If $sv_1\omega/v_0^2$ is small in comparison to 1 we may reintroduce the approximation of (7) to write

$$i_2(t_2) = i_0\left(1 + \frac{Sv_1\omega}{v_0^2}\cos\omega t_1\right) \quad \text{if} \quad \frac{Sv_1\omega}{v_0^2} \ll 1 \tag{13}$$

which is a sinusoidal function.

Equation (12) requires careful consideration. At first glance it appears that the current is negative over part of each cycle if $Sv_1\omega/v_0^2 > 1$. Such a negative current is contrary to fact, because the direction of the motion of electrons through the catcher never reverses. The apparent discrepancy is clarified by reference to Figs. 7 or 6. It is seen that fast electrons are able to overtake slower ones which preceded them before reaching the catcher. The wave of current which results when i_2 is plotted with respect to time is thus doubly peaked in each cycle.

The situation is further clarified by reference to (9) which shows that dt_2/dt_1 is also negative over part of each cycle for the conditions in question. A sort of folding back or overlapping process takes place so that electrons which left the buncher over three separate intervals of dt_1 pass through the catcher in the same time interval.

We are therefore correct in using the magnitude of current represented by (12) but in taking it as always positive. If $Sv_1\omega/v_0^2 = 1$, the current value indicated is infinite once per cycle. For $Sv_1\omega/v_0^2 > 1$, the current value indicated is infinite twice during each cycle. In the physical tube, of course, these peaks of current are lowered and spread out in time by the action of space charge. The peaks are relatively sharp and the presence of the double peak is particularly interesting in that it suggests efficient operation as a high-order frequency multiplier.

IX. Phase Shift in the Klystron

The Klystron differs from all conventional types of tubes in that it operates with an intrinsically large amount of phase shift. This phase shift is fundamental to the operation of the tube itself and, therefore, may not be removed by any ordinary correction method. For most purposes it is not objectionable, but a number of unusual phenomena result from its presence.

The phase shift with which we are dealing is caused by a time delay and is most readily calculated on that basis. Electrons require an appreciable time to travel from the buncher to the catcher. A signal which is suddenly applied to the buncher is thus able to affect the catcher only when the electrons controlled by the signal arrive at the catcher.

The time required for electrons to travel from buncher to catcher is readily calculated because the voltage applied to the buncher grid is usually small compared to the applied direct potential. Accordingly the velocity with which electrons travel the drift space may be taken as

$$v_0 = \sqrt{\frac{2eV_0}{m}} \tag{14}$$

from (1). In practical units, (14) becomes

$$v_0 = 6.0 \times 10^7\sqrt{V_0} \text{ centimeters per second.} \tag{15}$$

The time required for electrons to cross the drift space is, in terms of Fig. 8,

$$t = \frac{S}{v_0} = \frac{S}{\sqrt{V_0}}\cdot\frac{1}{6} \times 10^{-7} \text{ second} \tag{16}$$

where t is the time in seconds, S is the distance in centimeters, and V_0 is the applied potential in volts.

For a typical commercial Klystron, $S=3$ centimeters. Accordingly

$$t = \frac{1}{\sqrt{V_0}}\cdot\frac{3}{6} \times 10^{-7} = \frac{5}{\sqrt{V_0}} 10^{-8} \text{ second.} \tag{17}$$

The time delay just calculated is readily converted to a phase shift θ by means of the formula

$$\theta = \omega t. \tag{18}$$

For a typical Klystron such as we are considering the operating frequency may be taken as $f = 3 \times 10^9$ and $\omega = 6\pi \times 10^9$. Accordingly we write

$$\theta = \frac{300\pi}{\sqrt{V_0}} \text{ radians.} \tag{19}$$

For an operating voltage of 900 volts, a reasonable value, $\sqrt{V_0} = 30$ and

$$\theta = 10\pi \text{ radians.} \tag{20}$$

This is recognized as 5 complete cycles or 1800 degrees.

An additional phase shift of 90 degrees may be explained in terms of the mechanism of bunching. Let us consider three electrons that pass through the buncher at three successive instants of time corresponding to minimum, zero, and maximum acceleration, respectively. The electron which passed through the buncher at zero alternating voltage will arrive at the catcher at the same instant as the decelerated electron which preceded it and the accelerated electron which followed it. That is, the current peak at the catcher is associated with zero field in the buncher. Since the voltage maximum of the catcher must coincide with the current peak there for best operation, we conclude that a 90-degree phase shift between the two resonators must exist in addition to the phase shift of delay.

When the Klystron serves as an amplifier the phase shift is relatively unimportant. The output signal is delayed a small fraction of a microsecond with respect to the input, but no other effect is observed. Even if negative feedback is to be applied the situation is not serious because the phase shift is largely a function of the applied direct voltage rather than of frequency.

X. The Klystron as an Oscillator

The Klystron oscillator is in some ways considerably more complex than most low-frequency oscillators. A typical circuit is shown in Fig. 9. Its operation is best analyzed in terms of the functional block diagram of

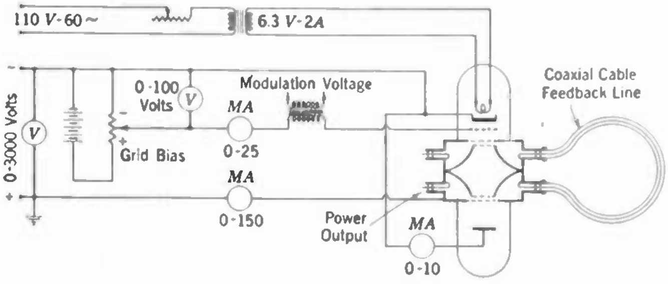


Fig. 9—Circuit diagram for Klystron oscillator with plate collector terminal brought out.

Fig. 10. This diagram is quite general and is useful in the study of all feedback oscillators. The amplification is provided by the electron beam as previously discussed. The limited action takes place as a result of the mechanism of bunching. When the voltage at the buncher exceeds a certain value the current at the catcher forms a double rather than single peak, and the effective transconductance is decreased.[9]

The frequency-control mechanism is complicated by the fact that there are two separate resonators and that they are coupled relatively tightly together. The behavior is best explained in terms of low-frequency coupled-circuit theory. It will be recalled that the

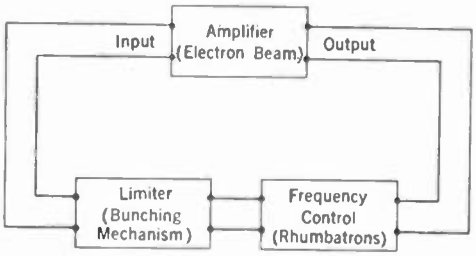


Fig. 10—Functional block diagram of a Klystron oscillator.

transmission characteristic of two identical tuned circuits shows a double hump provided that the coupling exceeds a certain value known as the critical value. (See Fig. 11.) Associated with this transmission characteristic is a curve of phase shift which is zero at three different frequencies. Oscillations are readily produced at the two frequencies of peak transmission. The reduced transmission associated with the mid-point is relatively unfavorable to oscillation.

In any oscillator it is necessary that the total phase shift around the loop be zero or an integral multiple of 2π at the operating frequency and that the voltage amplification of the entire system be unity. In the ordinary oscillator the phase shift of the system is relatively independent of applied voltages. Accordingly the conditions for oscillation are relatively independent of the applied voltage. The frequency of such an oscillator then adjusts itself until the total loop phase shift is zero.

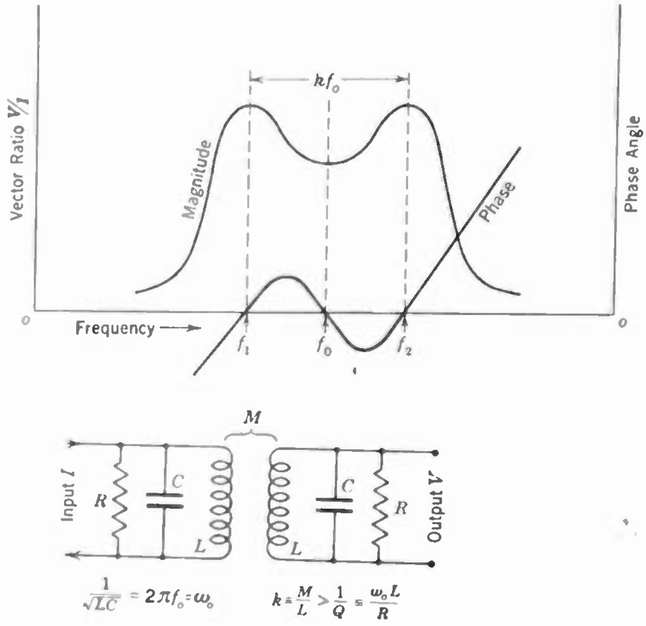


Fig. 11—Properties of identical coupled tuned circuits.

In the Klystron oscillator the total phase shift is the sum of three components. These are

$$\left.\begin{aligned} \theta_1 &= \frac{\pi}{2} \text{ due to inherent buncher-catcher relationship} \\ \theta_2 &= \frac{\omega S}{6\sqrt{V_0}} \times 10^{-7} \text{ due to transit time in drift tube} \\ \theta_3 &= \frac{l\omega}{v_c} \text{ due to feedback cable} \end{aligned}\right\} \quad (21)$$

where ω is the operating angular velocity, S is the drift distance, l is the effective length of the feedback cable, V_0 is the applied direct voltage, and v_c is the velocity of transmission of the feedback cable. Because only one of these terms involves the applied voltage, it is to be expected that only certain values of this voltage

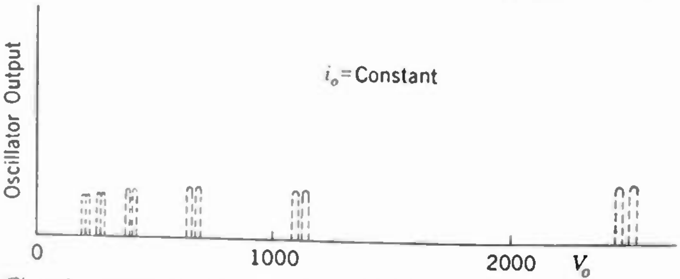


Fig. 12—Oscillation characteristic of a Klystron showing critical operating potentials.

will produce oscillation. Such is actually the case, and it is usually possible to identify two distinct sets of these values. One set of voltages reduces the total net phase shift of the system to a multiple of 2π for the

[9] Strictly speaking, the best operation occurs when the catcher current just begins to show a double peak. The above statement refers to an excessive broadening of the peaks.

frequency f_1 of Fig. 11; the second set reduces the phase shift to a multiple of 2π for the frequency f_2.

A typical plot of oscillation output versus applied voltage is shown in Fig. 12. The oscillations are in general somewhat more powerful as the applied voltage is raised, and the voltage increments between successive points of oscillation are increased.

The above relationships may be put into numerical form. The total phase shift must equal $2n\pi$ where n is an integer.[10] Accordingly we write

$$2n\pi = \frac{\pi}{2} + \frac{\omega S}{6\sqrt{V_0}} \times 10^{-7} + \frac{l\omega}{v_c} \tag{22}$$

or

$$\frac{\omega S}{6\sqrt{V_0}} \times 10^{-7} = 2n\pi - \frac{\pi}{2} - \frac{l\omega}{v_c}. \tag{23}$$

Dividing by 2π we have

$$\frac{fS}{6\sqrt{V_0}} \times 10^{-7} = \left(n - \frac{1}{4} - \frac{lf}{v_c}\right). \tag{24}$$

If the left member of the equation is evaluated for the various voltages at which oscillation occurs and if these values are plotted against a suitable series of integers, a straight line results. In this plot the larger values of

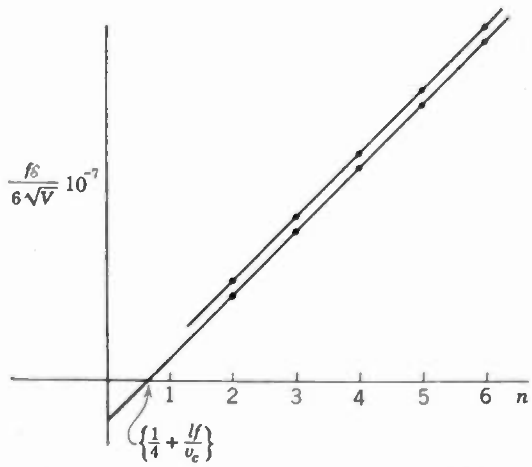


Fig. 13—Critical voltages for oscillation of a Klystron.

voltage are associated with the smaller integers. That two such lines are often observed is evidence that oscillation occurs at two separate frequencies. Such a plot is shown in Fig. 13. It is not always possible to make V_0 large enough to obtain the point $n=1$. The intercept on the horizontal axis is the value $(1/4+lf/v_c)$. Accordingly a means is at hand for evaluating the total phase shift of the circuit which couples the two resonators.

XI. The Cavity Resonator

We have now shown how the process of velocity modulation takes place and how the electrons gather themselves into an intensity-modulated wave of bunches. It remains to show how the output resonator or catcher derives power from the rhythmic passage of these bunches of electrons. From our previous work it is clear that the passage of these clumps of electrons across the space between the two grids is equivalent to the passage of a succession of pulses of electric current. It is necessary only that a potential difference exist between the grids in such a time phase as to oppose the passage of these electrons in order to derive a power output. Such a voltage will automatically result if the second resonator is tuned to the frequency of the input signal.

The situation is remarkably similar to that existing in a class C amplifier at lower frequencies. The grid produces a current in the plate-cathode circuit that flows in periodic short pulses. These pulses act upon the tuned tank circuit to build up a large voltage which opposes the flow of current and so produces the useful power output.

It is apparent that the flexibility of the Klystron is seriously limited by the fact that cavity resonators are permanently attached to the grids and thus form an integral part of the tube itself. By their very nature these resonators are adjustable over only a narrow frequency range, and accordingly the entire tube is limited to this particular narrow band for which the resonators are designed. Ordinary triodes and pentodes, on the other hand, are not so limited and therefore operate over a wide ratio of frequencies as controlled by the external oscillatory circuits or resonators which are attached.

A pair of grids suitable for producing a velocity-modulated beam in a Klystron have an area in the order of 1 square centimeter and a spacing of approximately 1 millimeter. The net capacitance of such a pair is in the order of 1 micromicrofarad (10^{-12} farad). If the wavelength to be produced is approximately 9 centimeters the frequency is 3.3×10^9 cycles per second and the natural angular velocity is 2×10^{10}. Using the basic relation

$$\omega_0^2 = \frac{1}{LC} \tag{25}$$

we have

$$4 \times 10^{20} = \frac{1}{10^{-12}L} \tag{26}$$

or

$$L = \tfrac{1}{4} \times 10^{-8} = 0.0025 \text{ microhenry}. \tag{27}$$

It is immediately evident that such a low value of inductance is not readily achieved and that the simultaneous achievement of a high Q is not possible by ordinary methods. It is rather illuminating to see how one form of cavity resonator results as a logical extension

[10] An ambiguity of π radians may easily arise in this work. In order to avoid such an error it is necessary to know the orientation of the coupling loops within the resonators so as to relate the field across buncher or catcher grids to the current flow in the feedback cable. Because the effective length of the feedback cable is seldom known exactly, the error is not of great importance.

of a low-frequency oscillatory circuit. This process is illustrated in Fig. 14.

The simple circuit of Fig. 14(a) is well known as the Hertzian oscillator and is readily recognized as a lumped capacitance associated with a single-turn inductor. At frequencies up to several hundreds of megacycles such a resonator is quite practical. At higher frequencies, however, the coil is necessarily small and the Q is degraded by radiation loss. The arrangement of Fig. 14(b) is capable of operation at somewhat higher frequencies because the two coils are essentially in parallel and the effective inductance is halved. Also radiation loss is reduced because each coil tends to cancel the field of the other.

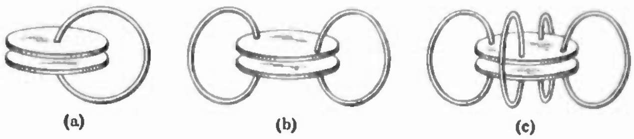


Fig. 14—Development of toroidal cavity resonator from parallel-plate condenser and single-loop coil.

In Fig. 14(c) two more loops of wire are added with an additional improvement in Q and reduction of inductance. The limit which is approached as more and more loops of wire are added is the cavity resonator in the form described by the Varians in their original paper. In such a resonator the electric current may still be thought of as flowing along the direction of the wires of the original structure. The magnetic field produced is confined entirely within the resonator, and accordingly no radiation loss exists. Typical field distributions are shown in Fig. 15. The Q of such systems is therefore limited only by the conductivity of the metal and is readily made quite high. The Q of a resonator without external loading is in the order of 10,000, and the external load is typically adjusted to such a value as to reduce this to about 1000.

The conclusion to be drawn from the foregoing is that the use of the cavity resonator is necessary in order to achieve a satisfactory tuned circuit at the desired frequency. Moreover, the frequency range over which any particular Klystron is operative depends to some extent upon the spacing between the two pairs of grids. If the frequency is relatively low the process of conversion from velocity modulation to intensity modulation requires a length that is excessive. Finally, the use of the cavity resonator is advantageous in that the high values of Q serve to give good frequency stability when the unit is used as an oscillator.

The resonators shown in Fig. 5 differ in form from those just shown, and the effective Q is probably reduced by this alteration. The mechanical structure, however, is improved thereby, and the modification of the electrical performance is not serious. Evidently the basic nature of the resonance as illustrated in Fig. 15 is not changed.

XII. Design and Applications of the Klystron

Since the Klystron is probably the most flexible and the most important of the velocity-modulated tubes, it is well to consider its practical operation in more detail. In particular there is no other device available at the present time which successfully replaces the Klystron as an amplifier of weak signals at hyper frequencies. Let us examine this operation with a consideration of the design features involved.

