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ELECTRICAL COMMUNICATION

*Technical Journal of the
International Telephone and Telegraph Corporation
and Associate Companies*

STABLE LABORATORY-STANDARD CAPACITOR MICROPHONE

MANUFACTURE OF TYPE-7400 TELEPHONE SELECTOR

CHARACTERISTICS AND APPLICATIONS OF THERMISTORS

GERMANIUM POSITIVE-GAP DIODE

BROAD-BAND REFLECTOMETERS AT HIGH FREQUENCIES

PULSE GENERATOR FOR MECHANOELECTRONIC SWITCHING SYSTEM

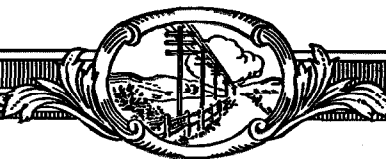
UNITED STATES PATENTS ISSUED TO THE INTERNATIONAL SYSTEM



Volume 32

JUNE, 1955

Number 2



ELECTRICAL COMMUNICATION

*Technical Journal of the
International Telephone and Telegraph Corporation
and Associate Companies*

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Published Quarterly by the
INTERNATIONAL TELEPHONE AND TELEGRAPH CORPORATION
67 BROAD STREET, NEW YORK 4, NEW YORK, U.S.A.

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Volume 32

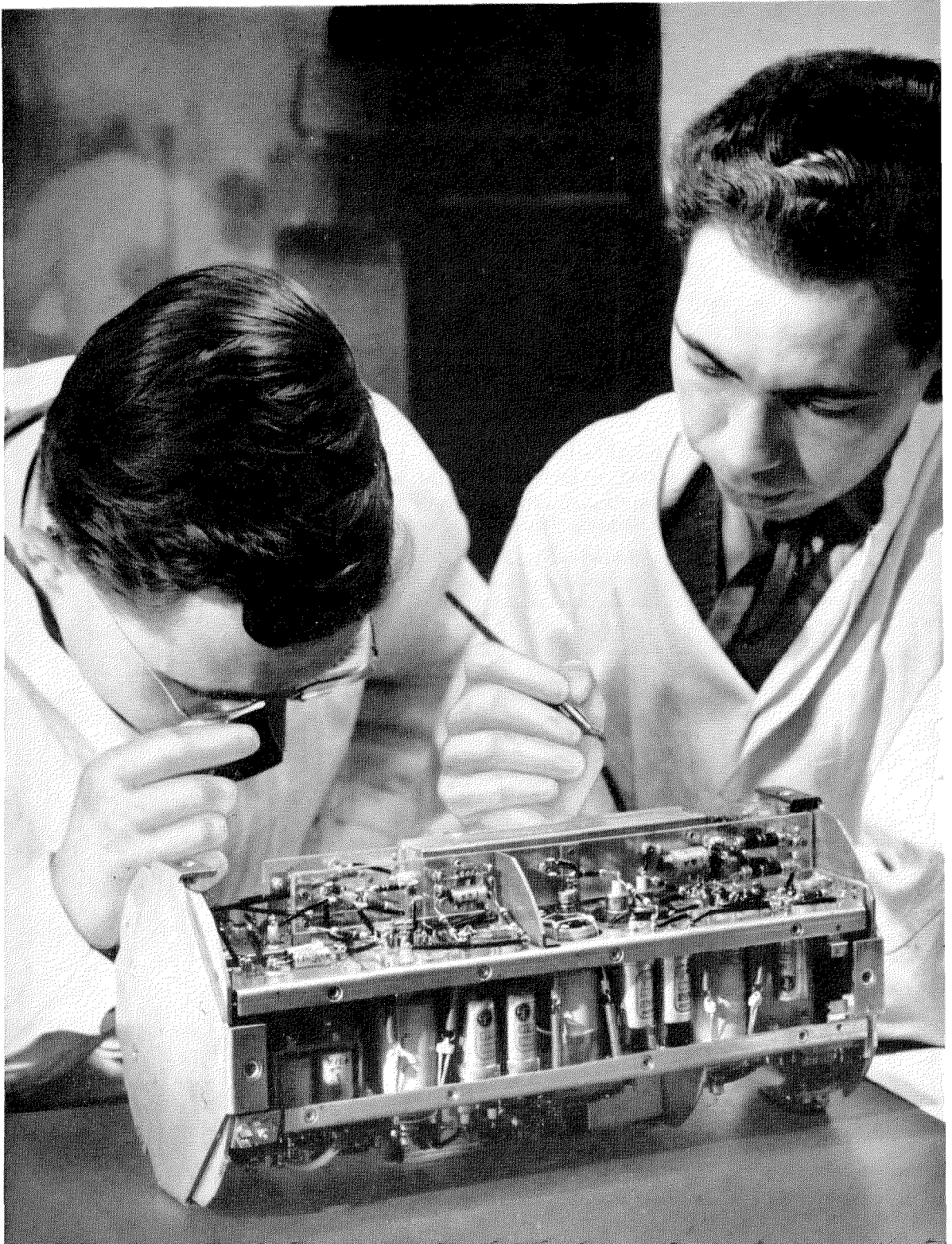
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The amplifier unit of a submerged repeater is shown undergoing inspection after assembly. Equipment similar to the above will be used in submerged repeaters designed jointly by the British Post Office and Standard Telephones and Cables, Limited for use in the Nova Scotia-Newfoundland portion of the forthcoming transatlantic telephone cables. Twenty-one of these repeaters are being manufactured by Standard Telephones and Cables in a special air-conditioned manufacturing unit. Each repeater will carry 60 two-way telephone circuits.

Stable Laboratory-Standard Capacitor Microphone*

By J. F. HOUDEK, JR.

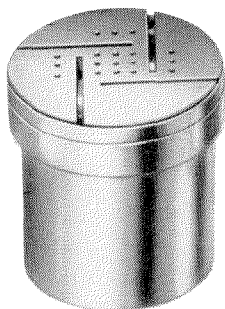
Kellogg Switchboard and Supply Company, a division of International Telephone and Telegraph Corporation; Chicago, Illinois

LABORATORY-STANDARD microphones must provide dependable calibration stability, satisfactory performance, and reliability. Rugged construction together with a spring-loaded rear-electrode assembly permit the maintenance of constant diaphragm tension for improved stability. Appropriate modifications of the acoustic controls provide either a pressure-type or a free-field type of response. Each microphone has uniform characteristics and meets the American Standards Association performance requirements in its own applications. Supporting performance data and other information relative to capacitor microphones are presented.

. . .

The microphone to be described does not differ in principle from the usual precision capacitor microphone in that it is essentially a stretched circular membrane near and parallel to a circular electrode. Since the theoretical aspects of these microphones and the effects of basic acoustic controls on their performance have been extensively treated^{1,2} in the literature, this phase will not be covered here. Historically the development of capacitor microphones in the laboratory of the Kellogg Switchboard and Supply Company was started during 1932. In 1935, two practical designs were in use and as may be seen in Figure 1, they differed in size by about four to one. Both types were used for

more than ten years before preference and development of the art caused the larger to lapse into obsolescence. Subsequently, only the smaller model of this microphone has been manufactured. Microphones of the early designs were incorporated in different types of test equipment where they gave satisfactory performance especially with respect to stability of calibration and adequacy of response-frequency characteristics. Design improvements of a minor nature have been made from time to time as a result of experience in the use and application of the microphone. Beginning with the second world war, Kellogg microphones were made available to other users, such as governmental laboratories, educational institutions, and manufacturing companies.



New capacitor microphone in actual size.

The present preferred method of calibration, the reciprocity method,^{3,4} was not in general use when the microphones were first made. Level calibrations were made by the comparison method, using a calibrated microphone for reference. Response-frequency performance was determined by the electrostatic-actuator method. After developing the reciprocity calibration apparatus shown in Figure 2, it was found that agreement between the two methods was within 0.5 decibel up to and including 13 000 cycles per second. Even after the reciprocity method became available, it was found to be advantageous to use the actuator to obtain relative sensitivities to frequencies extending well above the accurate range of the reciprocity apparatus, since calibrations by the actuator are not limited by frequency and are unaffected by the characteristics of the intervening gas.