The past discussion dealt primarily with electron beams in which the velocity modulation was relatively large. Such beams upon drifting only a short distance are converted into beams which are fully intensity-modulated. This condition corresponds to maximum power output and is desirable in the oscillator or power amplifier.

In the low-power amplifier a very small signal is applied to the grids of the buncher and relatively small velocity modulation results. For any moderate length of drift tube such a condition results in a beam having relatively small, nearly sinusoidal, intensity modulation. As the degree of velocity modulation is increased

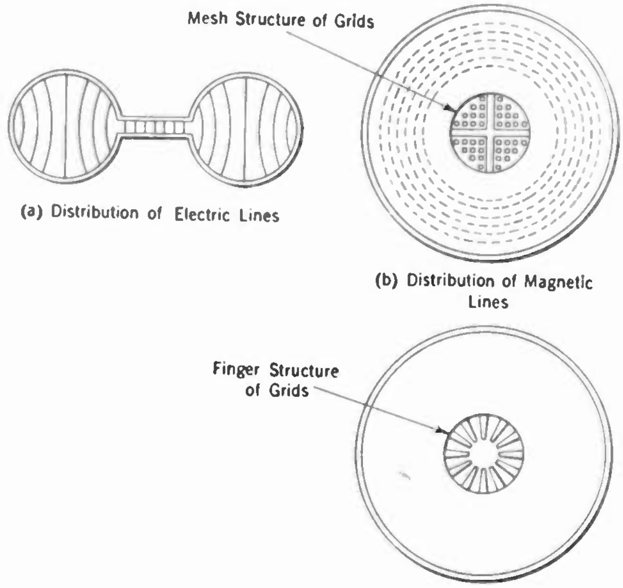


Fig. 15—Sections of a form of cavity resonator showing electric and magnetic lines.

the corresponding intensity modulation becomes less sinusoidal and of greater amplitude. This deviation from a sinusoidal current wave is of great interest since it leads to the possibility of frequency multiplication or of certain forms of modulation. It may be shown that the Klystron as a frequency doubler suffers somewhat less severe drawbacks than the ordinary class C triode doubler at lower frequency. When the Klystron is operating under ordinary conditions as an amplifier or oscillator, the buncher and catcher rhumbatrons are tuned to the same frequency. In frequency-multiplier applications, however, the catcher is tuned to some harmonic of the buncher frequency.

A rigorous mathematical treatment of the action of the Klystron is very difficult since the action becomes nonlinear at relatively small signal strengths. Also space charge and a variety of other effects greatly complicate the problem. Space charge is particularly important in two ways. It tends to spread any form of electron beam because each electron repels those beside it. This action causes electrons to deviate from the main path and to be lost on the inner surface of the drift space. The use of a longitudinal uniform magnetic field causes the electrons to move in helical paths and renders this effect unimportant.

Electrons also exert repulsive forces upon those electrons which precede and follow them in the beam. This action cancels to zero in a uniform beam and so tends to reduce an intensity-modulated beam into a uniform one. In a drift tube, therefore, the tendency of a small velocity modulation to produce an intensity-modulated beam is opposed by the action of space charge. Hahn[11,12] treats this problem in great detail, and his work is recommended to the serious student of velocity-modulated devices.

Two major limitations on the performance of these units are set by the effects of space charge. The intensity modulation produced by a given small signal may not be indefinitely increased by increasing the length of the drift tube since loss by space charge presently exceeds the gain by drift. The current density in the electron beams may not be increased above a certain definite value so that a specific maximum power output is obtainable with a structure of given size.

In the Klystron tubes at present commercially available, the catcher and buncher rhumbatrons are made slightly adjustable by corrugating their side walls. Adjustment screws are provided to expand or contract these walls, thus changing the volume and shape of the cavities. The rhumbatrons are thus exposed and subject to adjustment while operating, and since they are at a high positive potential with respect to the cathode it is customary to ground the positive terminal of the voltage supply. If good frequency stability is desired it is necessary to regulate both the plate and filament voltages. The grid bias may be obtained from a voltage-dropping resistor connected across the plate voltage supply when the operating plate voltages are under 1000 volts. At higher operating voltages it is better to employ a separate source.

Before a Klystron oscillator can be set in operation it is necessary to tune both buncher and catcher to the same frequency. In view of the high selectivity of these elements, this may be quite difficult to do, because the correct operating voltage is not known until the system is in oscillation. The problem may be greatly simplified by inserting an alternating voltage in series with the direct-voltage plate supply. The amplitude of this voltage should be sufficient to insure that the range of variation of the pulsations covers one of the operating potentials. (See Fig. 12.)

XIII. The Reflex Klystron Oscillator

The reflex Klystron oscillator shown in Fig. 16 employs a single rhumbatron which performs the functions of both buncher and catcher. The electrons emerging from the buncher are turned back into the buncher by the plate which is held at a negative potential. When the system is properly adjusted so that the electrons returning to the cavity resonator deliver their energy to it in the correct phase, oscillations are sustained. In many respects the reflex Klystron is similar to the well-known positive-grid oscillator.

XIV. The Inductive Output Amplifier[13,14]

We shall now describe another tube which resembles the Klystron in some respects but in other respects is

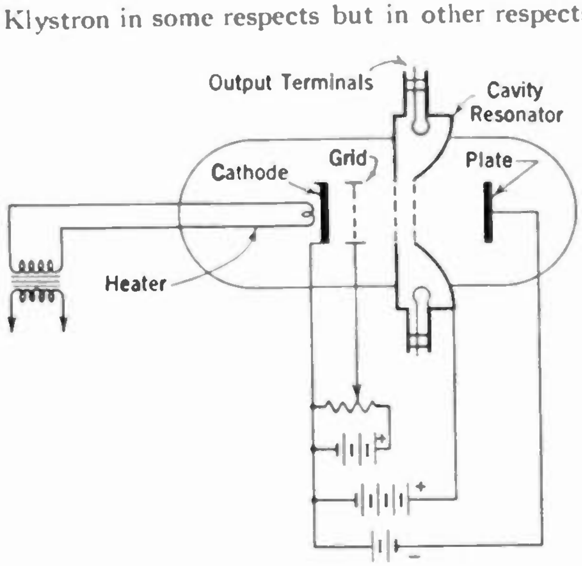


Fig. 16—Reflex Klystron oscillator.

quite different. It appears to promise considerable commercial importance in the band of frequencies between 100 and 1000 megacycles. It is not a velocity-modulated tube in the same sense as the Klystron because the cathode current is controlled directly by a grid of the ordinary type. An electron beam is used, however, and the output is derived from a cavity resonator in a manner very similar to that in the Klystron. Because of the relatively long path used it has been found necessary to provide a certain amount of magnetic focusing to prevent undue spreading of the electron beam.

Fig. 17 shows the basic elements of the system and Fig. 18 illustrates the additional features embodied in the practical tube. The number of electrons which leave the cathode is controlled by the grid. After leaving

[11] W. C. Hahn, "Small signal theory of velocity-modulated electron beams," *Gen. Elec. Rev.*, vol. 42, pp. 258–270; June, 1939.
[12] W. C. Hahn, "Wave energy and transconductance of velocity-modulated electron beams," *Gen. Elec. Rev.*, vol. 42, pp. 497–502; November, 1939.

[13] A. V. Haeff and L. S. Nergaard, "A wide-band inductive-output amplifier," Proc. I.R.E., vol. 28, pp. 126–130; March, 1940.
[14] A. V. Haeff, "An ultra-high frequency power amplifier of novel design," *Electronics*, vol. 12, pp. 30–32; February, 1939.

the grid the electrons are formed into a beam and accelerated by the high voltage applied to the electrodes *c* and *d*. This high-velocity beam which is intensity-modulated by the control grid moves past the aperture or slot in the output resonator. This resonator differs somewhat in form from that of the Klystron

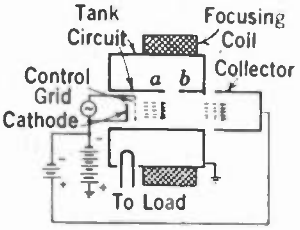


Fig. 17—Essential elements of the inductive output tube.

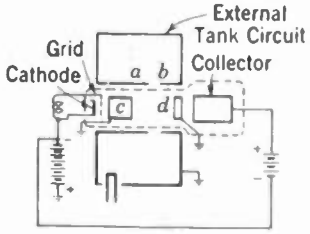


Fig. 18—Experimental form of the inductive output tube.

and in particular lacks the capacitance formed by the parallel-grid structure. Accordingly, it has a higher effective L/C ratio and therefore a higher impedance than that of the Klystron resonator. The mechanism of excitation is the same in both. The potential difference existing across the gap creates a field which retards the electrons, thus absorbing power from them. Electrons which pass through the resonator structure are somewhat decelerated before being captured by the final anode or collector. Thus the plate dissipation is reduced and the over-all efficiency is improved.

Fig. 19 shows the arrangement of cathode, grid, and

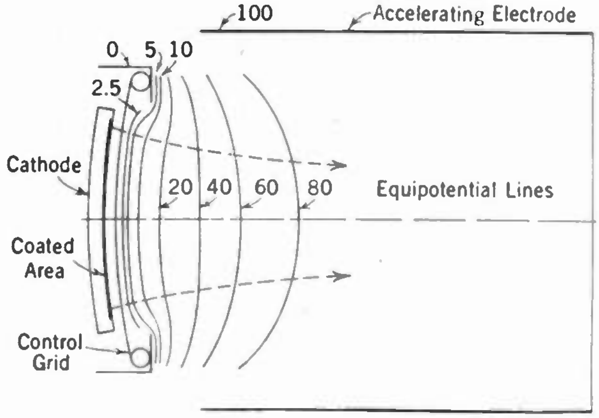
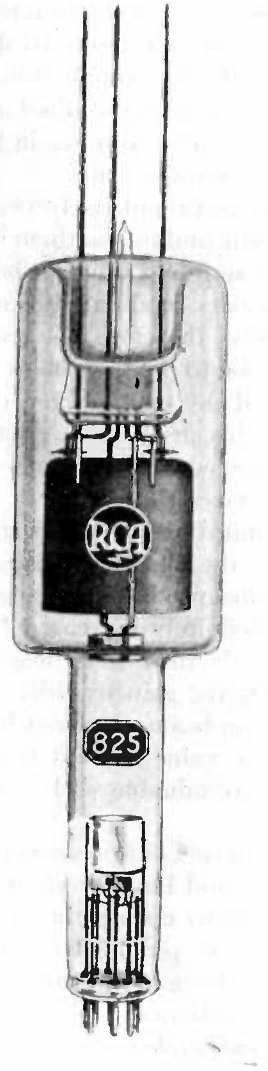


Fig. 19—Control grid and cathode structure of the inductive output tube.

accelerating electrodes with equipotential lines expressed in per cent of the potential applied. The control grid is of relatively fine mesh and is placed very close to the cathode so that a large value of transconductance is obtained. The high value of the accelerating field and the small spacings result in a very short transit time and therefore reduce the effective input conductance to a reasonable value. Fig. 20 shows the commercial model of this tube.

It is appropriate to investigate the exact source of the output power in this tube, for the result is not obvious. Electrons are emitted from the cathode in bunches under the action of the control grid. These bunches are accelerated to a high velocity by the action of the accelerating electrodes, but a negligible number reach these electrodes. The electrons are decelerated by the relatively high voltage existing across the gap of



(Courtesy of RCA Victor Division of Radio Corporation of America)

Fig. 20—Commercial form of the inductive output tube.

the resonator and are finally collected by the plate. If the accelerating voltage is designated V_a and the voltage across the output resonator is V_0, then electrons which leave the resonator have a velocity corresponding

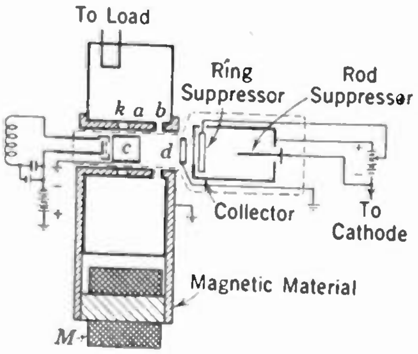


Fig. 21—Circuit for use of the 825 inductive output tube.

to an applied voltage of $V_a - V_0$. They are, however, still in a region of potential equal to V_a, and therefore a plate or collector voltage at least V_0 above the cathode voltage is necessary if the electrons are to be captured at the plate. Thus we conclude that the output power is taken directly from the plate-supply voltage and that a good conversion efficiency is at least theoretically possible.

The arrangement of Fig. 21 using the 825 tube shown in Fig. 20 gives the following experimental results:

Accelerating potential	3000 volts
Beam current	40 milliamperes
Transconductance	6000 micromhos
Input capacitance	6 micromicrofarads
Frequency	500 megacycles
Effective bandwidth	10 megacycles
Equivalent shunt capacitance of output resonator	2 micromicrofarads
Input of power	1 watt
Output of power	10 watts
Efficiency	25 per cent

Similar tubes under other conditions give larger power outputs and somewhat larger values of high-frequency power amplification.

The low equivalent shunt capacitance of the output resonator mentioned previously has an important practical significance. In the theory of video amplifiers it is shown that a low shunting capacitance is necessary if high amplification is to be obtained over wide bands of frequencies. This proposition is a perfectly general one, equally applicable here. At these frequencies, however, it is seldom practical to use the relatively elaborate compensating networks often used in video amplifiers. Accordingly, wide-band amplification is achieved most directly by reduction of the capacitance. In the present instance a bandwidth of 10 megacycles is achieved by reduction of the effective load impedance. An important property of this particular tube is that the efficiency is not seriously affected by such an adjustment of the impedance.

XV. Associated Tubes

A variety of tubes more or less similar to the Klystron are described by Hahn and Metcalf.[15] They use grid structures such as are shown in Figs. 1 and 2 and utilize drift and retarding field methods of converting velocity modulation into intensity modulation. These tubes are used as oscillators, amplifiers, and modulators at frequencies from 60 to 6000 megacycles. Certain of the tubes give large power outputs, good efficiency, and excellent stability.

One of the first to publish on the subject of velocity-modulation tubes was Heil.[16] His arrangement is not greatly dissimilar to the Klystron. Two pairs of grids are used and a drift space between the two transforms velocity modulation into intensity modulation. Only one resonator rather than two is used, however, and the device serves only as an oscillator and not as an amplifier or modulator.

XVI. Conclusions

The velocity-modulated tube, in common with the positive-grid and magnetron oscillators, depends for its operation upon electron transit time. The positive-grid tube serves only as an oscillator, no practical arrangements for producing amplification being known. As an oscillator the efficiency is low because the transit time is affected by the existence of an output voltage. The undesired phenomenon of phase selection sets in to limit the output and efficiency. Moreover, the output frequency depends upon the applied voltage and the tube structure as well as upon the tuning of the resonant circuit.

The magnetron as a transit-time oscillator has two major advantages over the positive-grid generator. The direct and important loss of electrons to the grid mesh does not exist, and the process of phase selection is absent. Accordingly, the magnetron is ideally capable of 100 per cent efficiency. Although this ideal is not closely approached in practice, the efficiencies of commercial magnetrons are high in comparison to the efficiencies of positive-grid tubes. In the magnetron the frequency depends primarily upon the tube structure and the magnetic field, and only secondarily upon the tuning of the resonant circuit. Relatively good frequency stability is achieved in practice, particularly if permanent magnets are used to produce the required field. No practical arrangement for the amplification of hyper frequencies by means of the magnetron has yet been disclosed.

Thus the Klystron, in common with similar velocity-modulated tubes, has the distinction of being the only known device for the amplification of hyper frequencies. This success may be attributed to the fact that transit time, although essential to the operation of the tube, does not contribute any critical time interval equivalent to a resonance. The two cavity resonators function in a manner which is essentially independent of the electron beam. Because of this independence of the resonators and the electron stream it is possible to excite one resonator with a small signal and to abstract a large signal from the second resonator.

In many respects the Klystron is simpler than other widely accepted types of tubes. No magnetic field is required, and no extremely high voltage is required. Relatively strong oscillations are produced with an applied potential of only a few hundred volts. With improvements comparable to those which have been made in other tubes there seems no reason to doubt

[15] W. C. Hahn and G. F. Metcalf, "Velocity-modulated tubes," Proc. I.R.E., vol. 27, pp. 106–116; February, 1939.

[16] A. Arsenjewa-Heil and O. Heil, "A new method of producing short undamped waves of great intensity," *Zeit. für Phys.*, vol. 95, pp. 752–762; July 12, 1935.

that maximum ratings below 1000 volts will soon be available.

At the present time the most severe drawback of the Klystron is the high cost of the tube and the associated tuner. Because the tube itself is not essentially more complex than other available types it seems probable that mass-production methods will reduce the cost to a value well within the reach of all. The tuner design now employed requires machine work of high precision. It is thus intrinsically expensive. Past experience, a relatively trustworthy guide, indicates that modified designs will be produced which accomplish the same results less expensively.

In the Klystron we find a device capable of oscillation, amplification, modulation, and detection. Accordingly the situation at hyper frequencies now is closely comparable to that which existed at lower frequencies some twenty-five years ago when the triode was relatively new. The developments in communication, navigation, aviation, and industry which will result are beyond the power of any prophet to predict. We may be sure, however, that the steadily accelerating advance in the knowledge and use of electronic devices will not halt and that hyper frequencies will play a leading part in this advance.