* Reprinted from *Journal of the Audio Engineering Society*, volume 2, pages 234-238; October, 1954.

¹ P. M. Morse, "Vibration and Sound," 1st edition, McGraw-Hill Book Company, New York, New York; 1936; pages 160-173.

² T. R. Bonn, "Ultrasonic Condenser Microphone," *Journal of the Acoustical Society of America*, volume 18, pages 496-502; October 1946.

³ L. L. Beranek, "Acoustic Measurements," 1st edition, John Wiley and Sons, New York, New York; 1949; pages 113-148.

⁴ A. L. DiMattia and F. M. Wiener, "On the Absolute Pressure Calibration of Condenser Microphones by the Reciprocity Method," *Journal of the Acoustic Society of America*, volume 18, pages 341-344; October 1946.

More recently, emphasis was placed on free-field comparisons of two microphones, slightly different in contour, to determine and if necessary to eliminate any undesirable characteristics

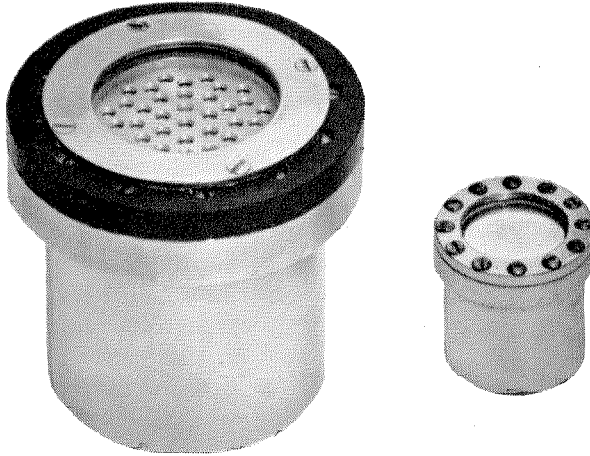


Figure 1—Two designs of microphones in use about 1935.

attributable to the size and shape of a microphone. The tests indicated no important undesirable characteristics and only a slightly greater directivity in the field pattern of the present free-field type over a proposed improved type.

A. The device should faithfully transduce speech and music at the intensities and frequencies normally encountered in audio work. Since it was to be the standard for the company in both the laboratory and the factory, special emphasis was placed on the ruggedness and stability of the unit.

B. The diaphragm size was restricted so as to be compatible with the preferred dimensions of the 6-cubic-centimeter artificial ears and couplers used in telephone receiver work.

C. From the diaphragm diameter so determined the design of the body was developed to be as small as possible, but with ample wall thickness to insure stability of performance. A flange was provided to simplify the mounting and positioning of the microphone with respect to the coupler or cavity with which it was to be used. This flange also added considerable rigidity to the body and further protected the diaphragm against possible strains due to mounting in couplers.

D. For free-field applications, such as the calibration of artificial voices, measurement of sound levels, studio and broadcast pickup, and

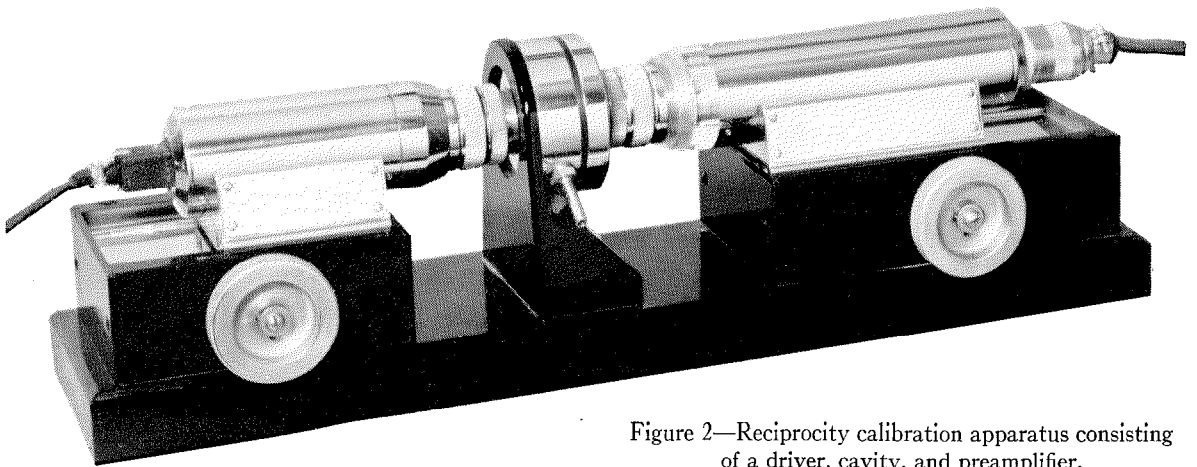


Figure 2—Reciprocity calibration apparatus consisting of a driver, cavity, and preamplifier.

1. Design Considerations

With regard to performance, all of the essential design considerations deemed necessary for the microphone were not specifically stated at the beginning of the development. With time and study, the following requirements were evolved.

the like, it was deemed advantageous to use the same microphone construction and to alter only the acoustic controls to attain more-uniform free-field characteristics.

E. In later stages of the development, commercial

requirements led to higher insulation resistance, an extension of the range of relatively uniform frequency response, and to linearity at higher sound levels. The need for an unquestionably

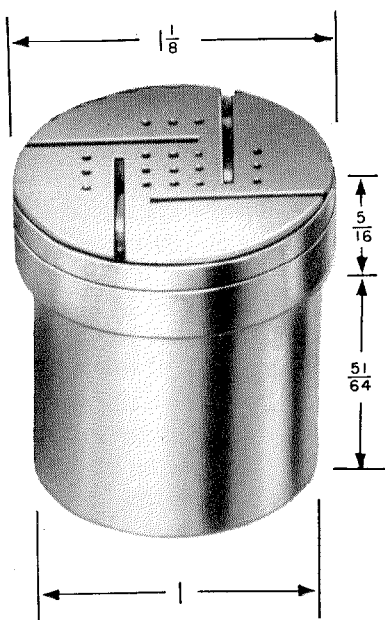


Figure 3—Dimensions in inches of the present model.

accurate method of calibration of the microphone led to the development of the reciprocity calibration apparatus. A minor refinement was required to prevent measurable hydrogen leakage when the microphone is used in its cavity.

F. In the capacitor microphone, a compromise is involved between sensitivity and other performance factors. With emphasis on reliability and lowered distortion, the resulting sensitivity lies slightly below that of the highest-output microphones of this class.

2. Design

Externally, the microphone shown in Figure 3 has an over-all length of $1\frac{1}{8}$ inches (29 millimeters) without the grille. The main body diameter is 1 inch (25 millimeters) and the flange at the front end has a diameter of $1\frac{1}{8}$ inches (29 millimeters). The heavy wall thickness and the flange to which the diaphragm is clamped provide excellent dimensional stability. The microphone is furnished with a removable protective grille, which should not be used in couplers.

Figure 4 illustrates the internal construction and discloses the spring loading used to maintain constant tension on the diaphragm under slight dimensional changes (which might result from temperature variations and/or stretching of the diaphragm). The movable back-electrode assembly is shown well separated from the diaphragm to clarify constructional details. Also evident is the large contact area between the tensioning edge and the diaphragm, which serves as an air seal and obviates the necessity of placing the microphone in a desiccator between periods of use. The resonant cavity, although not clearly shown, is located behind the fixed electrode and is coupled to the cavity of 0.0008 to 0.0012 inch (0.02 to 0.03 millimeter) in height at the rear of the diaphragm by means of 8 intercommunicating holes. The reaction of

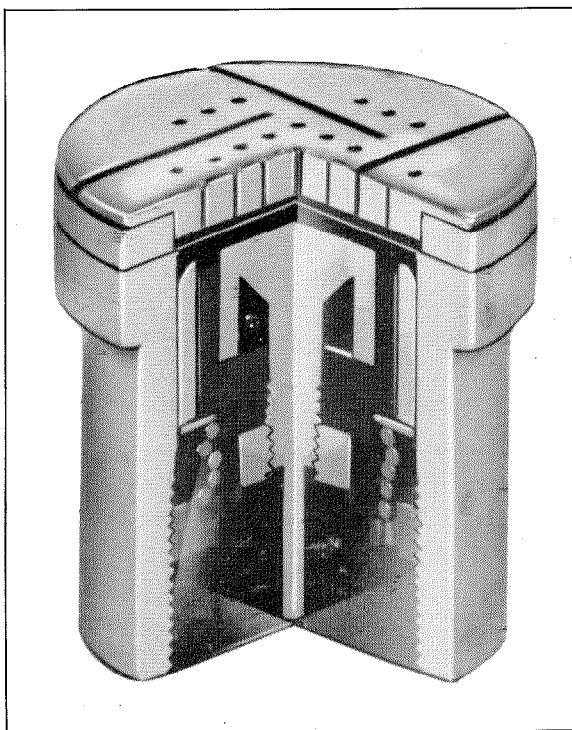


Figure 4—Sectional view of microphone.

this acoustical circuit has been carefully chosen to provide a compromise that flattens the resonance of the diaphragm and extends the frequency range of response.

The microphone has good shock resistance to all forces except those directed approximately perpendicular to the diaphragm. The weight is approximately 80 grams (2.8 ounces) and the unit is mounted by a substantial thread. Pressure should not be applied to the central electrode pin; a sliding-action spring-clip connection as shown in the adapter unit of Figure 5 is recom-

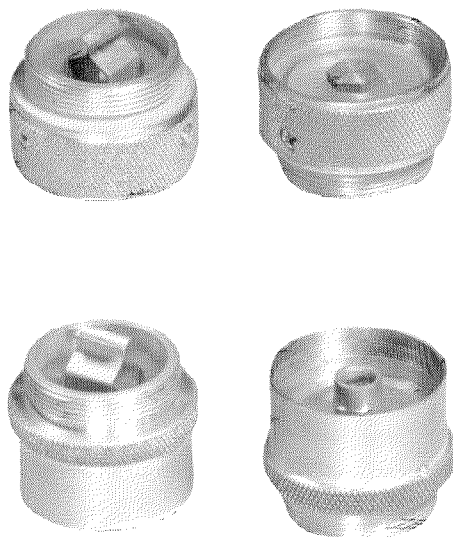


Figure 5—Adapter units.

mended. An adapter unit is used when it is necessary to couple the $\frac{25}{32}$ -inch, 32-thread-per-inch fitting of the microphone to preamplifiers that have a $\frac{29}{32}$ -inch, 60-thread-per-inch coupling.

3. Performance Characteristics

In studies on the capacitor microphone under conditions that were more severe than are encountered in normal use, it was found difficult in many cases to segregate the effects of environmental influences on the associated equipment. At other times, these factors cannot be extracted since the changes occurring in the microphone are of the same order as the accuracy of the testing equipment and it is necessary to express the results in terms of data obtained

from the entire system. It is advantageous to attach the preamplifier directly to the microphone to obtain consistent and reproducible laboratory results.⁵ Since the output impedance of the microphone is high in the audio-frequency range and requires a very-high load impedance, the usual cathode-follower is recommended for audio work. Various commercial preamplifiers are available and adaptable as an extension to the microphone, which then forms the nosepiece. At present in the laboratory, use is made of a modified Cruft cathode-follower-type circuit that is flat from about 1 to over 50 000 cycles.⁶ For free-field work, both the above and smaller preamplifiers made by the Audio Instrument Company have been used.

Typical response-frequency data on type *A* and *B* microphones are summarized in Figures 6 and 7. Referring to Figure 6, the open-circuit voltage response as a function of the frequency of a constant sound source is shown for free-field and cavity conditions. The performance of the type-*A* microphone in cavities, such as that of the reciprocity coupler in an atmosphere of hydrogen, is shown as curve *C*. The response rise of the first resonant mode of vibration of the diaphragm is not removed entirely but by appropriate control, uniformity of response with respect to the low and middle frequencies as a base is improved.

However, if the cavity-type microphone is used in a free field, the remaining resonant-frequency rise plus the diffraction effects to sounds at perpendicular incidence to the microphone diaphragm result in the usual somewhat excessive peak in the 10-kilocycle region. This field pattern is, however, the usually accepted characteristic for a microphone designed for both cavity and free-field applications. Again with reference to curve *C*, it is clearly indicated that the microphone provides a uniform response in cavity work. Also included in this design is the consideration that higher sound intensities are obtained in couplers, especially in tests concerned with overload and high-intensity linearity tests on receivers. Therefore, the type-*A* cavity micro-

⁵ L. L. Beranek, footnote reference 3, pages 221-223.

⁶ United States Office of Scientific Research and Development, "Report 3105 on Sound Pressure Meter"; 1944: pages 155-170.

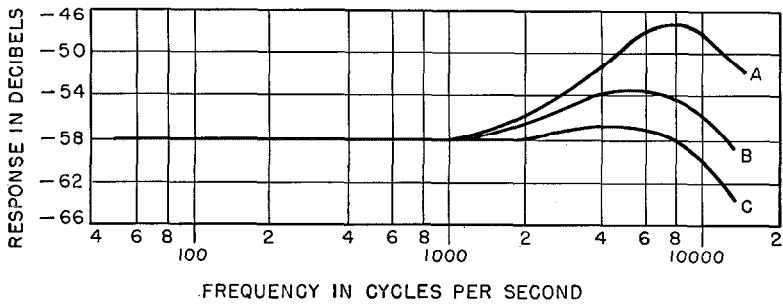


Figure 6—Voltage-frequency characteristic of type-A microphone. Open-circuit output voltage is plotted in decibels referred to 1 volt per dyne per square centimeter. *A* is for free-field operation with the grille and for sound incidence perpendicular to the diaphragm; *B* is for parallel incidence. *C* is for pressure response without the cavity.

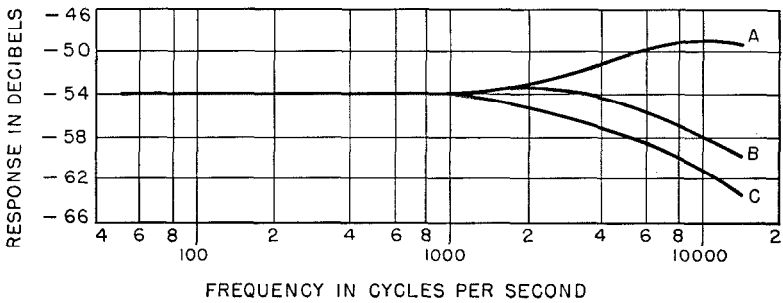


Figure 7—Characteristic of type-B microphone under the same conditions as in Figure 6.

phone has a slightly greater electrode spacing and a resulting lowered sensitivity to handle linearly sound pressures of 1500 to 2000 dynes per square centimeter.

As is evident from curve *C* of Figure 7, the type-*B* free-field microphone has a gradual roll-off characteristic when tested in a reciprocity coupler, showing that the first resonant mode is deliberately suppressed to provide the uniform free-field response indicated by curves *A* and *B*. A flat response characteristic in a free field is desired for high-fidelity recording, where a single capacitor microphone suitably located picks up the sounds of a large orchestra. Here, since the sounds reach the microphone at random incidence, the deviation from linearity is small and consists of very-uniform changes over a broad range of frequencies; it is undetectable audibly.

As shown in curves *A* and *B*, the response to frequencies above about 3500 cycles can be emphasized, de-emphasized, or rendered relatively flat by positioning the microphone diaphragm with respect to the incidence of the sound. This may appear as a fine point just after

mention of the fact that the response is very uniform to orchestral sounds reaching the microphone in random incidence. Nevertheless, the suggestion here is concerned with close-microphone recording, where it is desired to provide some high-frequency emphasis or a slightly accentuated brilliance to an orchestra, an orchestral section, or a voice. Similarly, where de-emphasis of brilliance is desired without increasing the distance from the microphone (and thus losing presence), proper positioning may be employed. If slightly increased microphone spacing is preferred to prevent blasting, microphone orientation can be employed to recreate the timbre qualities of close placement.