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Institute News and Radio Notes

Board of Directors

The regular meeting of the Board of Directors took place on June 2, 1943 and those present were: L. P. Wheeler, president; F. S. Barton, vice president; S. L. Bailey, W. L. Barrow, E. F. Carter, I. S. Coggeshall, W. L. Everitt, H. T. Friis, Alfred N. Goldsmith, editor; G. E. Gustafson, R. A. Heising, treasurer; F. B. Llewellyn, Haraden Pratt, secretary; F. E. Terman, B. J. Thompson, H. M. Turner, H. A. Wheeler, W. C. White, and W. B. Cowilich, assistant secretary.

The following applications were approved: for transfer to Member in the names of Rawson Bennett, S. D. Eilenberger, H. C. Harvey, R. P. Jutson, F. V. Long, J. L. Richey, H. C. Singleton, and A. H. Reesor Smith; for admission to Member in the names of S. L. Davis, G. O. Milne, K. R. Patrick, F. H. Pumphrey, and A. G. Tynan; and, 141 for admission to Associate, 81 for Student, and 6 for Junior.

The resolution, quoted below, was unanimously adopted and relates to the Kilgore-Patman Bills S-702 and HR-2100 proposing the mobilization of the scientific and technical resources of the Nation and the establishment of an Office of Scientific and Technical Mobilization:

Resolution of the Board of Directors of the Institute of Radio Engineers Regarding the Kilgore-Patman Bills S-702, HR-2100

June 2, 1943

"*Whereas*, The Board of Directors of The Institute of Radio Engineers is of the opinion that the scientific and technical resources of the Nation and in particular the radio personnel and facilities of the country are mobilized to a high degree and are working efficiently in the war effort; and

"*Whereas*, It appears that enactment of The Kilgore-Patman Bills S-702, HR-2100 to mobilize the scientific and technical resources of the Nation, to establish an Office of Scientific and Technical Mobilization, and for other purposes, would actually endanger the war effort by a reorganization of these resources along totally untried lines, and

"*Whereas*, It is the opinion of the Board of Directors that the premises given in the declaration of policy of S-702 are unsound and not representative of the facts; and

"*Whereas*, The enactment of these bills would establish a postwar bureaucracy inimical to the best interests of scientific and technical progress and thus also to the best interests of these United States; therefore, be it

"*Resolved*, That the Board of Directors of The Institute of Radio Engineers finds no valid reason for enactment of Senate Bill S-702 and House Bill HR-2100 and strongly opposes such enactment because these bills if enacted will result in confusing the war effort; and furthermore be it

"*Resolved*, That the Board of Directors of The Institute of Radio Engineers expresses its general opposition to any proposals which would have the effect of placing the scientific and technical personnel and facilities of the Nation under government supervision and control."

It was further decided to send copies of this resolution to the Congressional Committees on the Senate Bill S-702 and to distribute appropriate publicity on the subject bill to the technical and lay press.

Secretary Pratt, as chairman of the special committee on the subject, presented an interim report on the progress in promoting the Radio Technical Planning Agency, according to the postwar plan recently proposed by the Board of Directors. The report was unanimously accepted.

It was noted that industry is widely interested in the broad R.T.P.A. plan and that a number of replies have already been received from proposed sponsors in favor of the named organization.

President Wheeler stated that he and General Counsel Zeamans recently presented to the War Production Board, at Washington, the Institute's supplementary appeal for an exception to the modified WPB Paper Limitation Order L-244.

As an indication of co-operation with the War Production Board, Editor Goldsmith announced that the PROCEEDINGS would be printed on lighter-weight paper as soon as possible, but that the sharp increase in number of papers and number of I.R.E. members this year limits the extent to which paper economies can be made in the PROCEEDINGS without injury to the war effort.

Approval was granted to making translations of radio-and-electronic papers from certain foreign technical journals and to publishing such translations in the PROCEEDINGS.

Treasurer Heising gave a progress report on the study that is being made of the Institute's investments.

The employment of Klauser and Todt, Certified Public Accountants, to make the Institute's 1943 financial audit, was approved.

A budget was approved for the office alterations proposed by the Executive Committee.

President Wheeler stated that flowers were sent to the funeral of the late Past President John Stone Stone, who died on May 20, 1943.

The St. Louis Section was granted permission to affiliate with the Joint Council of the Associated Engineering Societies of St. Louis.

Mr. R. B. Shanck was appointed to the Membership Committee.

It was decided to postpone the September meeting from the first to the eighth of that month.

The resolution of the Philadelphia Section, recommending an expansion of the Institute's activities to include electronics, was read by President Wheeler and unanimously accepted.

Mr. Thompson, chairman of the Membership Committee, called attention to the letters and other material which were recently mailed to all members of that group, as part of the year's program to obtain new members.

The matter of arrangement of amendments was discussed relative to the ballots on the proposed Constitutional amendments which, prior to July 1, 1943, are to be mailed to the voting membership.

Executive Committee

The Executive Committee met on June 1, 1943 and those in attendance were L. P. Wheeler, chairman; Alfred N. Goldsmith, editor; R. A. Heising, treasurer; F. B. Llewellyn, Haraden Pratt, secretary; H. A. Wheeler, and W. B. Cowilich, assistant secretary.

The assistant secretary reported on personnel matters, including the resignation of an office clerk and the employment of another to fill a previous vacancy.

Salary adjustments were approved in the case of three clerical employees who were simultaneously transferred from temporary to permanent basis, and of two other staff employees on the basis of individual merit.

These applications were recommended to the Board of Directors for approval: for transfer to Member in the names of Rawson Bennett, S. D. Eilenberger, H. C. Harvey, R. P. Jutson, F. V. Long, J. L. Richey, H. C. Singleton, and A. H. Reesor Smith; and for admission to Member in the names of S. L. Davis, G. O. Milne, K. R. Patrick, F. H. Pumphrey, and A. G. Tynan.

The 141 applications for Associate, 81 for Student, and 6 for Junior grades were also approved for confirming action by the Board of Directors.

Chairman Wheeler reported that, on May 26, 1943, he and General Counsel Zeamans personally appeared before the War Production Board, at Washington, and presented the supplementary data requested by that agency relative to the Institute's appeal for exception to the modified Paper Limitation Order L-244.

As a paper-economy measure, Editor Goldsmith announced that the PROCEEDINGS would be printed on lighter-weight paper and that this would be done as soon as the particular stock could be obtained by the printer.

As additional steps toward conserving paper, consideration was given to withholding the publishing of further editions of the Yearbook and the Cumulative Index until after the war.

It was decided to increase the advertising rates of the PROCEEDINGS in view of the substantial increase in membership and subscribers.

The translations of papers on radio and electronic subjects, appearing in foreign-language technical journals, were considered for publication in the PROCEEDINGS.

It was unanimously agreed to recommend to the Board of Directors that R. B. Shanck be appointed to the Membership Committee.

Requests from the Chicago and St. Louis Sections were discussed and referred to the Board of Directors.

Chairman Wheeler reported that the Institute had telegraphed flowers to the funeral of the late John Stone Stone, Institute's President in 1915, who died on May 20, 1943, at San Diego, Calif.

It was decided to hold meetings of the Executive Committee during July and August, if necessary.

Approval was granted to having the Institute's 1943 financial audit made by Klauser and Todt, Certified Public Accountants.

Secretary Pratt announced that arrangements had been completed for the installation of the payroll-check system, previously approved, and that it would become effective July 1, 1943. Certain office alterations were approved.

Further discussion was held on the subject of the Institute's investments.

Arthur Van Dyck on Active Naval Duty

Arthur Van Dyck, Junior Past President of the Institute of Radio Engineers, entered active duty in the Navy, as

ARTHUR VAN DYCK

Lieutenant Commander, stationed in Washington. He is in the Office of Naval Operations, Communications Division, with duties pertaining to Electronic Matériel.

Commander Van Dyck was commissioned in the United States Naval Reserve in 1925. For the first year of the war he continued in his prewar position in the RCA Laboratories, engaged in war radio and radar developments, and visited England during the year.

Oswald F. Mingay

One of the I.R.E. members in Australia, Oswald F. Mingay, is now visiting the United States and Canada. He has been here for about four months and has met many members of the Institute and inspected a considerable number of radio plants engaged on electronic war work. Some of them are producing equipment destined for Australia, where General MacArthur commands all the American and Australian troops in the Southwest Pacific Area.

OSWALD F. MINGAY

Mr. Mingay, as a Captain in the Australian Signal Corps, was loaned by the Army to the Radio and Signals Directorate of the Ministry of Munitions, (equivalent to the War Production Board in the United States) and was manager of the Radio Division. He was sent by his Directorate to investigate radar-radio production and development in America. He was general secretary of the Australian Institute of Radio Engineers since its inception in 1932, and until he joined the Ministry of Munitions last year.

W. M. Angus

W. M. Angus, formerly designing engineer of the receiver division of the General Electric Electronics Department at Bridgeport, Connecticut, has been named engineer of the division, according to an announcement by I. J. Kaar, managing engineer.

W. M. ANGUS

A graduate of the University of Toronto, Canada, in 1926, Mr. Angus was employed by General Electric in 1936 as an engineer in the company's radio and television receiver division. In 1940, Mr. Angus received the Coffin Award, highest honor bestowed upon a General-Electric employee, for the execution of an idea for the automatic winding of coils used for touch-tuning of radio receivers and transformers. In the fall of 1941 he was appointed designing engineer for the General-Electric receiver division which position he held until his recent appointment as engineer of the division. Mr. Angus is an Associate member of the I.R.E.

John K. Hilliard

John K. Hilliard, chief transmission engineer of the Metro-Goldwyn-Mayer sound department and Member of The Institute of Radio Engineers, recently consultant to the Radiation Laboratories of Massachusetts Institute of Technology, has joined the war production staff of Altec Lansing Corporation, Los Angeles, as chief engineer of the Radar and Motion Picture Division. He is chairman of the Theater Standards Committee of the Research Council of the Academy of Motion Picture Arts and Sciences, a member of the Motion Picture Standards Committee of the Royal Scientific Society of Great Britain, and is the author of many publications on technical subjects in the communications and motion-picture fields.

JOHN K. HILLIARD

Books

Electric and Magnetic Fields, by Stephen S. Attwood

Published by John Wiley and Sons, Inc., 440 Fourth Avenue, New York, N. Y. 420 pages+9-page index+xiii pages. 215 figures, $6\frac{1}{4}\times9\frac{1}{4}$ inches. Price, $4.50.

This book deals almost exclusively with electric and magnetic fields separately; the last chapter however contains a brief discussion of two important examples of interaction between electric and magnetic fields, waves guided by a pair of parallel wires, and radiation from an accelerated charge. The presentation is elementary with a strong emphasis on field sketching and with abundant graphical illustrations. Laplace's equation is barely mentioned; but in so far as electrostatics and magnetostatics are concerned the book is quite adequate for ordinary practical purposes. A large number of problems is included.

The book is divided about equally between electric and magnetic fields. Part I, dealing with electric fields, begins with an elementary discussion of the electrical constitution of matter, experimental facts, and definitions. Then, the author passes to the principles of field sketching and illustrates these principles by sketching the fields of several simple combinations of point charges. From there on the author develops conventional electrostatics: energy storage, boundary conditions, fields with simple geometric properties, polarization, electrostatic induction, the method of images, etc. There is a brief chapter on space charge and thermionic emission in which equations are developed for a steady stream of charge between parallel electrodes.

Parts II and III deal with magnetic fields, Part III being concerned particularly with fields in ferromagnetic media. Part II is, of course, analogous to Part I and for this reason it is not as extensive. Ferromagnetic fields are discussed in considerable detail and copiously illustrated with diagrams and charts.

The remaining Part IV has two chapters: one on Units and Standards and the other on Interaction of Electric and Magnetic fields. This last chapter contains a section dealing with oscillation in an electric circuit consisting of a resistance, inductance, and capacitance in series; a section on waves along parallel wires; a section on the field of a uniformly moving electric particle; and a section on radiation from an accelerated particle. The book is concluded with two interesting sketches, one on electrons in atoms and the other giving an historical outline.

S. A. Schelkunoff
Bell Telephone Laboratories, Inc.
New York, N. Y.

Notice

In the case of certain texts, it is believed to be helpful to the readers of the Proceedings to have available one or more expressions of opinion, in book-review form, from competent reviewers. There are accordingly here presented in sequence two reviews of a recently published book.

The Editor

Frequency Modulation, by August Hund

Published by the McGraw-Hill Book Co., 330 West 42 Street, New York 18 N. Y. 368 pages+7-page index+X pages 113 illustrations. 6¼×9¼ inches. Price, $4.00.

This is the first book that purports to be an engineering text on frequency modulation covering both basic principles and the design of commercial apparatus.

The need for such a book is apparent when it is realized that the system of frequency modulation invented and developed by Professor E. H. Armstrong was demonstrated before the radio art in 1935 to provide a method of reducing disturbances in radio signaling. Nevertheless, the art has been slow to take advantage of the improvement frequency modulation offers in radio communication. In part this can be attributed to a lack of understanding of the fundamental principles of this new development and the engineering-design considerations involved in good engineering practice.

This reviewer is disappointed to find that while Hund's book contains much that is of value, the author has obscured the fundamental relations and failed to present the subject in a clear concise manner. This failure to provide a clear-cut presentation of frequency modulation is disappointing to a greater degree than it would be under normal circumstances because of our dependence on having the best communication during this national emergency.

The author does not seem to realize that the purpose of a textbook is to dispel and not create confusion. This is shown by (1) the treatment of "frequency" and "phase" modulation, and (2) the handling of the historic facts, which are inaccurate to a degree that is hard to understand since the author has a most complete compilation of publications on the historical background of frequency modulation.

With regard to (1), there are scores of pages under the heading "Fundamental Relations" whose main effort appears to be to establish a basic difference between what is termed "frequency" modulation and what is termed "phase" modulation. The subject first came up in Roder's paper entitled "Amplitude, Phase, and Frequency Modulation" published in the Proceedings of the I.R.E. in December, 1931, and in the subsequent discussion with David Luck in the Proceedings for May, 1932.

After Luck had straightened out some errors in Roder's mathematical operations, they both agreed that "phase" modulation was simply a "frequency" modulation in which the extent of the deviation increased directly as the frequency of the modulating current. In other words, the question of whether a transmitter radiates phase or frequency modulation resides in the frequency characteristic of an audio amplifier. This is the concept accepted by the radio art of the relation between phase and frequency modulation.

Nowhere does the author give this simple explanation. He sets up equations for phase and frequency modulations and talks about the difference in the equations. Having done that, he proceeds to insist that, there is a *great* difference between phase modulation and frequency modulation. The absurdity of this artificial distinction appears at once when pre-emphasis (renamed "pre-accentuation") is introduced in the characteristic of the audio modulating system. The result is confusion of the concepts of frequency modulation and for the average man the physical properties of the modulation become obscured.

The second point referred to, the historical inaccuracy, appears on page 231 in the following statement: "Some of the important theoretical contributions that were made prior to any important engineering applications did not exactly predict great advantage for FM over AM."

Both Carson's paper and the paper by Roder, that has been referred to, make definite predictions of great disadvantages for frequency modulation. The history of the invention and development of frequency modulation contains a lesson about invention of such fundamental importance to the younger part of the engineering profession that it would be a great pity to keep from them the opportunity of learning from it what they may. The author not only succeeds in covering this lesson up, but, in effect, in teaching the opposite one. To the reader of Hund's book who is not familiar with the history of frequency modulation, Professor Armstrong, the inventor, is portrayed as the designer of an "indirect" frequency-modulation transmitting system. Armstrong's paper, which overturned so many of the basic concepts long held by the art, is referred to in the bibliography as "an unusual non-mathematical description of the Armstrong indirect FM system, describing many of the auxiliary apparatus employed."

The book directs attention exclusively to the broadcasting application of frequency modulation, creating the impression that this is the only field of use. That the contrary is true the art knows, from the many applications in the mobile and point-to-point services.

The basic relation between frequency deviation and the audio band in relation to improvement over the equivalent amplitude-modulation system is hardly presented and is obscured in a maze of mathematics. Only the most discerning reader will realize that voice communication in the range of 0 to 3000 cycles requires only a 30-kilocycle bandwidth to achieve the same reduction in disturbance that is obtained in the high-fidelity broadcasting system using a 150-kilocycle bandwidth.

"Frequency Modulation" will perhpsa delight the mathematical physicist but it is not the work on frequency modulation that will give the radio art and the classroom a clear understanding of the fundamental principles and the basic engineering concepts so necessary to speed the development of this new method of communication.

Paul A. de Mars
Navy Department
Bureau of Aeronautics
Washington, D. C.

The author of "Frequency Modulation" who has had a varied experience as a teacher, as a research worker in both commercial and government laboratories and as a consulting engineer has prepared this engineering text on frequency modulation to present both the basic principles and the commercial design of frequency-modulated radio equipment as it was just prior to the present war. Approximately the first half of the book discusses the fundamentals of frequency modulation but includes some information on the circuits of three different commercial radio transmitters as well as a section on wave propagation in the 40- to 50-megacycle band. The remainder of the book deals with auxiliary apparatus employed in frequency-modulation systems, gives detailed descriptions of specific commercial types of frequency-modulation radio broadcast transmitters and receivers, and includes a chapter on antennas and transmission lines.

The subject matter in the first part of the book makes a careful distinction between amplitude, phase, and frequency modulation. The discussion is to a degree mathematical and possibly somewhat advanced for most engineering undergraduates. The entire book assumes a good knowledge of alternating-current circuit theory and vacuum tubes. Working formulas for spectrum distributions with various types of modulation are presented together with tables of Bessel functions of the first kind to facilitate application of the formulas. (Note that in Table VI J_{10} (18) obviously should be -0.07317. See Jahnke and Emde "Tables of Functions.")

The examples and much of the text material are based upon the present Rules of the Federal Communications Commission governing the use of frequency-modulated broadcasting. No mention is made of the use of or equipment for frequency-modulated systems in other fields of communication. The author makes frequent reference to his earlier published books for explanations which are needed for an understanding of the subject matter but are not included in this text.