Extension of the reciprocity and free-field comparison curves to 15 000 cycles shows

no sharp peaks or deviations from uniformity. Carrying the investigations further by the actuator method has shown that the uniform roll-off carries on to ultrasonic frequencies on the free-field version. A secondary peak occurs at about 16 000 cycles on the type-*A* microphone and is of little concern in cavity work. Since the response-frequency characteristic is relatively uniform over the range of component frequencies of transients, the response of these microphones is excellent for highly transient sources, such as speech and musical percussive sounds of the piano, cymbals, and the like. Listening tests verify that sounds of this character are reproduced with clarity, crispness, and liveness.

The following data may be of interest. A sound pressure of 154 decibels (referred to 0.0002 dyne per square centimeter), which is the output limit of our present facilities, was applied to the microphone and no electrical failures resulted. Although the microphones were found to be linear in output voltage with respect to the applied signal, the system distortion in this test as measured on a harmonic-distortion meter for

type *A* was about 3 percent and for type *B* about 4 percent when 200 volts of polarization were used. The system distortion dropped to below 1 percent for type *A* at 142 decibels and the similar distortion point for type *B* is about 134 decibels. Since the distortion values obtained were well below the American Standards Association requirement, no attempt was made to eliminate the distortion introduced by the sound source and preamplifier circuits used.⁷ Therefore, our rating of the microphones based on the overload point for type *A* is 140 decibels and for type *B*, 130 decibels; values that allow a good margin for faithful reproduction. To indicate that distortion is a function of loudness of signal, when two microphones were used in a coupler, one as the sound source and the other as a microphone transmitter, the total distortion in our reciprocity apparatus was less than 1 percent at 85 decibels. This is the maximum level obtainable. Assuming that half of the distortion was associated with each microphone, the distortion at this level was considerably less than 0.5 percent.

The insulation resistance of the microphones with 250 volts of direct current applied is over 200 000 megohms for normal room conditions and has been found to remain at that figure indefinitely. Type-*A* microphones usually withstand direct voltages at 385 volts and type *B* of 270 volts. Individual microphones deviate from the usual breakdown values within -10 to $+20$ volts. The recommended polarizing voltage is 200 ± 20 volts, which provides an adequate safety factor for a simultaneously applied polarizing voltage, loud signal, and physical jolt.

As a result of adequate factors of safety, namely: sufficient breakdown voltage, high insulation resistance, tensioned diaphragm, and special attention to cleanliness in manufacture: a low value of inherent noise is obtained. Measurements to obtain the inherent noise of the microphone showed that it is below the measuring capabilities of our equipment. The noise level of the preamplifier was below 0.0001 volt and this below-scale reading was not altered with or without the microphone in the circuit. This value is in agreement with published information on such a device, which for inherent noise may be

⁷ L. L. Beranek, footnote reference 3, page 220.

considered as a 40-micromicrofarad capacitor shunted by a resistor of 20 megohms.⁸

These microphones, when cycled through temperature changes from -50 to $+60$ degrees centigrade, were found to maintain calibration accuracy. Whereas the American Standards Association specifications allow 0.05-decibel change per centigrade degree, the microphones appeared to vary only 0.02 decibel or less. This value cannot be presented with a higher degree of certainty at present since it is difficult to extract all effects of temperature variations occurring in the associated equipment.

The construction is such that when the rear of the microphone is open to the pressure existing on the front of the diaphragm, equalization to changes in pressure results. In laboratory tests, a pressure change from 10 to 740 millimeters of mercury over a 10-second period caused no visible distortion of the diaphragm. However, this pressure change during a 5-second interval caused the diaphragm to touch the rear electrode. Although this treatment did not affect the performance of the microphone, it is recommended that such changes be extended over a period of time somewhat greater than 10 seconds. It is further recommended that the polarizing voltage be removed during extensive pressure changes to prevent possible sparking and resulting electrical deterioration of the microphone.

4. Performance Limitations

For accurate work, certain precautions are important when using capacitor microphones. Since they have the characteristics of high-impedance electrical devices and use air for the dielectric and working medium, corrections should be made when environmental influences are abnormal. Altitude changes affect the acoustic controls, and corrections for the altered response-frequency characteristics must be made when large changes are involved. When the microphones are used for calibration work requiring greatest accuracy, corrections for these variables and in some cases for the capacitance change should be applied. If not abused, the microphones are stable and have maintained calibration over many years. For all ordinary usages, such as in factory and laboratory equip-

⁸ L. L. Beranek, footnote reference 3, pages 221-223.

ments and in broadcast and recording work, the Kellogg microphones remain uniform. There is no need to store these microphones in a desiccator except possibly for long-time (5 years or more) storage of verification standards.

Some capacitor-microphone characteristics are useful indicators of questionable performance. Should the calibration or sensitivity shift sharply, the microphone should be suspected of being defective due to mishandling or for other reasons. A measurement of the insulation resistance (through a 1-megohm protective resistor) and of its capacitance should be made. If the insulation resistance changes are 15 percent or less, the damage is slight and the microphone is still serviceable. Each micromicrofarad variation in capacitance changes the sensitivity by about 0.1 decibel and if the change exceeds 4 micromicrofarads the sensitivity and frequency characteristic will be seriously altered. An unsatisfactory microphone is usually characterized by flash-overs or noise, which make it necessary to remove the microphone from service—usually before response-frequency deterioration would be suspected.

5. Conclusions

The stretched-diaphragm capacitor microphone has been recognized for a long time as

having a very-uniform response-frequency characteristic, a wide dynamic range, low distortion, and low noise. Present-day recording techniques and higher-quality reproducing systems demand these performance advantages. The microphones described herein meet these requirements and also provide a good degree of ruggedness and calibration stability without sacrificing desirable characteristics. Further, the bandwidth of the microphone has been extended by allowing slight deviations from linearity. Since these deviations are minor and are spread over a broad range, they are not audibly detectable. Studies of response-frequency characteristics show the particular designs to be advantageous and emphasize the necessity for the two types of microphones, each specialized for its field of application.

6. Acknowledgment

Indebtedness is acknowledged to Mr. C. M. Miller, who is responsible for much of the design and exacting assembly of the microphones, and to Messrs. R. V. Hagen and C. H. Royce, who participated in the more-recent technical and refinement studies. The research and development were performed under the general direction of Dr. G. H. Brodie, who along with the others aided in the preparation and editing of this paper.

Manufacture of Type-7400 Step-by-Step Telephone Selector

By J. L. DONCEEL

Bell Telephone Manufacturing Company; Antwerp, Belgium

THE NEW TYPE-7400 step-by-step selector has been developed for use in telephone- and teleprinter-type machine-switching exchanges.

The selector is a 30-point single-motion rotary switch with wipers arranged to provide 30 or 60 outlets. The brush carriage with ratchet wheel

is actuated by a stepping mechanism operated by self-interruption. This step-by-step selector incorporates the following improvements over previous designs.

A. Greater stability of operation.

B. Greatly reduced maintenance.

C. Interchangeability of parts.

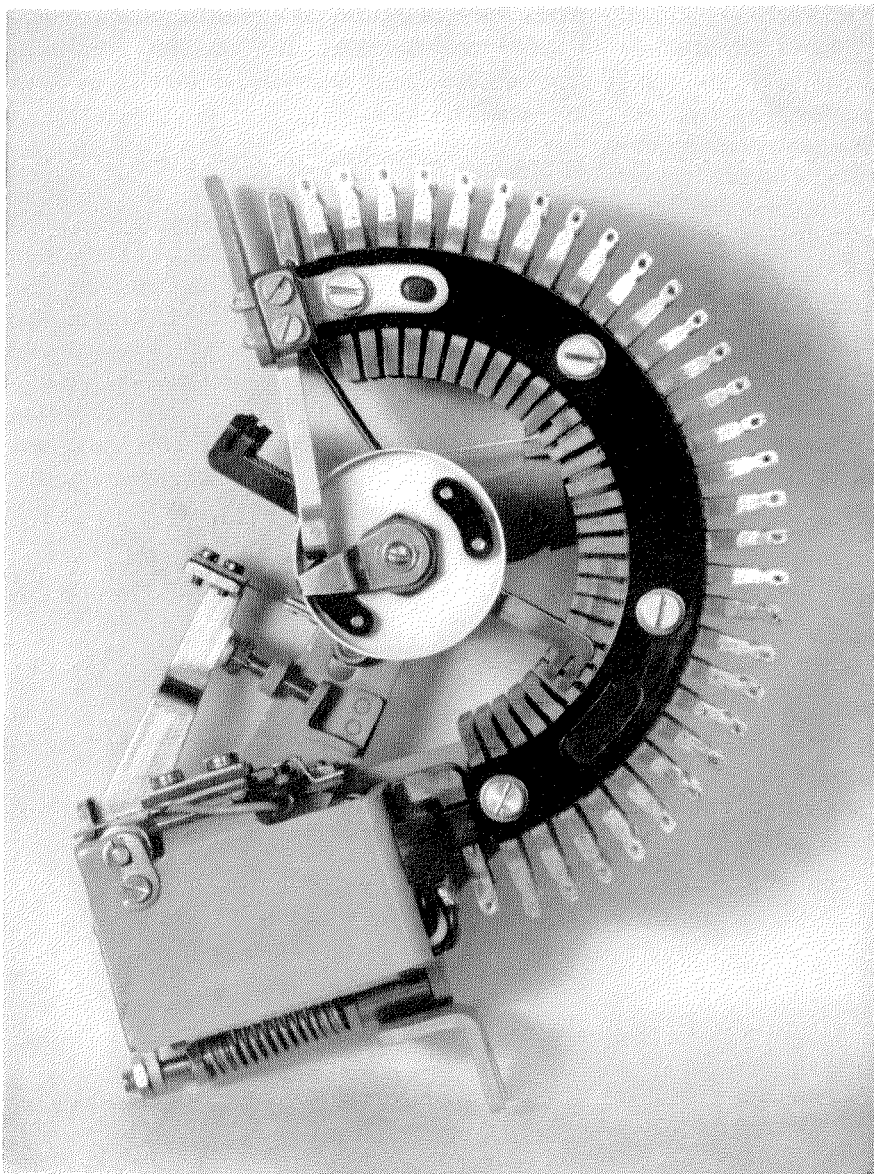
D. Stepping mechanism and rotor is a jack-in unit that can be adjusted before mounting on the contact bank.

E. Parts are very easily accessible and in plain view, allowing a visual check of operation.

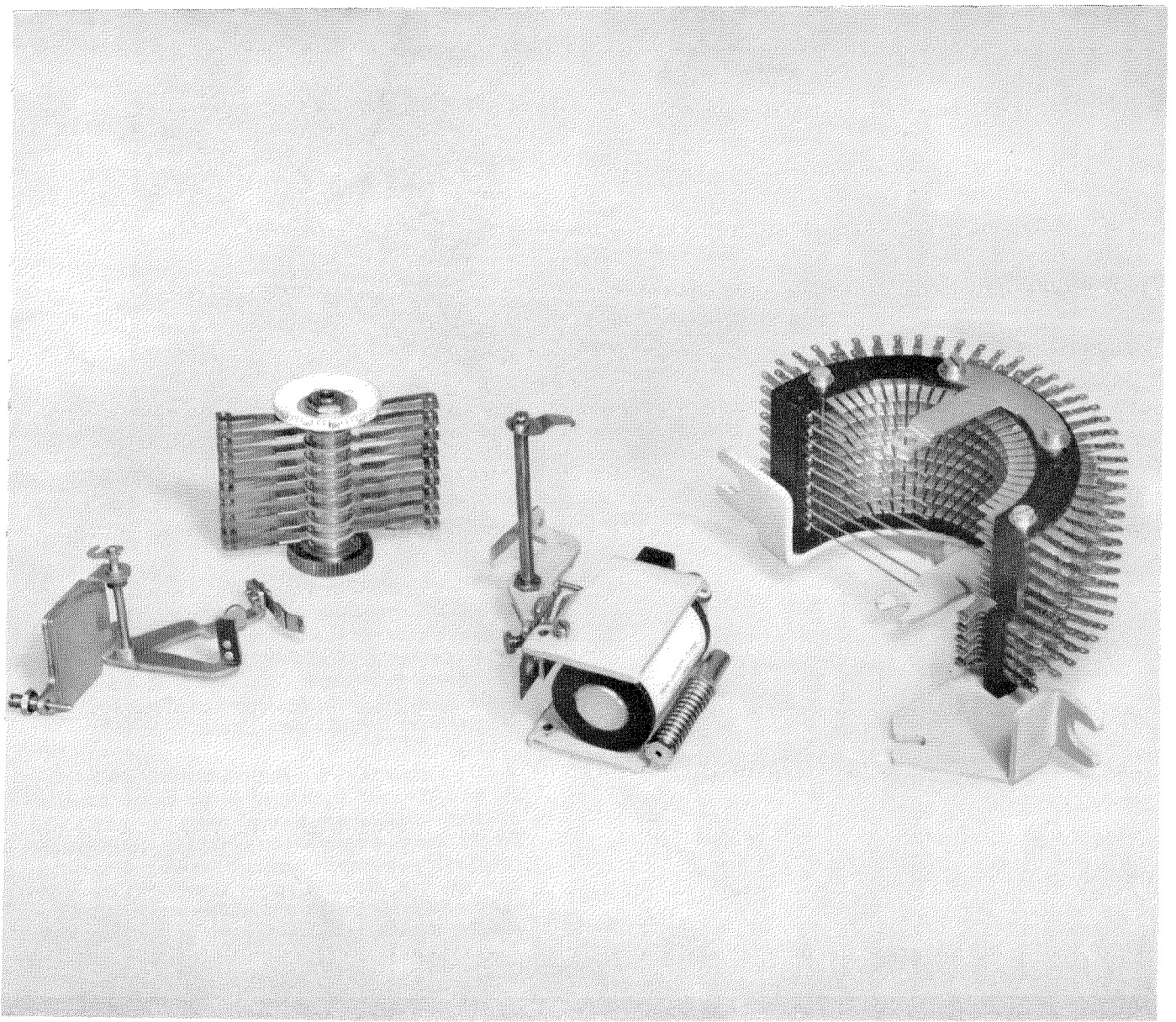
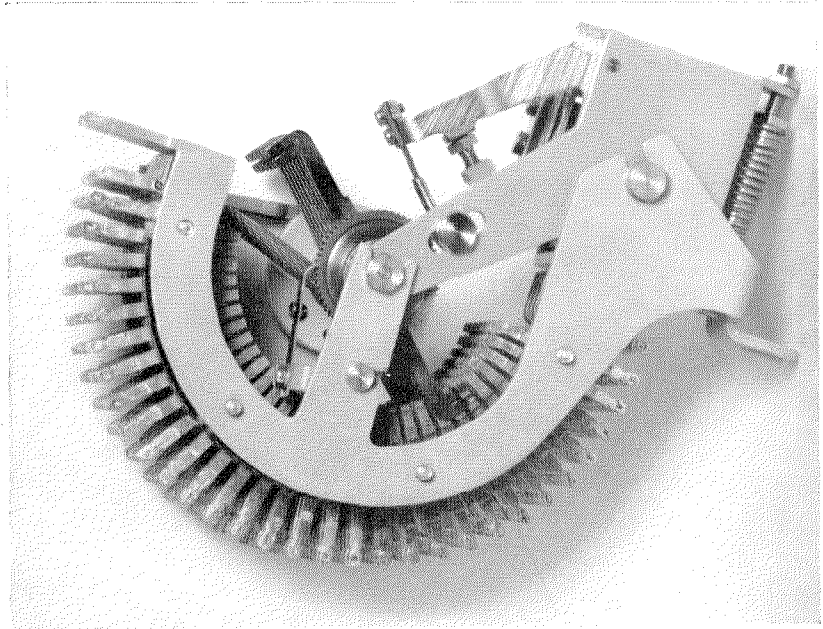
The different piece parts are manufactured within very accurate limits to ensure interchangeability of components.