The book is a new text and should be of value to frequency-modulation broadcasting engineers. A table of references at the back of the book should be extremely valuable to those engineers interested in more complete information than can be obtained from the text itself on any of the subjects discussed.

Austin Bailey
American Telephone and Telegraph Co.
New York, N. Y.

The Future of Television, by Orrin E. Dunlap, Jr.

Published by Harper and Brothers, 49 E. 33 St., New York, N. Y. 187 pages +6-page index+xi pages. 18 illustrations. 6×8¾ inches. Price, $2.50.

This book is one of a series written by this author in which he describes various aspects of radio communication—particularly broadcasting—of interest to the layman. In this series, this book is the second one relating to television, a book entitled, "The Outlook for Television," having been published in 1932. While the title of the present book refers to the future, it is in substantial part devoted to a review of television developments during the past ten years as a background to "the forward look."

One chapter, devoted to the problems of launching a new industry, deals in some detail with the activities of the Federal Communications Commission and the co-operating National Television Standards Committee. This discussion brings out the difficulties involved in steering a course between unhampered development and protection of the immediate interest of the purchaser of equipment. The investor is advised to scrutinize cautiously the financial prospectus of any unproved television enterprise.

Several very readable chapters discuss the place of television in the home, the unusual situations that arise in the production of television programs, and the relation between television and other industries such as motion pictures and the theater. The view expressed that the job of television is to bring "the greatest show on earth" into the home and that "newscasting is to television what the newsreel is to the theater screen." In his discussion of the outlook for sound broadcasting, the author emphasizes that television is an unsafe ground for prophecy but he seems to share the views of others in the industry whom he quotes as believing that broadcasting will go through a gradual transition period as sight is added to sound, first in broadcast stations in the large centers of population and later by some form of relaying system in broadcast stations throughout the country. Financial support for television is viewed as dependent on either the advertising sponsor or the "admission gate."

The author apparently assumes that the layman is not much interested in the technical side of television beyond the knowledge that the production of such programs in the home is the equivalent of showing at least 24 different "lantern slides" per second, about 200,000 separate bits of information being required to make up each slide. In the discussion of the evolution of television, he emphasizes the fact that this field is not the invention of one man but represents collective effort in research.

The book closes with a chronology of historic steps in television beginning with the mathematical prediction of electromagnetic waves by Maxwell in 1867. It is the author's prediction that after the war television will be one of the contributions of science to the new era promised by the Atlantic Charter, further breaking down the barriers between nations.

Laurens E. Whittemore
American Telephone and Telegraph Co.
New York, N. Y.

Basic Radio, by J. Barton Hoag

Published by D. Van Nostrand Co., 250 Fourth Avenue, New York, N. Y., 1942. 358 pages+21-page index+xii pages. 425 figures. 5¾×8¾ inches. Price $3.25.

This little book contains a great deal of factual information useful in the radio arts. The word "Basic" in the title refers to the fact that the author has attempted to select and deal with only the important and time-tested tubes and circuits.

It is inevitable that the author of this book has had constant exposure since the outbreak of war to the necessity of finding a way to use large numbers of men who are intelligent, but not technically trained, to operate and maintain the complex radio tools of modern war. This book can be valuable in helping such men to become useful for radio war services.

The treatment presumes no knowledge of trigonometry, only the barest and most elementary knowledge of high-school algebra, and a few concepts from elementary physics. However, it demands of the reader a high native intelligence, and a willingness to give concentrated and persistent study to the subject.

The author begins at the beginning and goes through to the end. The beginning is a description of the electron and of the most elementary electric-circuit concepts. The end is a description of the behavior of ultra-high-frequency electronic-power sources and of wave guides. It is obvious that to cover in 336 pages even the basic tubes and circuits of radio utility between these two extremes something must be omitted in the manner of treatment.

In general, the things that are omitted are rigorous proofs, qualifying phrases, finer details of circuit analysis, and any subject matter not immediately pertinent to widely used devices. Explanations are such as to convey mental pictures rather than to establish proofs. Careful study of this book will make the behavior of radio devices seem reasonable as well as orienting their uses effectively. However, the point of view so obtained will be that of a highly skilled tradesman, who knows how to use things, rather than that of an engineer, who has a genuinely firm knowledge as to why they work. In the discussions of reasons why certain technical functions are useful, the logic of the presentation is clear and concise. In the analysis of technical functions facts rather than logical points are emphasized.

The problems at the end of the text are graded as to difficulty, and are chosen to give the student familiarity with over-all practical considerations.

W. G. Dow
University of Michigan
Ann Arbor, Michigan

Report of the Secretary—1942

The following report has been prepared to inform the membership of the activities of the Institute during the calendar year 1942.

Membership

The paid membership of 8794 represents an increase of 25.3 per cent over that of 1941, and a new high in the Institute membership. Fig. 1 gives the membership figures during the life of the Institute.

The domestic membership during 1942 showed a gain of 29.4 per cent while the foreign membership, despite wartime restrictions, increased by 0.5 per cent.

The proportion of the membership outside the United States and its possessions dropped from the high of 24.2 per cent in 1935, to 11.6 per cent.

There were 2065 new members elected to the Institute during 1942, or an increase of 26 per cent as compared with 1643 for 1941.

TABLE I

MEMBERSHIP DISTRIBUTION BY GRADES

		Per cent
Fellow	168	1.9
Member	833	9.5
Associate	6205	70.6
Junior	55	0.6
Student	1533	17.4
	8794	100.0

During the year 1984 applications for membership were received, or 23 per cent more than the 1608 received during 1941. The applications for student grade increased 74 per cent and those for all other grades decreased 3.8 per cent.

work of the officers of the Papers Procurement Committee of the Institute, namely, D. D. Israel, chairman; and W. L. Everitt and J. R. Pierce, vice chairmen. The 1942 volume of the PROCEEDINGS (Volume 30) included 560 pages of technical and editorial material. The trend during the year was toward an increasing number of pages of technical material per issue.

Wartime conditions have made it difficult to place each issue of the PROCEEDINGS before the members prior to the 15th day of the month of issue. However, by the establishment of a definite routine procedure with firm deadlines, it has proved possible to place each issue of the PROCEEDINGS before the members of the Institute on or before the scheduled date, mentioned above.

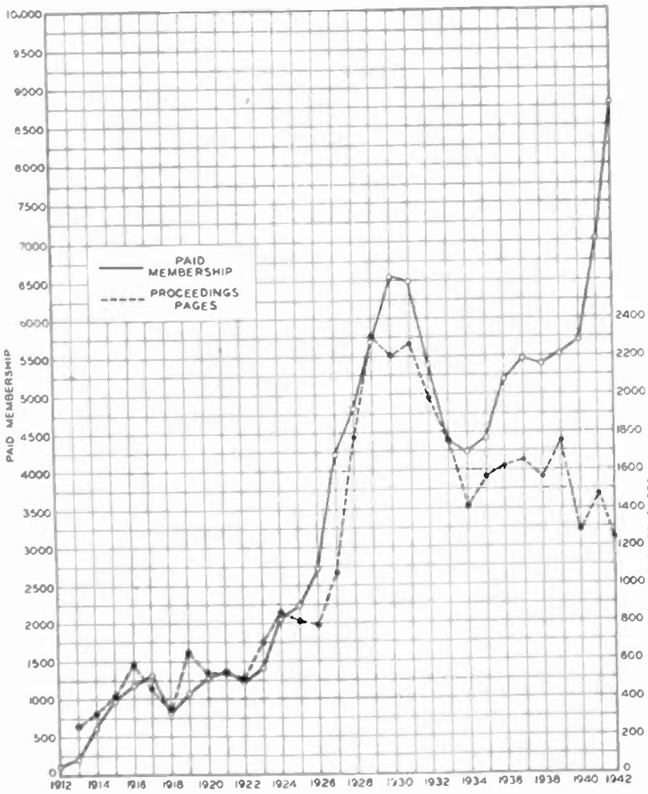

Fig. 1—The variation in paid membership is shown by the solid graph. The dotted line is for the number of pages of technical and editorial material in the PROCEEDINGS. Starting in 1939, a larger format was used and the scale of pages should be divided by 2.2.

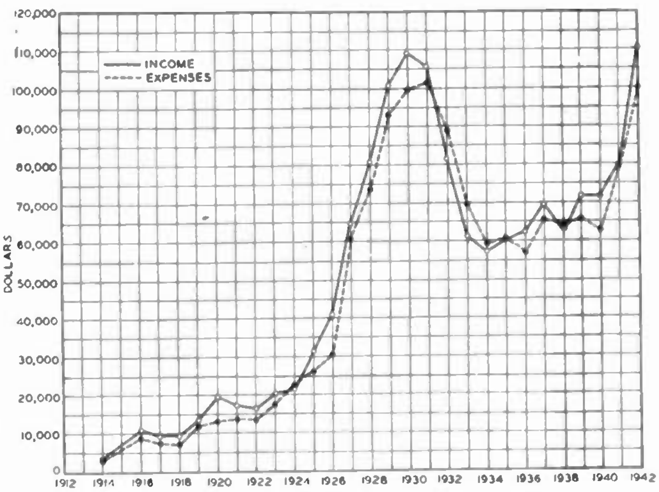

Fig. 2—Income and expenses are plotted for the life of the Institute.

During 1942, the membership in the British Empire rose 13.6 per cent as contrasted with the 8.3 per cent gain during the previous year. The European membership decreased 23.3 per cent during 1941 and 4.2 per cent during 1942.

Increases of 28 and 6 per cent took place in the South American membership in the years 1941 and 1942, respectively.

Table I shows the distribution of the membership by grades. The percentage of the total in each grade is also given.

Proceedings

During 1942 there were published in the PROCEEDINGS seventy-five technical papers. Despite certain necessary wartime restrictions, the flow of technical material of merit has tended to increase during the year, and the engineering quality of the papers is regarded as fully on a parity with those of the average peacetime year. The procurement of papers has been in large measure due to the systematic and effective

Under the continued direction of Editor Goldsmith, two marked trends have characterized the editorial content of the PROCEEDINGS during 1942. One is in the increasing number of papers dealing with the professional problems of the engineer and his position in modern society. The other is in the papers dealing with application of radio technique to noncommunication matters. A plan for previewing books was developed during the year and arrangements wemre made for its later initiation.

Tutorial papers of outstanding merit have been added to the contents of the PROCEEDINGS as an aid to the wartime instruction of radio technical personnel.

Sections

The New York Section was established during the year, bringing the total to 27 Sections.

The 179 meetings held by the Sections were 23 per cent less than in 1941. The number of members in Section territories reached the new high of 6525, or an 85 per cent gain over the 3517 in 1941.

Meetings

In addition to the 179 Section meetings, there were three conventions during the year.

The Winter Convention was held on January 12, 13, and 14 in New Hork City. A total of 25 technical papers was presented during the three days. The attendance totaled 1753 men and 37 women.

Cleveland, Ohio, was the headquarters for the Summer Convention which took place on June 29, 30, and July 1. During the five technical sessions, 33 papers were presented and a symposium was conducted. The registration totaled 256 men and 36 women.

The Rochester Fall Meeting was held on November 9 in Rochester, New York. The presentation of three technical papers, in addition to a symposium, occurred during the one-day session. A total registration of 310 was recorded.

Board of Directors

The Board of Directors held 10 meetings during 1942.

Executive Committee

The Executive Committee held 20 meetings during the same period.

Administrative Committees

The 16 committees and subcommittees, whose activities bear directly on the operation of the Institute and its relations with the membership, held 19 meetings during the year and, in addition, handled a considerable volume of work through correspondence.

Technical Committees

The technical activities of the Institute are conducted by 33 committees and subcommittees. The meetings of these groups, seven in all during the year, were curtailed by the wartime conditions.

Summary of Meetings

The other meetings, three in all, were conducted by a special group and are thus reported. Table II gives a summary of the number of meetings of all groups held during the year.

TABLE II

MEETINGS HELD DURING 1942

Board of Directors	10
Executive Committee	20
Administrative Committees	19
Technical Committees	7
Special Committees	3
	59

Constitution and Bylaws

No Constitutional amendments were submitted to the membership, but several changes in the Bylaws were adopted.

Awards

The Institute Medal of Honor for 1942 was presented to Albert Hoyt Taylor for his contributions to radio communication as an engineer and organizer, including pioneering work in the practical application of piezoelectric control to radio transmitters, early recognition and investigation of skip distances and other high-frequency wave-propagation problems, and many years of service to the government of the United States as an engineering executive of outstanding ability in directing the Radio Division of the Naval Research Laboratory.

The Morris Liebmann Memorial Prize was bestowed on Dr. S. A. Schelkunoff for is contributions to the theory of electromagnetic fields in wave transmission and radiation.

In recognition of their contributions to radio, the following nihe members of the Institute were transferred to the Fellow grade:

Wilmer L. Barrow
George H. Brown
Geoffrey Builder
Adolph B. Chamberlain
Ellsworth D. Cook
Hugh S. Knowles
Warren P. Mason
Harold O. Peterson
George C. Southworth

Finances

The comparative statement of income and expenses for 1942 and 1941, and the comparative balance sheet for the same years, are based on data from the auditor's reports for the periods, which were prepared by Klauser and Todt and by Patterson and Ridgway, respectively, both firms being certified public acountants. In addition, the income and expenses for each year of the life of the Institute are charted in Fig. 2.

Headquarters Staff

During the year considerable change in headquarters personnel took place including resignations, replacements, and one addition to the staff. One employee was on an extended leave of absence due to illness.

The PROCEEDINGS of the I.R.E. was admitted to membership in the Audit Bureau of Circulations in the same period.

At the close of the year the headquarters staff consisted of 17 permanent employees, including the addition of an editorial assistant, in contrast with 16 in the preceding year. Also, the advertising manager and his staff were continued in the employ of the Institute on a contractual basis.

Deaths

The deaths of one Fellow, two Members, twenty Associates, and one Student, whose names are listed below, were reported during 1942.

FELLOW

Pedersen, P. O., (F'15)

MEMBER

Pidgeon, H. A., (M'29)
Simmonds, E. J., (M'25)

ASSOCIATE

Becker, F. V., (A'37)
Campbell, A. G., (A'31)
Charrier, G. M., (A'37)
Dayton, B. L., (A'41)
Denman, R. P. G., (A'41)
Fortune, D. J., (A'40)
Guggenheim, S., (A'30)
Hanover, E. A., (A'27)
Hoar, J. E. C., (A'41)
Hollis, R., (A'38)
Hopfenberg, J. A., (A'40)
Maeder, A. S., (A'41)
Pedersen, A. A., (A'36)
Peters, T. A., (A'42)
Plummer, W. B., (A'40)
Roberts, E. L., (A'36)
Shreve, A. F., (A'29)
Strauss, S., (A'42)
Weis, C. L., Jr., (A'39)
Willging, L. F., (A'37)

STUDENT

Kohler, F. L., (S'41)

Acknowledgment

The substantial progress of the Institute during 1942 resulted in a large measure from the guiding efforts of President Van Dyck. The members of the Board of Directors, the Executive Committee, and the other committees also gave generously of their time and energies to, and are to be credited for their real part in, the rapid growth of the Institute which occurred despite the trying conditions during this wartime period.

Respectfully submitted,

Haraden Pratt

HARADEN PRATT
Secretary

June 30, 1943

Comparative Statement of Income and Expenses for the Years Ending December 31, 1942 and 1941

Income	1942	1941
Dues, Current and in Arrears	$ 49,363.60	$41,760.15
Entrance and Transfer Fees	3,551.00	2,926.00
Subscriptions	8,773.06	10,959.04
Advertising	38,010.00	16,560.40
Binders, Bound Volumes, Emblems, Reprints	3,643.59	3,266.15
Interest from Investments[1]	1,414.47	1,122.20
Conventions	2,970.37	2.938.50
Miscellaneous	2,527.48	178.17
	$110,253.57	$79,710.61

Expenses	1942	1941	1942	1941
Advertising Commissions, Salaries, Expenses			$ 13,214.89	$ 5,172.69
Bad Debts, Less Recoveries[2]			2,854.24	3,462.18
Binders, Bound Volumes, Emblems, Reprints			2,730.82	2,493.52
Conventions			4,447.60	4,172.31
Membership Solicitation			401.50	1,470.60
New York Meetings			866.81	881.09
Office			7,097.98	6,677.81
Depreciation	$ 775.05	$ 586.88		
Insurance	200.14	197.81		
Postage	2,858.56	3,064.91		
Stationery, supplies	2,417.40	1,935.49		
Telegraph, Telephone	846.83	892.72		
Printing			27,697.01	21,369.62
PROCEEDINGS	21,061.31	20,519.03		
Standards	2,533.35	303.06		
Yearbook	3,725.84	—		
Miscellaneous	376.51	547.53		
Professional Services, Accounting and Management			700.00	1,650.00
Rent and Electricity			3,650.32	3,228.00
Salaries			29,672.65	22,445.43
General	20,541.41	15,300.71		
PROCEEDINGS	4,153.78	5,683.12		
Standards	3,259.90	1,461.60		
Yearbook	1,717.56	—		
Sections			4,196.54	4,188.52
Securities Exchange Loss[3]			—	701.50
Miscellaneous			2,518.64	1,467.90
			$100,049.00	$79,381.17
Carried to Surplus			$ 10,204.57	$ 329.44

[1] Morris Liebmann Memorial Fund and Prize not included in this accounting.
[2] At least 99 per cent of these amounts represent nonpayment of dues.
[3] Reorganization of Chicago and Eastern Illinois Railway made exchange of securities compulsory.

The Institute of Radio Engineers, Inc.
Comparative Balance Sheet

December 31, 1942 and 1941

Assets

	December 31, 1942	December 31, 1941	Increase Decrease
Current Assets			
Cash	$ 60,773.89	$ 44,968.09	$15,805.80
Accounts Receivable—Current			
Dues	160.67	360.22	199.55
Advertising	5,175.80	2,341.32	2,834.48
Reprints	141.21	158.90	17.69
Exhibition Booths	—	150.00	150.00
American Standards Association	1,570.59	—	1,570.59
Inventories (as submitted by the management)			
Proceedings	6,942.48	7,363.28	420.80
Bound Volumes	186.00	207.00	21.00
Binders	—	216.55	216.55
Emblems	257.40	191.98	65.42
Accrued Interest on Investments	94.17	94.17	—
Total Current Assets	$ 75,302.21	$ 56,051.51	$19,250.70
Investments—At Cost (Market Value, Dec. 31, 1942, $23,765.18)	$ 36,227.04	$ 36,246.37	$ 19.33
Morris Liebmann Memorial Fund Assets			
Investments (Market Value, Dec. 31, 1942, $6,956.75)	10,012.45	10,012.45	—
	344.00	351.08	7.08
Total Fund Assets	$ 10,356.45	$ 10,363.53	$ 7.08
Furniture and Fixtures after Reserve for Depreciation	5,983.40	5,279.70	703.70
Prepaid Expenses			
Unexpired Insurance	83.27	49.76	33.51
Stationery Inventory—Estimated	200.00	200.00	—
Convention Expenses	—	327.19	327.19
Medal of Honor	101.00	—	101.00
Salaries	—	181.20	181.20
Yearbook Expense	—	98.70	98.70
Total Prepaid Expenses	384.27	856.85	472.58
Total Assets	$128,253.37	$108,797.96	$19,455.41

Liabilities and Surplus

	December 31, 1942	December 31, 1941	Increase Decrease
Accounts Payable	$ 4,142.07	$ 3,717.01	$ 425.06
Section Rebates	750.50	170.00	580.50
Advance Payments			
Dues	23,233.25	15,708.12	7,525.13
Subscriptions	3,331.78	1,862.51	1,469.27
Binders	98.00	—	98.00
Amount Withheld from Employees	630.37	335.41	294.96
Total Liabilities	$ 32,185.97	$ 21,793.05	$10,392.92
Morris Liebmann Memorial Fund			
Investments	10,012.45	10,012.45	—
Unexpended Income	344.00	351.08	7.08
Total Fund	$ 10,356.45	$ 10,363.53	7.08
Deferred Income—Conventions	—	1,135.00	1.135.00
Surplus—Donated	1,997.80	1,997.80	—
Surplus—Earned			
	73,508.58	73,464.60	43.98
	10,204.57	329.44	9,875.13
Total	$ 83,713.15	$ 73,794.04	$ 9,919.11
Less—Prior Year Adjustments (Net)	—	285.46	285.46
Total Earned Surplus—December 31	$ 83,713.15	$ 73,508.58	$10,204.57
Total Liabilities and Surplus	$128,253.37	$108,797.96	$19,455.41

Contributors

Warren H. Bliss (A'41) is an assistant professor of electrical engineering at the University of Maine, Orono, Maine, and spends the summer vacation period with the Terminal Facilities Research Group of

Warren H. Bliss

the RCA Laboratories in New York. During the school year, Mr. Bliss is on a part-time schedule conducting co-operative tests with the Terminal Facilities Research Group on communication problems.

⁂

George H. Brown (A'30–F'42) was born on October 14, 1908, at North Milwaukee, Wisconsin. He received the B.S. degree at the University of Wisconsin in 1930; the degree of M.S. in 1931; the Ph.D. degree in 1933; and his professional degree of E.E.

George H. Brown

in 1942. From 1930 until 1933, he was a Research Fellow in the electrical engineering department at the University of Wisconsin, and from 1933 to 1942 he was in the research division of the RCA Manufacturing Company at Camden, New Jersey. Since 1942, he has been at the RCA Laboratories at Princeton, New Jersey. He is a Member of Sigma Xi and the American Institute of Electrical Engineers.

⁂

E. D. Cook (A'21–F'42) received the B.S. degree in electrical engineering from Union College in 1920. Following this, he did graduate work both at Harvard University and at Union College and an M.S. degree was awarded to him by Union College for this work in 1921. He received the degree of Ph.D. from Union College in 1923.

Dr. Cook spent several years as a designer of alternating-current machinery with the General Electric Company, later transferring to the special development section of the radio department where his work was concerned with the application

William A. Edson

of electron tubes to power problems, general radio developments, and talking motion pictures.

In 1929, Dr. Cook resigned from the General Electric Company to become chief engineer of a development laboratory which the Brunswick Balke Collender Company of Chicago was then in the process of forming. He later assisted with the establishment of the United Research Corporation of Long Island City, a subsidiary of that company, whose work was in the field of radio developments and talking movies. This organization was subsequently purchased by Warner Brothers. In 1932, Dr. Cook joined the advanced development division of the photophone department of the Radio Corporation of America as a development engineer and consultant on talking movie problems.

In 1936, he again joined the General Electric Company to work on high-frequency problems. Shortly thereafter he was placed in charge of the high-frequency section of the general engineering laboratory of that company.

⁂

William A. Edson (M'41) was born on October 30, 1912, at Burchard, Nebraska. He received the degrees of B.S. and M.S. in electrical engineering in 1934 and 1935 respectively, both from the University of Kansas. In 1937 he received the degree of Doctor of Science in communication engi-

J. Epstein

neering from Harvard University. From 1937 to 1941 he served as a member of the technical staff in the systems development department of the Bell Telephone Laboratories. In 1941 he resigned from the Laboratories to accept the position of assistant professor of electrical engineering at the Illinois Institute of Technology.

⁂

J. Epstein (A'42) was born on December 18, 1909, at Cincinnati, Ohio. He

Lloyd Espenschied

received the E.E. degree from the University of Cincinnati in 1932 and the M.S. degree in physics in 1934. During 1934–1935, he was an instructor in physics at the Cincinnati College for Pharmacy. He was in the research division of the RCA Manufacturing Company at Camden, N. J., from 1935 to 1942. Since 1942, he has been at the RCA Laboratories at Princeton, N. J.

Lloyd Espenschied (M'13–F'24) was born on April 27, 1889, at St. Louis, Missouri. He was graduated from Pratt Institute in 1909. During 1909–1910 he was with the Telefunken Wireless Telegraph Company of America. Since 1910, Mr. Espenschied has been on the technical staff of the Bell System and is at present a consultant with the Bell Telephone Laboratories, Inc. He has taken an active part in the early development of vacuum-tube radiotelephony, broadcasting, and overseas services; and has also had a responsible part in the development of wire-line carrier telephony and telegraphy and coaxial-cable systems. He has many inventions to his credit and is the author of numerous technical papers. He participated abroad

E. W. Herold

in International Conferences on electrical communications. In 1935, Mr. Espenschied was the co-receiver of the A.I.E.E. award for a paper on Wide-Band Transmission over Coaxial Lines, and in 1940 he received the I.R.E. Medal of Honor. He is a Fellow of the American Institute of Electrical Engineers, a Member of the Acoustical Society, an Associate Fellow of the Institute of Aeronautical Sciences, and a Charter Member of the Wireless Institute.

E. W. Herold (A'30–M'38) was born on October 15, 1907, in New York City. He received the B.S. degree from the University of Virginia in 1930 and the M.S. degree from Polytechnic Institute of Brooklyn in 1942. From 1924 to 1926, Mr. Herold was with the Bell Telephone Laboratories and from 1926 to 1929 with E. T. Cunningham, Inc. In 1930 he entered the research and engineering department of the RCA Manufacturing Co., at Harrison, N. J. Since 1942, he has been with the RCA Laboratories at Princeton, N. J. He is a member of Phi Beta Kappa and Sigma Xi.

L. Malter (A'37) received the B.S. degree from the College of the City of New York in 1926 and the M.A. and Ph.D. degrees from Cornell in 1931 and 1936, respectively. He taught physics at C.C.N.Y. from 1926 to 1928. He was with the acoustic research and Photophone divisions of the RCA from 1928 to 1930, and during the summer of 1931. From 1933 to 1936 Dr. Malter was in the electronics research division of the RCA Manufacturing Co., at Camden, N. J.; from 1936 to 1938 in the high-vacuum section, and from 1938 to 1942 in the research laboratories of the RCA Manufacturing Co., at Harrison, N. J. From 1942 to date he has been at the RCA Laboratories, Princeton, N. J. He is a Fellow of the American Physical Society and a member of the American Association for the Advancement of Science and Sigma Xi.

Louis Malter

Donald W. Peterson

Donald W. Peterson (A'43) was born on June 18, 1914, at Amherst, Wisconsin. He received the degree of B.S. in electrical engineering at the University of Wisconsin in 1936. During 1936–1940 he was with the service division of the RCA Manufacturing Company in Camden, N. J. From 1940 to 1942 he was with the research division of the RCA Manufacturing Company. Since 1942, Mr. Peterson has been with the RCA Laboratories at Princeton, N. J.

Robert I. Sarbacher was born in Baltimore, Maryland on September 6, 1907. He received the B.S. degree in electrical engineering from the University of Florida in 1933. After a year of study in the mathematics department at Princeton University, he entered the graduate school of physics and communication engineering at Harvard University. He was awarded the M.S. degree in 1936 and the Sc.D. degree in 1939 at Harvard University, where, from 1936 to 1940 he served as research assistant and as instructor in physics and communication engineering. He was appointed to the faculty of the Illinois Institute of Technology in 1940, where he is associate professor of electrical engineering, and until recently was in charge of the communication engineering courses in the electrical engineering department. At present he is with the Navy department, serving as technical aid in the Bureau of Ships.

Robert I. Sarbacher

Dr. Sarbacher is a member of the American Institute of Electrical Engineers, Society for the Promotion of Engineering Education, American Physical Society, and the Acoustical Society of America.

LITERATURE

New books of interest to engineers in radio and allied fields—from the publishers' announcements.

A copy of each book marked with an asterisk (*) has been submitted to the Editors for possible review in a future issue of the Proceedings of the I.R.E.

*ELEMENTS OF RADIO. By A. Marcus and W. Marcus, prepared under the direction of Dr. Ralph E. Horton • • • *Prentice-Hall, Inc., 70 Fifth Ave., New York, N. Y.* (699 pages+xiii, 6×9½ inches, cloth bound, 504 illustrations.) This book is a basic and elementary home-study course on the fundamentals of radio which requires no advance knowledge of physics and mathematics. The authors state that it was written to the exact prescription of the Government officials responsible for training of thousands of new radio technicians. It is written in simple, non-technical language and appears to fill the need for an extremely elementary but complete text for the radio beginner. Price $4.00.

*DYNAMICAL ANALOGIES. By Harry F. Olson, Acoustical Research Director, RCA Laboratories, Princeton, N. J. • • • *D. Van Nostrand Co., Inc., 250 Fourth Avenue, New York, N. Y.* (5¾×8¾ inches, xi+196 pages, 59 figures, cloth.) The author suggests that "Analogies are useful for analysis in unexplored fields. By means of analogies an unfamiliar system may be compared with one that is better known." This book deals with the analogies between electrical, mechanical rectilineal, mechanical rotational and acoustical systems. The text assumes on the part of the reader a familiarity with the alternating current system and physics. The material and chapters are organized with great clarity and terms are well defined. Price $2.75.

*CYCLOPEDIA OF TELEVISION FACTS. Compiled by M. N. Beitman, B.S., Radio Instructor, Englewood High School, Chicago • • • *Supreme Publications, 328 S. Jefferson St., Chicago, Ill.* (5½×8½ inches, 60 pages, offset printed from typewritten text, 50 illustrations.) This little book is a simple explanation of Television and how it operates. It is done entirely without mathematics and is designed as an introduction to television for high school students or the layman who wants to know more about television. Price $.40.

*PRINCIPLES AND PRACTICE OF RADIO SERVICING. (Second Edition) By H. J. Hicks, M.S., Associate Radio Engineer, Aircraft Radio Laboratory, Wright Field, Dayton, Ohio • • • *McGraw-Hill Book Co., 330 W. 42nd St., New York, N. Y.* (6¼×9 inches, xii+391 pages, 311 figures, cloth bound.) The author has had fourteen years experience in training men for radio service work, operators for sound motion picture machines and in vocational high schools. The first edition (March, 1939) was designed to teach servicing even up to complicated radio service problems. Mathematics has been reduced to a

(*Continued on page 60A*)

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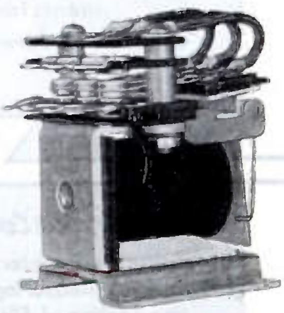

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ATLANTA

"Adjustment of Directional Antennas," by W. S. Duttera (Presented by A. W. Shropshire, Radio Station, WSB); April 16, 1943.

"The Electrical Strength of Nitrogen and Freon under Pressure," by H. H. Skilling and W. C. Brenner (Presented by Ben Akerman, Radio Station WGST); April 16, 1943.

"The Part That the Professional Engineer Plays in the War Effort," by Captain Floyd Fausett, Signal Corps, U. S. Army; May 21, 1943.

BUFFALO-NIAGARA

"Entertainment Radio—After the War: Broadcasters' Angle," by Karl Hoffman "Design Engineers' Angle," by B. E. Atwood; "Retailers' Angle," by A. J. Dykowski; and "Customers' Angle," by Members Present; May 26, 1943. Election of Officers; May 26, 1943.

CINCINNATI

"Radio Broadcast Studio Design," by C. B. Sloan, Crosley Corporation; May 24, 1943.

CONNECTICUT VALLEY

"Design, Testing, and Measurement of Solid Dielectric Transmission Lines," by H. H. Selvidge, Fournier Institute; Discussion of Paper by J. C. Coykendall, E. J. Butler, and R. E. Moe; May 28, 1943.

DETROIT

"Plastics—The New Engineering Materials," by W. B. Hoey, Plastics Processes, Inc.; May 21, 1943.

"Insulating Oils Used in Electrical Equipment," by J. C. Benedict, Detroit Edison Company; June 18, 1943.

LOS ANGELES

"The Mass Spectrometer, An Electronic Method of Chemical Analysis," by Harold Washburn, Consolidated Engineering Corporation; May 18, 1943.

"Available Commissions in the U. S. Marine Corps," by Captain G. H. Wilson, U. S. Marine Corps; May 18, 1943.

MONTREAL

Inspection of Training Facilities at No. 1 Wireless School; May 19, 1943.

NEW YORK

Get-Together Dinner and Smoker-Meeting with F. E. Butler, Guest Speaker; June 2, 1943.

PITTSBURGH

"Tin Flow by Radio-Frequency Heating," by C. J. Madsen, Westinghouse Electric and Manufacturing Company; May 10, 1943.

PORTLAND

"Analysis of the Tuned-Plate Oscillator," by R. M. Lichtenstein, Reed College; Paper Discussed by Messrs. Deardorff, Meissner, and Pritchett; April 22, 1943.

"Transmission-Line Theory," by Wilson Pritchett, Oregon State College; Paper Discussed by Messrs. Deardorff and Meissner; May 8, 1943.

(*Continued on page 30A*)

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SECTION MEETINGS

(*Continued from page 28A*)

"Practical Aspects of the Doherty Amplifier," by J. L. Corl, Oregon State College; Paper Discussed by Messrs. Bookwalter, Singleton, and Feikert; May 8, 1943.

ST. LOUIS

"Interference in the Presence of Strong Radio-Frequency Signals," by A. J. Ebel, Radio Station WILL; May 28, 1943.

SAN FRANCISCO

"A War-Time Training Program for Radio Technicians," by Lt. Commander P. J. Fritschel, U. S. Navy Radio Material School; April 22, 1943.

TWIN CITIES

"Ignition Research at Northwest Airlines," by Carl Swanson, Northwest Airlines; May 19, 1943. Election of Officers; May 19, 1943.