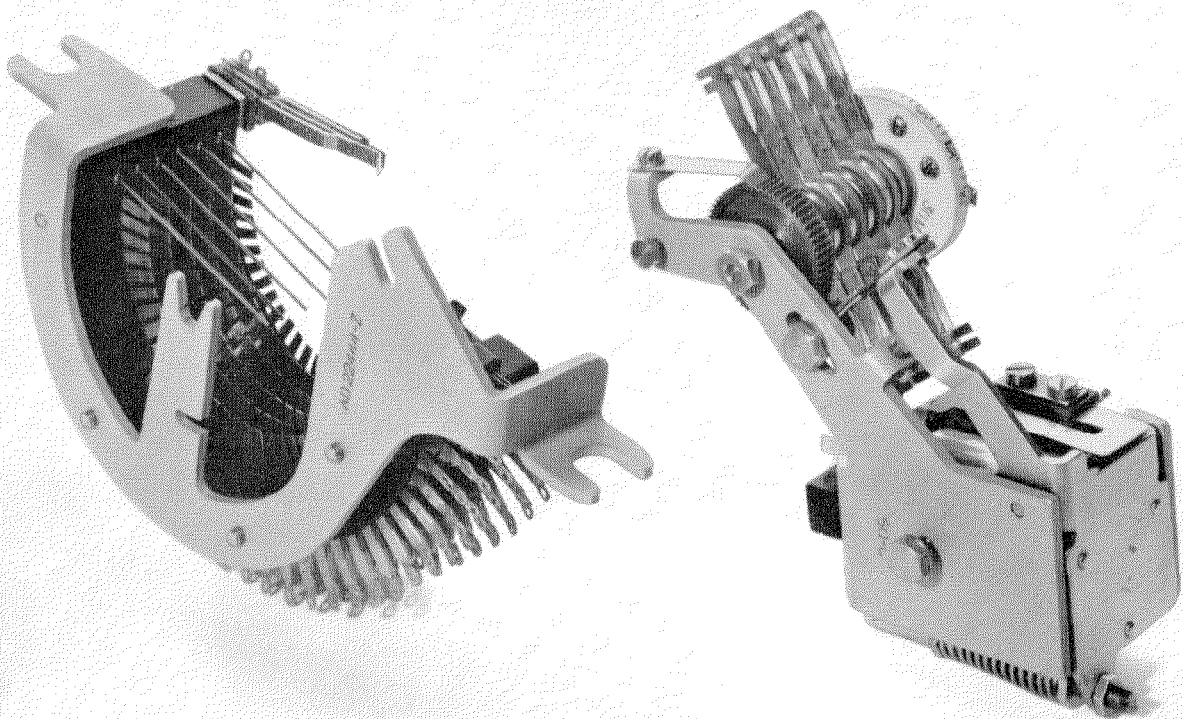
The maximum capacity of the selector is 10×30 points or 6×60 points. The switch operates at normal potential at a speed of 60 steps per second under self-interruption.

1. The assembled selector.



2. Another view of the selector from the base-plate side is given at right. Below, the major assemblies of the selector are, from left to right, the armature with driving pawl and adjusting screw for the armature spring; the rotor; the yoke with retaining pawl, armature spring, power coil, and rotor axis; and the contact bank.



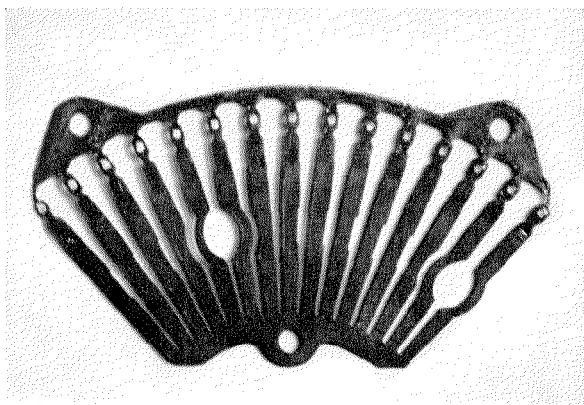


3. The selector can be disassembled into the two main parts shown here; at the left is the contact bank with the off-normal contact and, at the right, the stepping mechanism with the rotor.

Each of these two parts forms a complete unit that is preadjusted by means of suitable gages, after which they

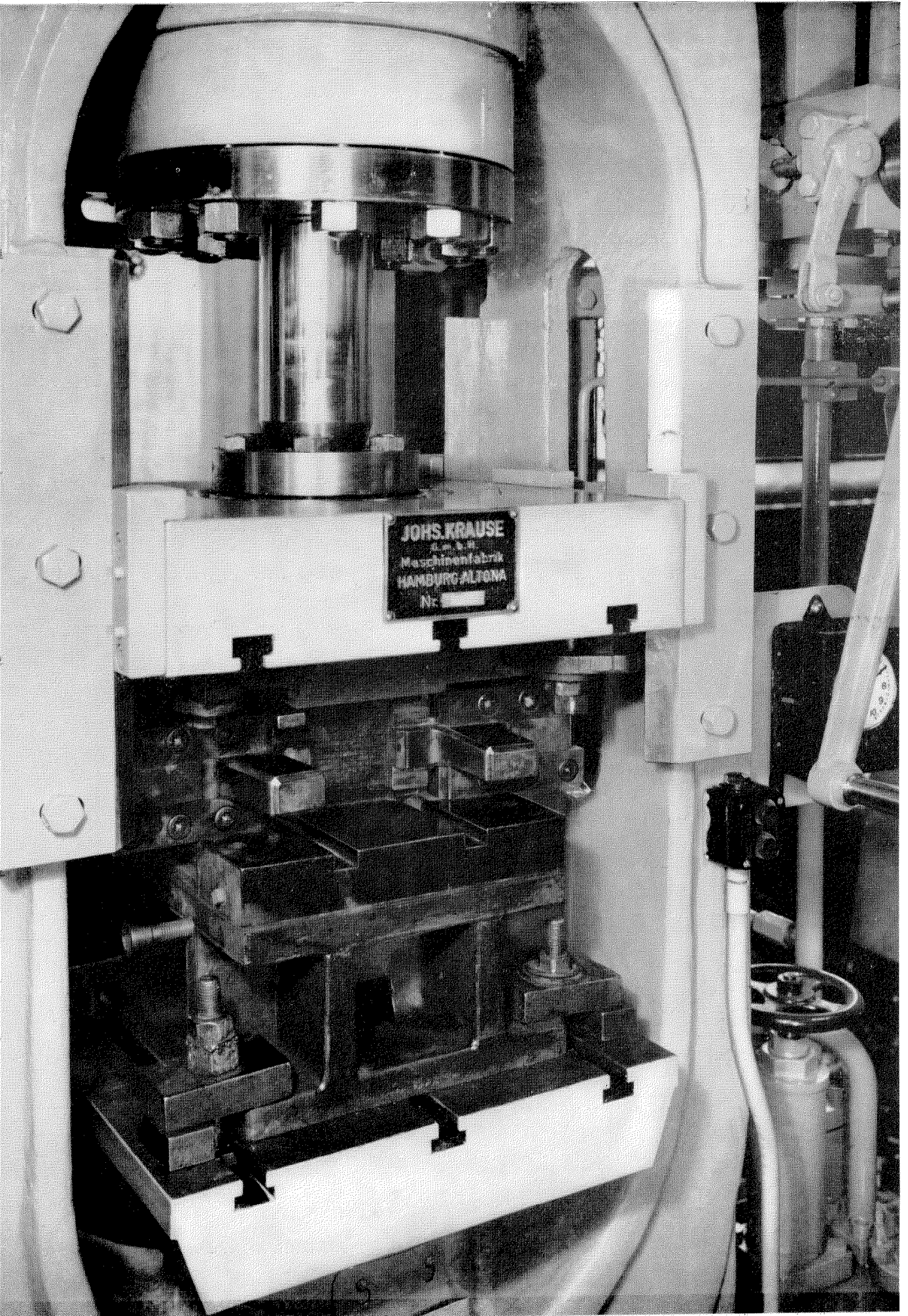
may be assembled easily and rapidly without further adjustment. The stepping mechanism is attached to the bank by one screw.