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(*Continued on page 32A*)

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(*Continued from page 30A*)

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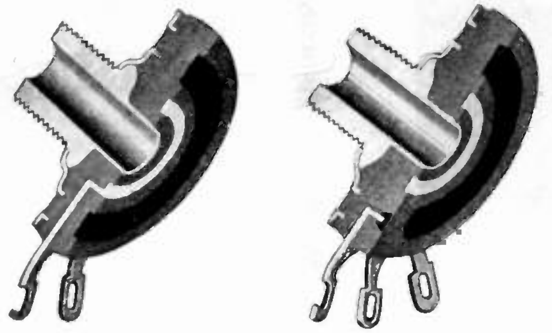
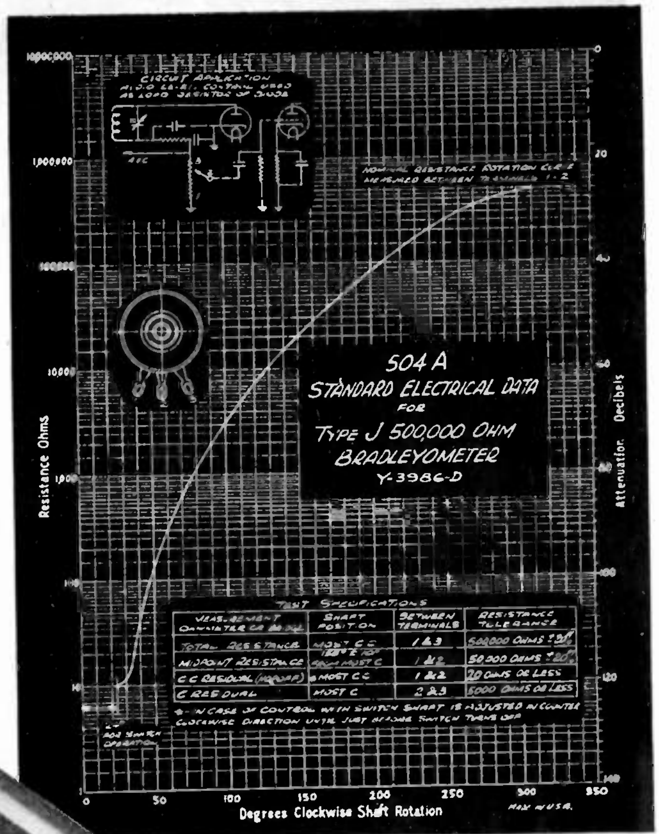
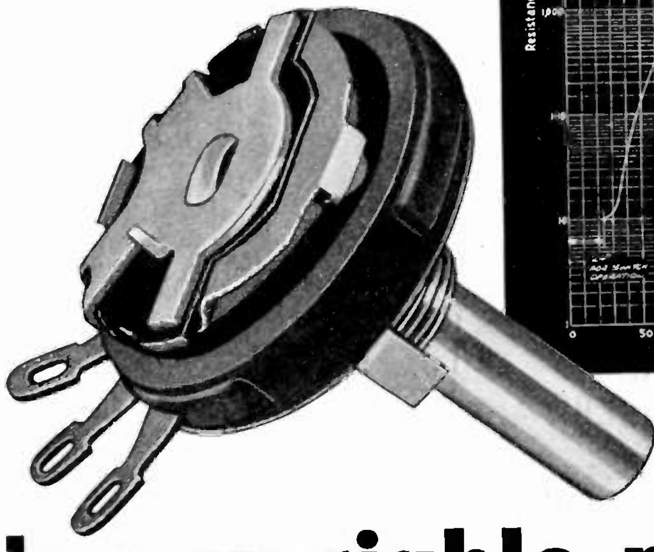
(*Continued on page 34A*)



Cross-section views of Type J Bradleyometers showing how terminals are connected to the solid, molded resistor element.





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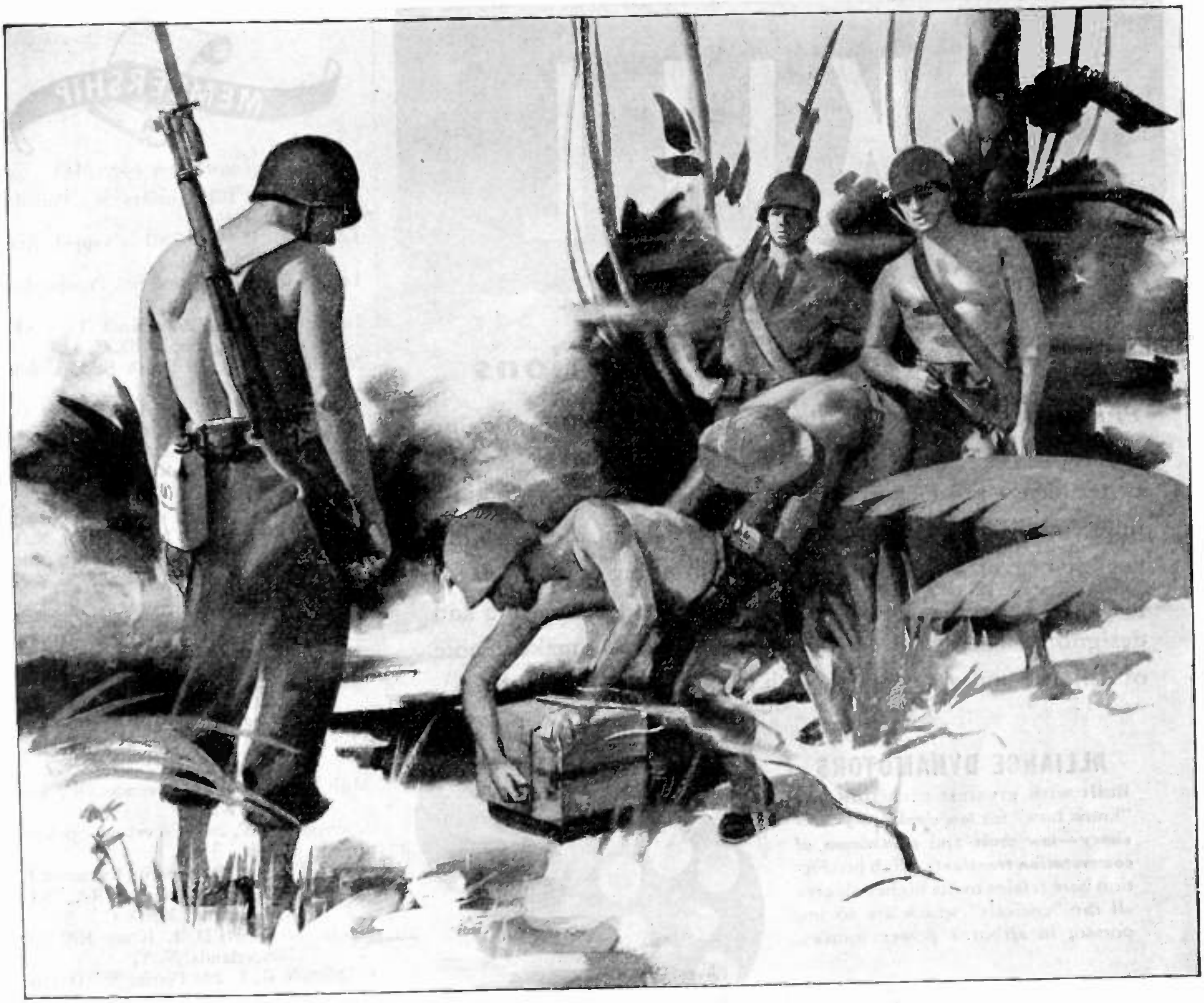

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White, M. W., Naval Radio Station Dupont, Charleston, S. C.
Wilewski, H. M., 67 Stockman Pl., Irvington, N. J.
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Telephone Development: Graduate electrical engineer needed; good theoretical knowledge of acoustics desirable; some experience with magnetic circuits, acoustical measurements.

Laboratory Engineer: For laboratory measurements and tests on a-f communications apparatus, a graduate electrical engineer. To design and supervise equipment for production testing; must maintain laboratory standards. Familiarity with government specifications and inspections procedure desirable.

Communications-Signaling: Design engineer familiar with application and molding technique of plastics. Should have several years of design experience on small electrical apparatus; understanding of magnetic and electrical circuits desirable.

Positions concerned now with war contracts and planning postwar developments. Those now employed at highest skill in war industry need not apply. Communicate with Chief Engineer, giving background and salary requirements. Write, phone or wire to Great American Industries, Inc., 70 Britannia Street, Meriden, Conn.

PATENT ATTORNEYS

Patent attorneys, who are electronic physicists and electrical or radio engineering graduates who have maintained contact with the field of high-frequency electronics, radio manufacture, carrier-current telephony, and light-current circuit design and computing, can make a substantial contribution in research or development jobs with one of the National Defense Research Committee laboratories located in the East. The project is secret but is one of the most urgent of all research jobs now under way for the Government.

An electrical engineering background in light currents is essential, and amateur radio experience, inventive ability and ingenuity in the design and layout of radio equipment would be of considerable help.

Facilities for specialized refresher training and orientation in the particular field may be available. Anyone who possesses these qualifications and is interested in a vital wartime development job for the duration may get further details on request. All inquiries will be held confidential. Address Box 299.

AUDIO ENGINEER OR EXPERIMENTAL PHYSICIST

Work: Design of microphones, earphones, vacuum-tube hearing aids, general audio-development work. Essential war work and manufacturing on peacetime products.

Experience Preferred: Some electroacoustical design and manufacturing with enough general audio experience to be useful when there are no acoustical problems on hand. College degree. Over five years experience along these lines.

Would also consider a man with moderately broad audio-frequency design and experimental experience, good technical education, and good practical common sense. Over three years of good experience along these lines.

Location: Minneapolis, Minnesota.

Salary: Depends entirely on qualifications.

Interview: Will pay traveling expense of qualified applicants. Address letter of application to Box 300.

PATENT ATTORNEY

Patent attorney to join small Patent Department of mid-western concern engaged in development and production of electronic and allied equipment. Experience in electronics, acoustics or sound-recording field highly desirable. Salary open. Candidate must be citizen with proof of same. Write Personnel Director, The Brush Development Company, 3311 Perkins Avenue, Cleveland, Ohio.

(Continued on page 42A)

POSITIONS OPEN

(Continued from page 40A)

ELECTRONIC ENGINEER

Electronic engineer with M.A., Ph.D., or the equivalent in physics, for research and design in geophysics. Experience in filter design and sound recording is desirable. Write to Independent Exploration Company, 901 Esperson Building, Houston, Texas.

ENGINEER

AAA-1 eastern manufacturer, over 75 years operation and leader in growing industry, has immediate permanent position for chemical, electrical, electronic or chemical-metallurgical engineer to organize and increase efficiency of production activities. Send full details of experience to Box 301.

RADIO OR ELECTRICAL ENGINEER

Capacitor manufacturer located in New Bedford, Mass., wants an electrical or radio engineer—man or woman—for equipment- and circuit-development work.

Permanent postwar future for right person. This firm has excellent laboratory facilities and is a leader in its field.

Applicant should be college graduate with a degree—or equivalent experience—in radio engineering or electrical engineering.

Interview in Boston, New Bedford or New York can be arranged. Traveling expenses paid to place of interview.

Write fully, giving age, education, experience, etc. Address reply to Box 302.

RADIO ENGINEERS

Well-established international corporation, 100% in war work with definite postwar possibilities, needs several radio engineers who are familiar with the construction or use of automatic-receiving equipment. Also two transmitter engineers familiar with 40 k.w. equipment.

Applicants should have college degree or approximately ten years experience in radio. Openings in Chicago and New York. Salaries from $100.00 a week dependent upon experience and ability.

In reply please give complete details of experience, age, education, present and former employers, present earnings and your telephone number. Enclose recent photo if available. Address reply to Box 303.

RADIO ENGINEERS

Qualifications should consist of a Class A "ham" license, good electrical background, and knowledge of building your own transmitter, receiver and test equipment.

Positions available in laboratory, testing and design work with a mid-west manufacturer engaged 100 per cent in production of war equipment. "Ham's Paradise" with annual salary ranging from $3000 to $5000.

Applicants, not not employed at bighest skill and able to obtain release, are requested to send full data including draft status, salary required, and a small photo (whicb will not be returned). Address Box 295.

ENGINEER

Electrical or electrocbemical engineer, thoroughly versed in the theory of liquid and solid dielectrics, to direct a research and development program in the development and application of dielectrics to capacitors. This is an unusual opportunity for a capable engineer interested in his present and post-war future. Write to Industrial Condenser Corporation, 1725 W. North Avenue, Chicago, Ill.

ENGINEERING RESEARCH DIRECTOR

A leading manufacturer of small compact electrical parts, widely used in communication equipment and industry in general, seeks engineer to direct its research activities leading to development of new products.

Qualifications must include thorough knowledge of fundamental physical sciences, ability to do and direct research work, resourcefulness, originality, and a sufficiently practical slant to conserve time and energy.

Interesting work. Remuneration depending on the individual. Location in eastern Pennsylvania.

In reply, state complete qualifications which will be kept confidential. Those engaged in war work cannot be considered. Address Box 296.

(Continued on page 46A)


PERFECTION
from a
HUNDRED PAIRS
OF HANDS
The "human element" so consistently mentioned as a necessary part of every great accomplishment, is exemplified in the high quality of Thordarson transformers. From engineering laboratory to final inspection, countless pairs of hands and minds work in harmony to bring about in each complete transformer a coordination of effort that makes for superb performance.
ARMY E NAVY
THORDARSON
ELECTRIC MFG. COMPANY
500 WEST HURON ST., CHICAGO, ILL.
Transformer Specialists Since 1895
. . . ORIGINATORS OF TRU-FIDELITY AMPLIFIERS

POSITIONS OPEN

(*Continued from page 42A*)

ELECTROACOUSTIC ENGINEER

One experienced in laboratory development and research activities. New position open in the development of war-production items. Situation includes excellent opportunities for post-war employment. Salary open. Write Personnel Manager, Universal Microphone Co., Box 299, Inglewood, California.

ELECTRICAL ENGINEERS—PHYSICISTS

Radio or ultra-high frequency experience desirable but not essential. Interesting radio tube development work. Good location, Lancaster, Pennsylvania. Address replies to Personnel Dept., RCA, Victor Division, Harrison, N.J. If you are using your full skill, full time on war work, please do not apply.

ENGINEERS—PHYSICISTS

Radio engineers, electronic engineers, electrical engineers, physicists. A non-profit research laboratory engaged in urgent war research must increase its scientific staff. Men or women (college graduates) with experience in vacuum tube circuit design, construction of aircraft radio equipment and design of small electromechanical devices are needed. Salaries range from \$3,000 to \$8,000 depending upon experience, ability, education and past earnings.

Apply by mail to Airborne Instruments Laboratory, Columbia University Division of War Research, Box 231, Mineola, N.Y.

RADIO ENGINEERS AND TECHNICIANS

A progressive company with a sound background in radio and electronics needs, at once, several men with training and experience in any phase of the radio industry. The work open is vital to the war effort but offers a promising post-war future for the right men. College degree or equivalent experience necessary. Men now engaged at highest skill on war production should not apply. Write Box 294.

RADIO ENGINEER

Wanted for general work covering design and installation with major communication company chiefly in New York area. Technical graduate preferred. Must have car. Position not limited to duration. Address Box 298.

RADIO ENGINEERS AND TECHNICIANS

In critical war industry. Opportunity for several competent men in research and production engineering on Government contracts. Work is with a company well known in the radio industry, located in a Michigan city. Send full particulars of your experience and photo. Address Box 284.

RADIO ENGINEER

Experienced in the manufacture and testing of ultra-high-frequency apparatus; must be capable of taking complete charge of war projects. Splendid opportunity. War workers at highest skill need not apply.

Inquiries will be kept confidential. Please state age, experience, and salary expected. Write Box 288.

RADIO INSTRUCTION-BOOK WRITERS

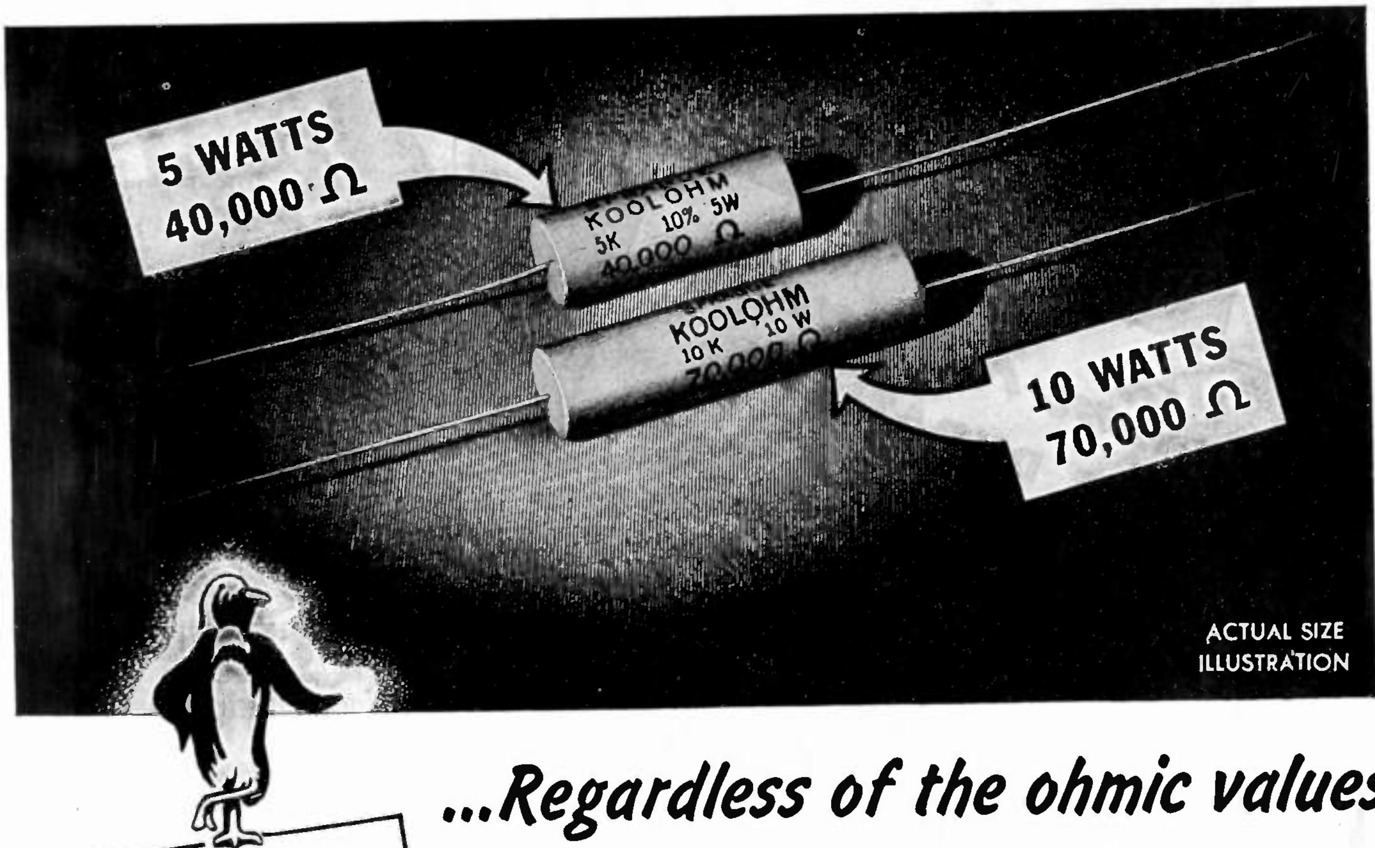
Thorough knowledge of radio principles and ability to describe in simple terms the operation of UHF circuits required. Work relates to instruction books for electrical apparatus. Excellent opportunity to do essential work in very essential war industry. Salary depends upon qualifications and experience.

Replies solicited from Engineers, Patent Attorneys, Teachers and others qualified.

Write complete qualifications and salary desired to Hazeltine Electronics Corporation. 1775 Broadway, New York, N.Y.

(*Continued on page 56A*)

THESE RESISTOR WATTAGE RATINGS MEAN WHAT THEY SAY



...Regardless of the ohmic values

KOOLOHM WIRE IS 2¼ TIMES LARGER

Although Koolohm Resistors themselves are no larger—often smaller—than ordinary resistors, their wire is 2¼ times larger in actual cross-sectional area! This is especially important in the high resistance values where very fine wires so often fail to carry the currents involved in power resistor operation, or to withstand the corrosive effects of operation in high humidities *Play safe* with Koolohms

If it's a 5-watt Koolohm, use it at its full 5-watt rating—regardless of whether it has a 1 ohm or a 40,000 ohm value! If it's a 10-watt Koolohm you can count on it dissipating a full 10 watts whether the resistance value is 1 ohm or 70,000 ohms!

In brief, there's no need to "play safe" with Koolohms. You don't have to use a larger resistor than you actually require. You can forget your worries as to whether the wire size is big enough to carry the current and the resistor body large enough to withstand the temperature rise involved. *You can use any Koolohm at its full wattage rating—any time, anywhere!*

This freedom of use is made possible because Koolohm design is based upon a time-tested, inorganic insulating material. This is sintered on the wire *before it is wound*—at 1000° C.! The insulation is flexible, and has a dielectric strength of 350 volts per mil at 400° C.!

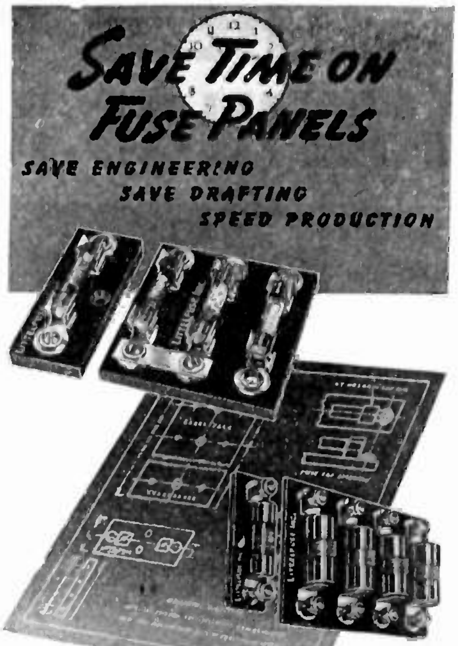
Samples free to industrial users. Catalog on request to all who are interested in better, more dependable resistors.

SPRAGUE SPECIALTIES CO., Resistor Division, North Adams, Mass.

KOOLOHM

REGISTERED TRADEMARK

POWER WIRE WOUND RESISTORS AND METER MULTIPLIERS



Send NOW for New Blueprint of

LITTELFUSE

***Universal* FUSE PANEL No. 1505**

Blueprint comes to you without charge. With it you can quickly designate the Fuse Panel for your requirements—exactly to your specifications. One number covers your mounting. #1505 specifies Standard Panel Mounting: first dash number, size of fuse required; second dash number, the number of poles required. If bus bars are required, specify separately, and specify poles to be bussed.

READY TO MOUNT

These strong light panels are equipped with terminals and Beryllium Copper Fuse Clips, or terminal studs. They meet all Air Corps requirements. They insure the utmost in durability.

MANUFACTURERS, ENGINEERS, DRAFTSMEN, PRIME CONTRACTORS, PURCHASING AGENTS:

Here is important aid in saving hours of time and labor, speeding production. Send for as many copies as you need of new B/P for Littelfuse Universal Fuse Panel No. 1505. Special panels to your specifications made promptly.

LITTELFUSE
INCORPORATED

4782 Ravenswood Ave.,
CHICAGO, ILLINOIS

242 Ong Street
EL MONTE (Los Angeles suburb)
CALIFORNIA

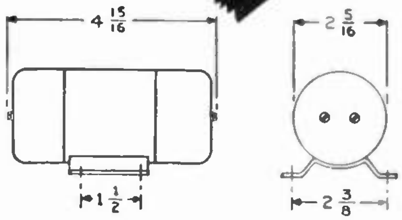
GIVE HIM A BREAK!

Evening is about his only chance to telephone home. He can get through easier if the wires aren't crowded—and his calls mean so much to him and the home folks. So please don't call Long Distance between 7 P. M. and 10 P. M. unless your calls are really necessary. Many thanks.

BELL
TELEPHONE
SYSTEM

Now—A Tiny Power Supply Unit Weighing Only 34 Ounces

EICOR



OUTPUT		INPUT		EICOR PART NO.
VOLTS	AMPS	VOLTS	AMPS	
200	.050	28 14	1.0 2.0	2316-21 2316-22
150	.067	28 14	1.0 2.0	2316-23 2316-24
100	.100	28 14	1.0 2.0	2316-25 2316-26

Continuous duty. 50°C temperature rise. Regulation 20% from no load to full load.

Here is Eicor's answer to your need for a power supply that is much smaller, much lighter, and completely dependable. This tiny Dynamotor is now available to manufacturers of electronic equipment for critical applications where space and weight requirements are of utmost importance.

SAMPLES AVAILABLE

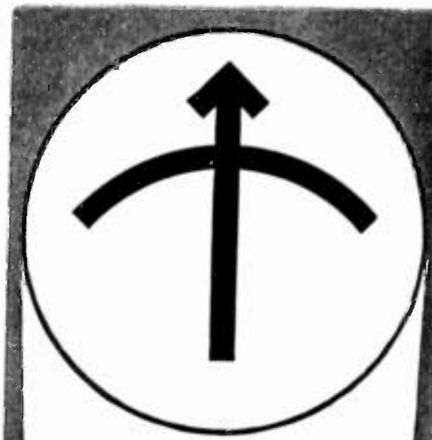
Our specialized experience can be of help to you. Samples of this exclusive Eicor product in the types listed at left furnished quickly for development purposes on priority order.

Write, wire or phone

EICOR INC. 1501 W. Congress St., Chicago, U.S.A.

DYNAMOTORS • D. C. MOTORS • POWER PLANTS • CONVERTERS

Export: Ad Auriema, 89 Broad St., New York, U. S. A. Cable: Auriema, New York



LABORATORY STANDARDS

Standard Signal Generators
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Square Wave Generators
•
Vacuum Tube Voltmeters
•
U. H. F. Noisemeters
•
Pulse Generators
•
Moisture Meters
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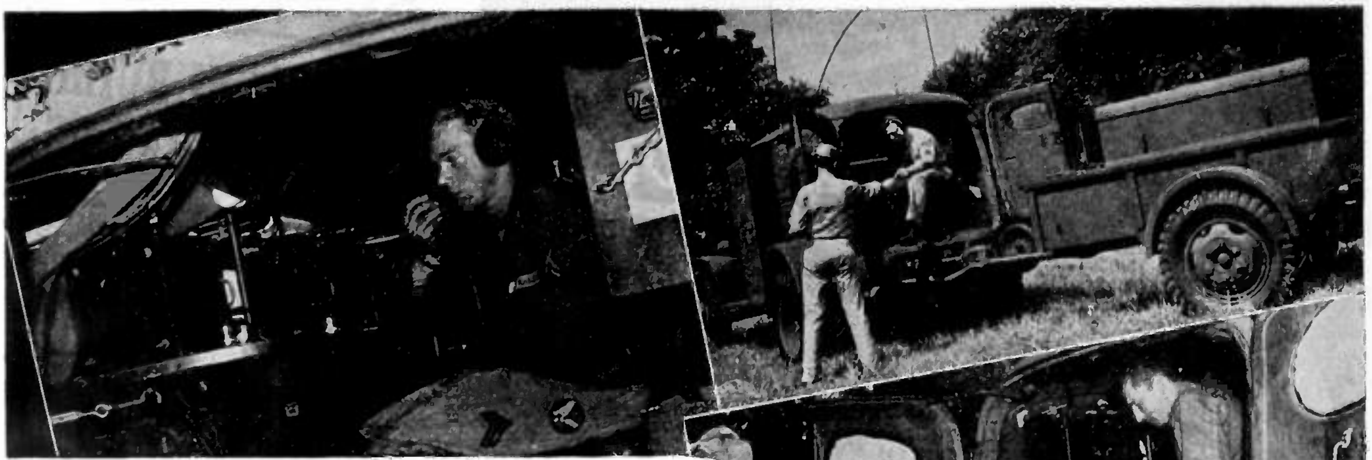
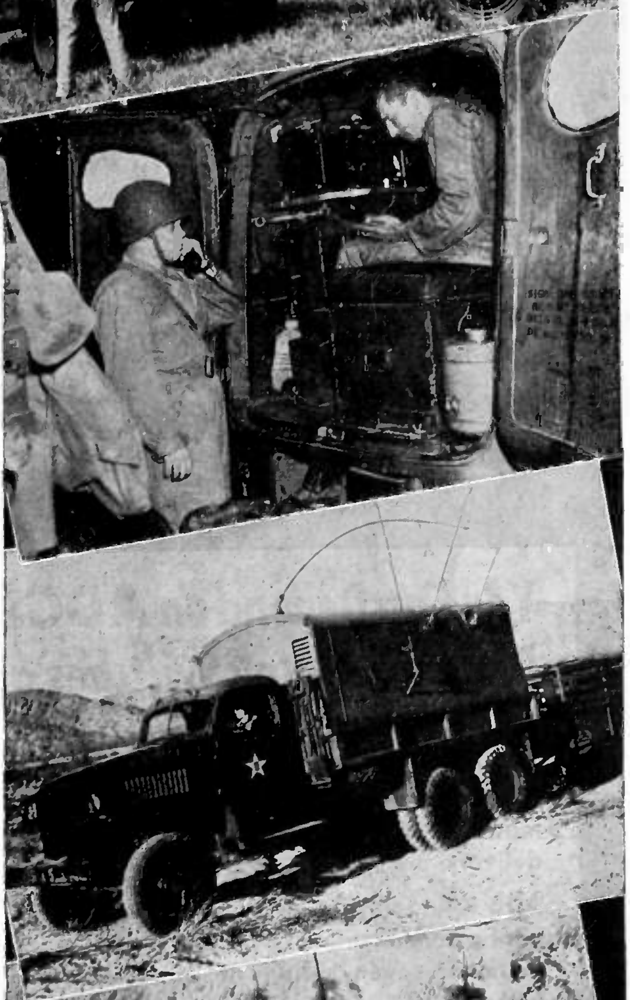
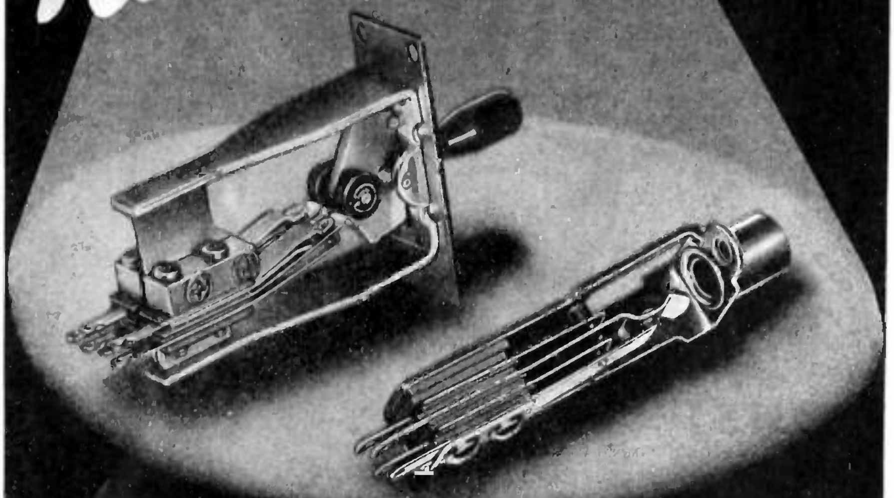
MEASUREMENTS CORPORATION

Boonton, New Jersey





Need them Now?



You Can Depend on ADC for Quick Delivery on Key Switches and Jacks

Audio Development's reputation for precision quality keeps orders coming—expanded facilities send them out in a hurry. We offer immediate service on standard units made from stock parts to your specifications—prompt delivery on special types.

ADC Key Switches assure quiet, dependable operation in vital communication circuits. Standard types allow a maximum of seven springs in each quadrant, providing a wide variety of locking and non-locking switching combinations. Silver alloy contacts are standard—special contact materials can be supplied if desired. Available with or without mounting plates.

ADC Jacks are of approved welded box construction assuring rigid alignment of all parts. Non-aging springs provide permanent, proper tension. Additional springs allow for switching of auxiliary circuits. Available for use with all standard two and three circuit telephone type plugs.

Send us your detailed specifications and requirements. We are equipped to serve you promptly and efficiently.

ADC

Audio Development Co.

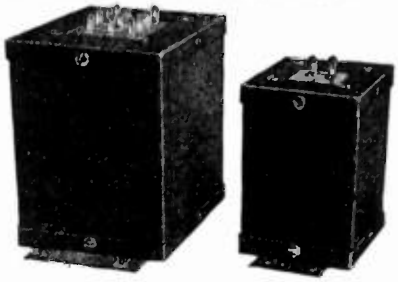
2833 13th Ave. S., Minneapolis, Minn.

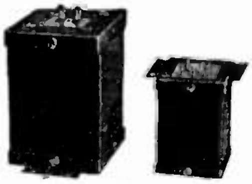
ACME

PRECISION-BUILT TRANSFORMERS

FOR ELECTRONIC PERFORMANCE

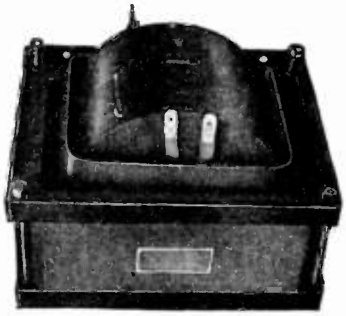
Controlling electrons to a useful purpose requires transformers of exact performance characteristics. Acme precision-built transformers for electronic applications, when submitted to unbiased tests, invariably win top honors for performance. If your electronic application is out of the ordinary, let Acme transformer engineers help in its solution.



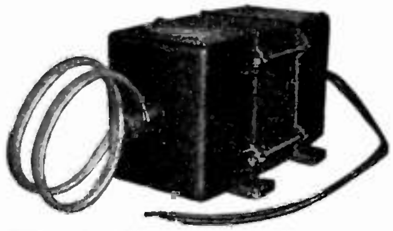


FOR EXAMPLE

Acme compound-filled transformers for short wave communication, public address systems and other radio applications are preferred for their serviceability under temperature variations from —40° to +120°.



And preferred for rugged construction, trouble-free long-life. Typical, high voltage plate supply transformer for transmitter. 33,000 volts, 1.8 ampere secondary.



ISOLATING TRANSFORMERS

For use wherever radio, communication, or other electrical equipment must be tested with complete freedom from outside interference. Shielded secondary winding and shielded secondary cable isolate primary fluctuations and interference. Write for details.

THE ACME ELECTRIC & MFG. CO.
44 WATER ST. CUBA, N.Y.

Acme ACME Electric
TRANSFORMERS

POSITIONS OPEN

(Continued from page 46A)

RADIO ENGINEER

Mathematical knowledge to include trigonometry and elementary calculus. Must be familiar with all types of circuits for reception and transmission on frequencies up to 200 mc/s. Knowledge of transmission lines, aerial arrays, television and cathode-ray-tube technique desirable. Must also have knowledge of radio test equipment and must understand any circuit diagram.

Prefer man conversant with manufacture, repair and fault location on all types of transmitters and receivers. Address Box 289.

RADIO ENGINEER

EDUCATION: Minimum of two years college in Electrical Engineering.

EXPERIENCE: Minimum of two years in radio test or engineering, or five years in electrical control work (power station or telephone central-office wiring, etc.).

Must be of a type qualified to interpret and clarify with inspectors and responsible executives electrical specifications, problems of manufacture, test and inspection. Address Box 290.

RADIO ENGINEER

Opening in engineering department of concern making communication apparatus. Prefer college graduate with ultra-high-frequency experience, but lack of experience in this field will not bar an adaptable man.

Duties involve developing, designing and carrying the product through the shop with a minimum of supervision. If inexperienced, the duties would be the same under supervision. The company is engaged in war activity exclusively and expects to continue this work after the war.

Salary $2,500 to $5,000 depending upon qualifications. Pleasant surroundings and working conditions. If now employed in war work, a release from the employer must be obtainable. Address: Chief Engineer, Templetone Radio Company, Mystic, Conn.

INSTRUCTORS IN ADVANCED ARMY-NAVY PROGRAM

Prominent Eastern technical institute needs additional instructors in officer training program in modern electronics and radio applications. An excellent opportunity to acquire advanced knowledge and to render important service in war effort. Men having various degrees of qualifications are needed, from recent graduates in Electrical Engineering or Physics to those with long experience in radio engineering or teaching. Salary according to qualifications and experience. Applicants must be U. S. citizens of unimpeachable reputation. Any inquiries will be treated as highly confidential. Please send personnel data and photograph to Box 292.