The jack-in connection between the winding of the power magnet and the interrupter on the bank greatly facilitates maintenance work.



4. The arc terminals of the contact bank are of phosphor bronze punched out in a form of a sector, an example of which is shown in the photograph at the left. Each sector incorporates 15 terminals. These are stacked in various combinations to provide the desired number of levels and contacts in the molded bank.

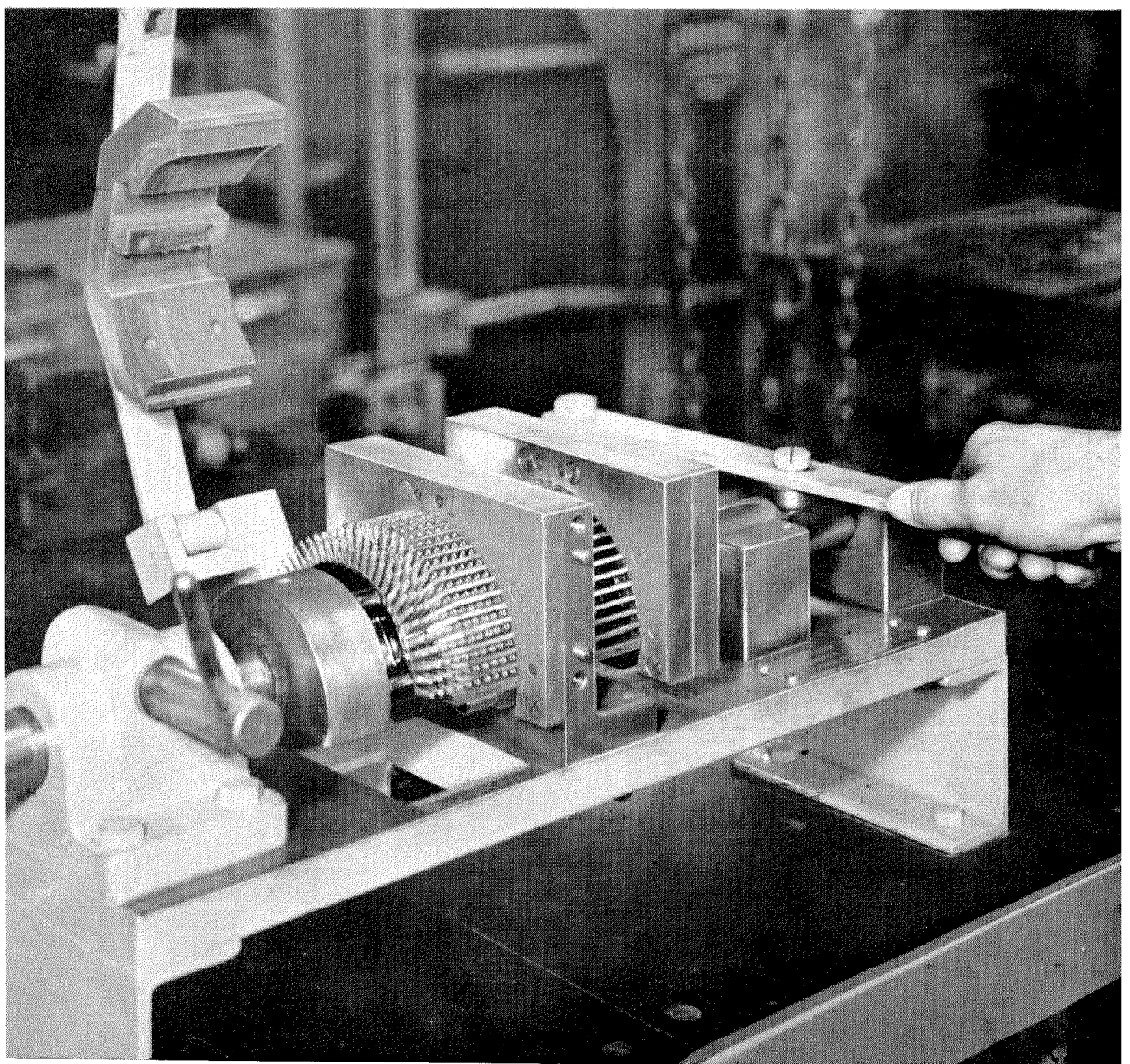
5. The contact bank is a single injection-molded bakelite unit. The mold is designed so that contact banks of different capacities may be made using the same mold. For the insulating material between the terminals, preference has been given to a thermosetting compound, as this gives the best guarantee against temperature distortion and may be molded within sufficiently accurate limits. The molding press in operation is shown on the facing page.