RADIO AND ELECTRONIC ENGINEERS

First, we are seeking the services of one or two trained engineers who have had ample experience in electronic engineering. The men selected will not only be concerned with present war production, but should eventually develop key positions in postwar operation.

Second, we are also looking for a few young engineers who have had good schooling and background to be trained for specialized work with us.

This is an excellent opportunity for men who qualify to connect with a progressive, highly regarded manufacturer of transmitting tubes. Many special benefits will be enjoyed in your association with this company.

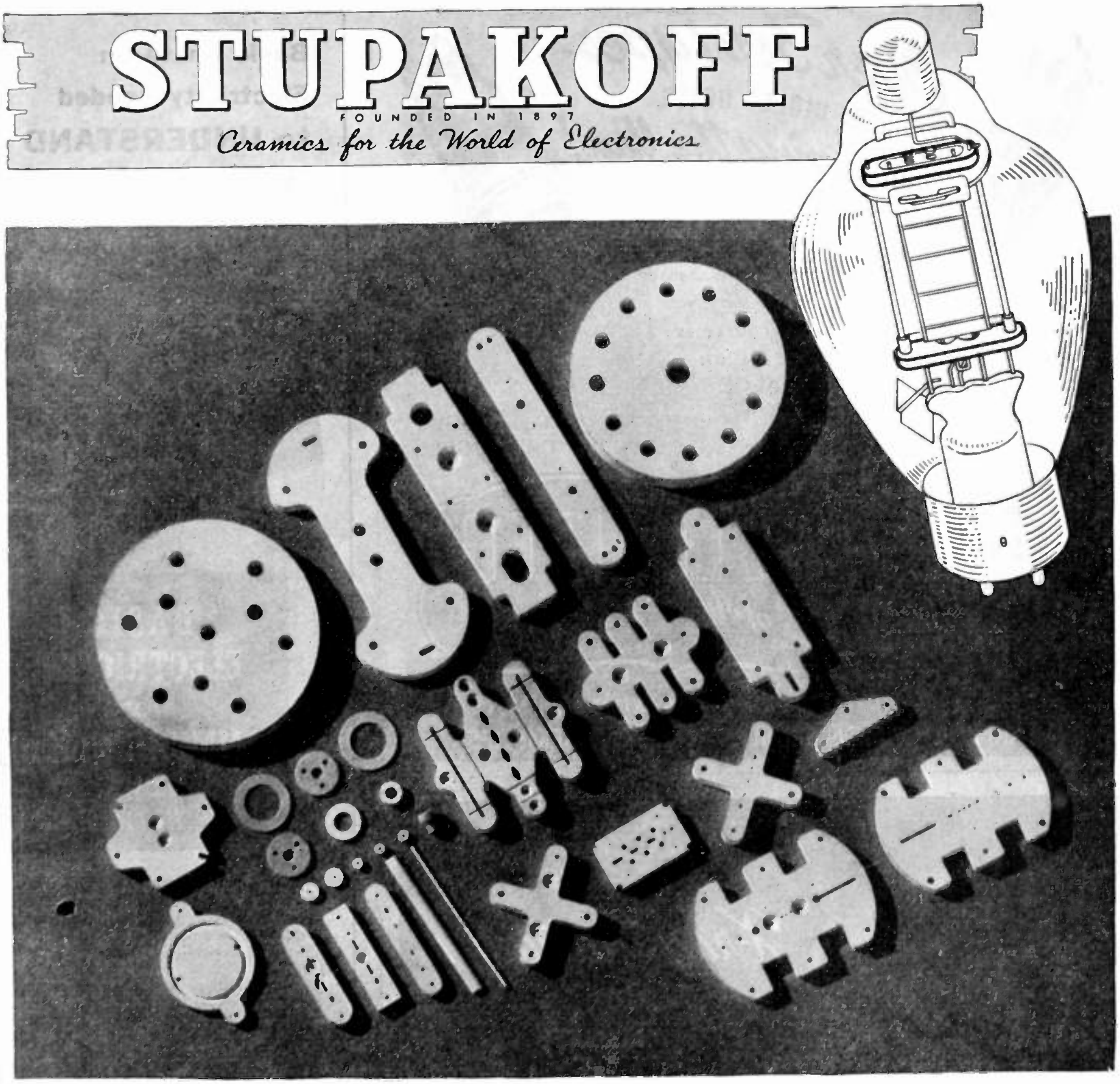
Write at once giving complete details of past experience. Interviews will be promptly arranged. Persons in war work or essential activity not considered without statement of availability. Chief Engineer, United Electronics Company, 42 Spring Street, Newark, New Jersey.

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

STUPAKOFF

FOUNDED IN 1897

Ceramics for the World of Electronics



20 Years' Experience...

ENGINEERING AND MANUFACTURING INSULATORS FOR *Radio Tubes*

STUPAKOFF ceramics were used as heater insulators in the first A. C. radio tube. In the years following we have been directly associated with practically every change and improvement and have developed many materials which have contributed to the progress of the radio tube industry.

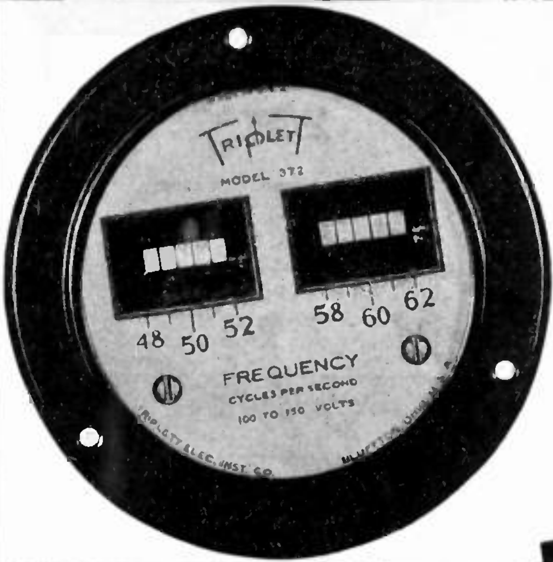
Insulators vary in specifications, depending upon their application. We have the experience necessary to prescribe the correct material for a given application and the manufacturing facilities to produce extremely large quantities of precision made ceramic insulators to your specifications.

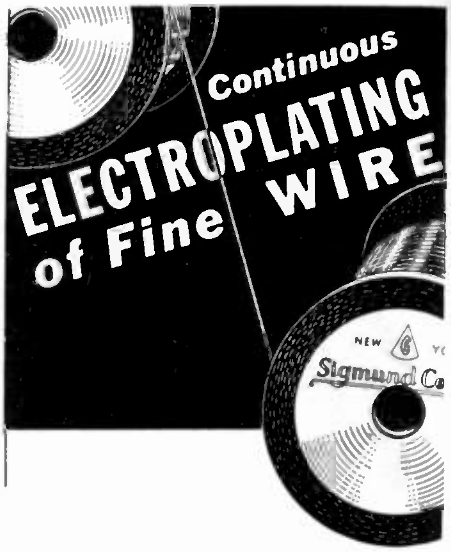
ARMY E NAVY

"for great achievement"

STUPAKOFF
Steatite
INSULATORS

STUPAKOFF CERAMIC AND MANUFACTURING CO., LATROBE, PA.





LITERATURE

(*Continued from page 26A*)

minimum. Ths second edition is really a new book brought up to the minute with current radio circuits. It is a teaching text book. Price $3.50.

Two Beitman Service Books

By M. N. Beitman, B.S., Radio Instructor, Englewood High School • • • *Supreme Publications, 328 S. Jefferson St., Chicago. Ill.* SIMPLIFIED RADIO SERVICING BY COMPARISON METHOD. (8½×11 inches, 108 pages+covers, paper bound.) Subtitled "By Comparison Method" the author has developed many time saving ways to check for faults in receiver parts. The book contains the "Sylvania Radio Tube Characteristics" tables and diagrams and 16 trouble-shooting blue-prints of circuits and over 1000 hints on radio servicing Price $1.50.

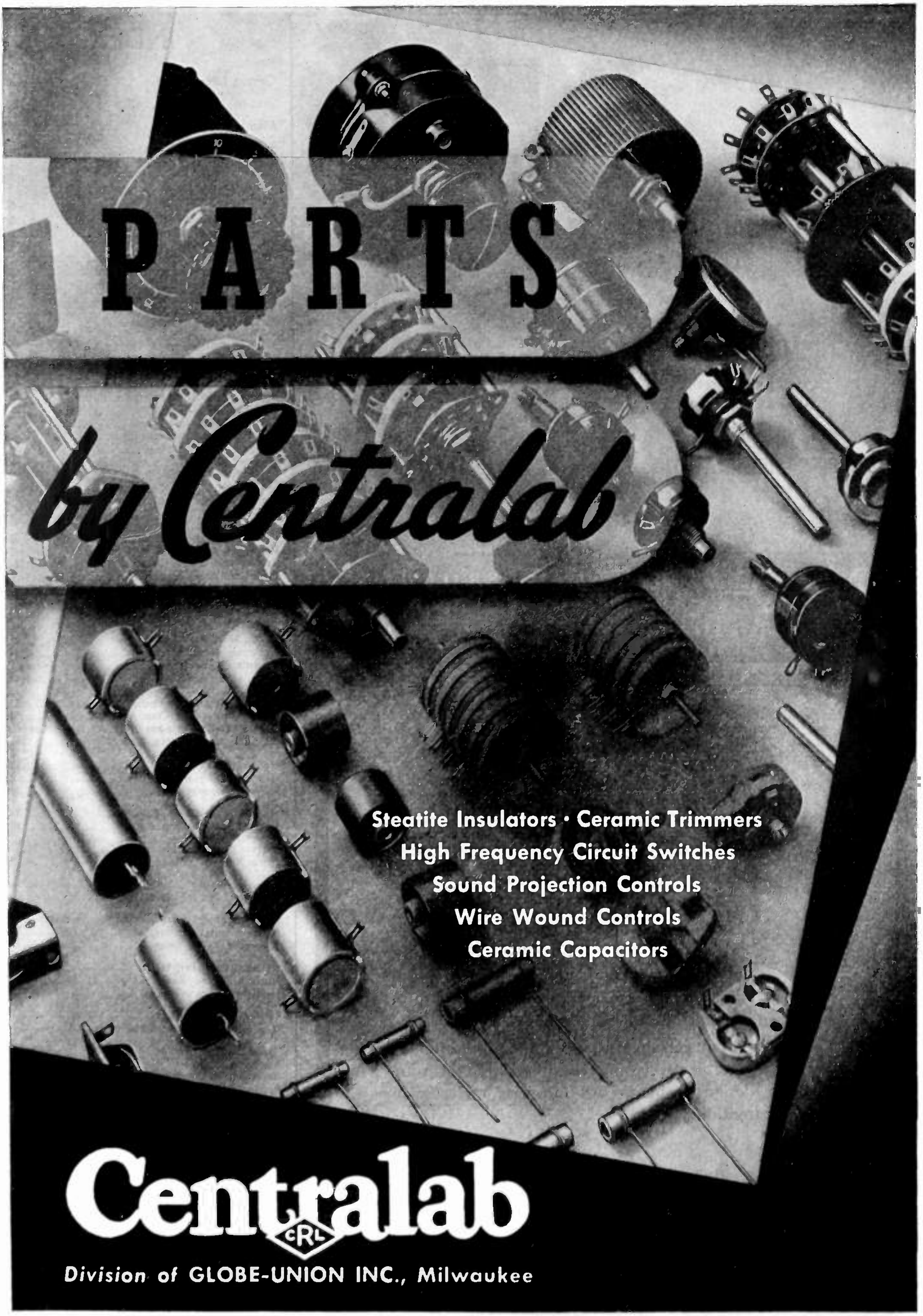
HOW TO MODERNIZE RADIOS FOR PROFIT (8½×11 inches, 32 pages +covers.) The author gives practical job sheets, schematics and photographs to show ways to improve older radio sets. Price $1.00.

Two "Most-Often-Needed" Books

By M. N. Bietman, B.S., Radio Instructor, Englewood High School • • • *Supreme Publications, 328 S. Jefferson St., Chicago, Ill.* MOST-OFTEN-NEEDED GENERAL ELECTRIC RADIO DIAGRAMS. (5½×8½ inches, 64 pages, paper bound.) Gives circuits and information on 41 General Electric Radio Receivers manufactured from 1935 to 1941.

MOST-OFTEN-NEEDED 1942 RADIO DIAGRAMS. (8½×11 inches, 192 pages, paper bound.) This volume, number 5 of an annual series gives circuit diagrams for from one to thirty 1942 models for 42 radio receiver manufacturers together with alignment data, service hints and replacement parts data for service men. Price $2.00.

*FIRST PRINCIPLES OF RADIO COMMUNICATIONS. By Alfred P. Morgan • • • *D. Appleton-Century Co., 35 W. 32nd St., New York, N. Y.* (6×8½ inches, ix+366 pages, 183 figures, cloth bound.) This book is designed for a simple teaching and studying text for beginners in radio. It starts with basic electrical principles and unfolds each subject to give a complete basic understanding of the principles that make each part of a transmitter and receiver work. Mathematics has been almost entirely avoided. A few important electrical calculations are explained but they require only ordinary arithmetic or simple algebra. The book is designed principally for usefulness and training men for the armed forces.


PARTS
by Centralab
Steatite Insulators · Ceramic Trimmers
High Frequency Circuit Switches
Sound Projection Controls
Wire Wound Controls
Ceramic Capacitors
Centralab
CRL
Division of GLOBE-UNION INC., Milwaukee

INDEX

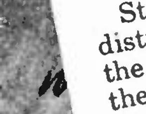
DISPLAY ADVERTISERS

lectronic briefs: FM

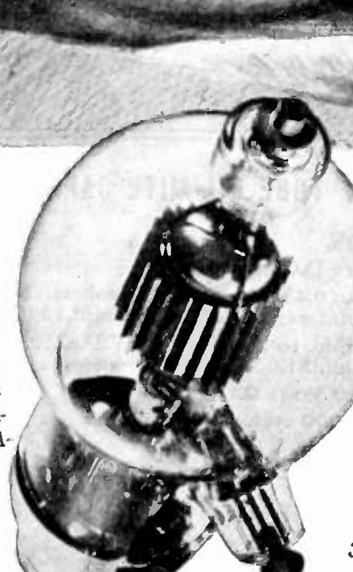
adio is simply a method by which lectrical energy is transmitted through space. By varying the intensity or frequency of this electrical energy, an intelligible signal can be created. The principle is the same whether dot dash code messages or voice and music are being transmitted. In the case of voice and music transmission the radio wave must be varied (modulated) at the same speed as the vibrations of the voice or music. The characteristics of electrical energy which can be varied or modulated are three: voltage, frequency and phase. Radio transmitters which vary the intensity (voltage) are called amplitude modulated and those which vary the frequency are called frequency modulated. The differences of these two systems can be understood easily by visualizing a beam of light. An audible signal can be transmitted by varying the light intensity (amplitude modulation) or by varying the color of the light beam (frequency modulation).

Static and other man-made electrical disturbances are identical in character to the amplitude modulated signal. Hence these disturbances are extremely bothersome to AM broadcasts. On the other hand these electrical disturbances do not essentially vary in frequency and consequently do not interfere with FM transmission. Another fortunate characteristic of FM is the fact that the stronger of two signals predominates, thus eliminating much inter-station interference and cross-talk. Further, and of great importance, the fidelity of tone can be made nearly perfect even when the heaviest of musical scores is being broadcast.

In frequency modulation as in all things in the field of electronics, vacuum tubes are the most important component. Eimac tubes have the distinction of being first choice of most of the leading electronic engineers throughout the world. They are consequently first in the most important new developments in electronics...FM for example.



rmy-Navy "E" ag awarded for igh achievement n the production f war material.



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TUBES

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SAN BRUNO, CALIFORNIA

Export Agents: Frazar & Hansen
301 Clay Street, San Francisco, California

P.O. 645

SEA DOGS

Hats off to the Sea Dogs . . . those intrepid bluejackets who are making naval history! Tales of their courage, endurance and sacrifice will live long after the last gun is silenced and the sea is safe again for all free men.

They're the men who make possible the naval feats that are adding up to victory. They're the men who are at their posts when the cry of "*Battle Stations!*" is heard. They're the men who never lose hope or courage . . . who carry out orders unquestioningly. As one young seaman so aptly expressed it—"They give you a tough job to do, and you do it. That's all."

That's all.

Like the bluejackets on a United States battleship who shot down 32 attacking Jap planes and later sank three Jap cruisers and one destroyer.

Or the seamen on the little U. S. Destroyer Laffey who bravely tackled four Jap warships, and fought until their last gun was silenced.

Then there was the sailor who, lying on the blistering deck of the crippled aircraft carrier Hornet, with one leg broken and the other shattered, tried to climb off his stretcher to have another shot at the Japs.

And the sea dogs who stood unhesitatingly behind Admiral Callaghan as he led the cruiser San Francisco right into a hell of fire between two lines of Jap warships, and then finished off a battleship, a cruiser and a destroyer.

That's our Navy . . . youngsters, most of them . . . the generation the old fellows called "spoiled" and "soft."

That's our Navy . . . every man-jack of them the finest example of American manhood this country has ever produced.

That we may not let them down, Cornell Dubilier is using every minute of its thirty-three years of experience and every facility at its disposal to build capacitors that are absolutely reliable—that can be depended upon to stand up under tough treatment and battle action. We are proud to think that C-D Capacitors, in all types of communication systems, are helping to guide the destinies of America's great fleets.

A Tribute to the American Sailor

CORNELL DUBILIER ELECTRIC CORPORATION, SOUTH PLAINFIELD, N. J.