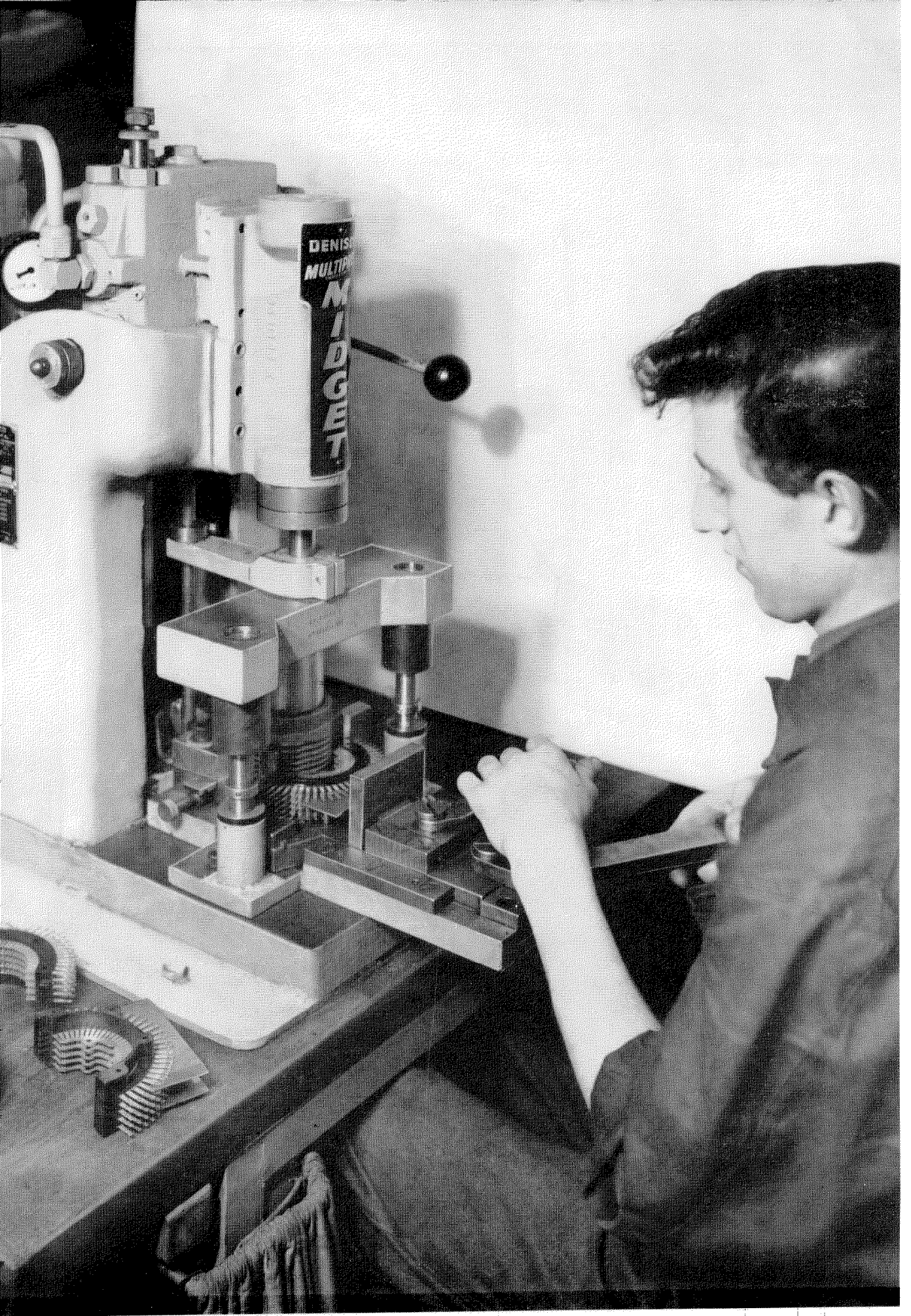


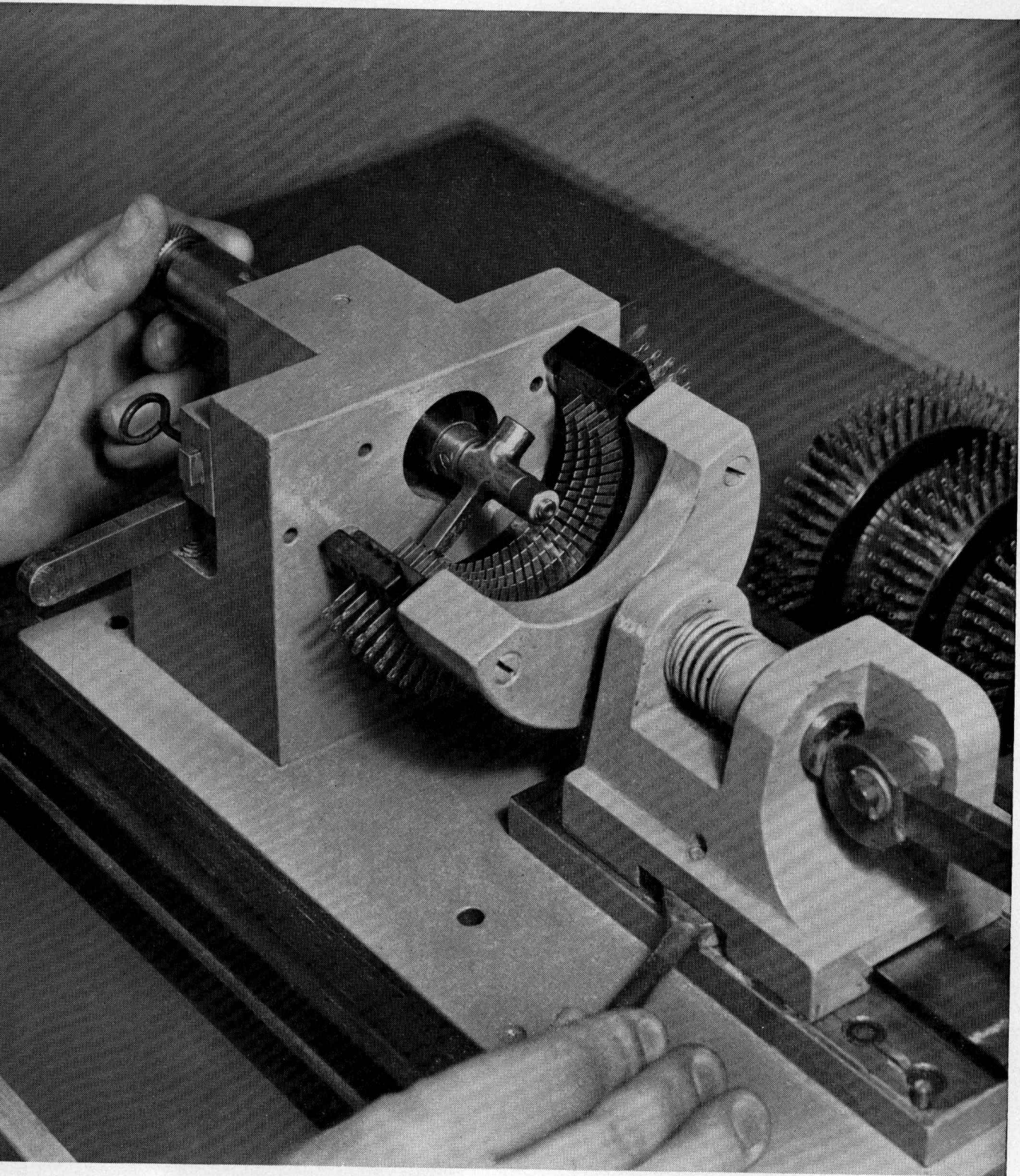
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6. Below, after the molding operation, excess bakelite is removed by a suitable tool. All bakelite particles between the terminals are removed in a single operation.

7. The next process is cutting out the metal sectors that hold the terminals together during molding. This is done in a midget multipress as shown on the facing page.





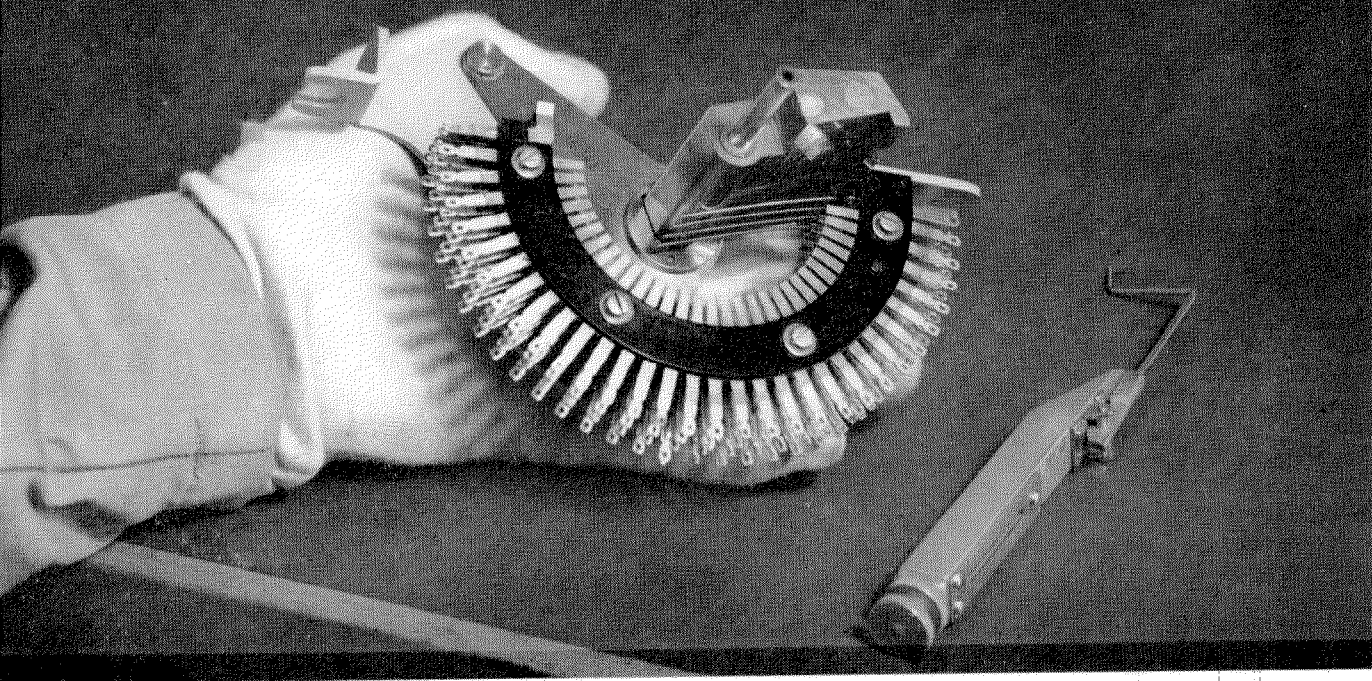
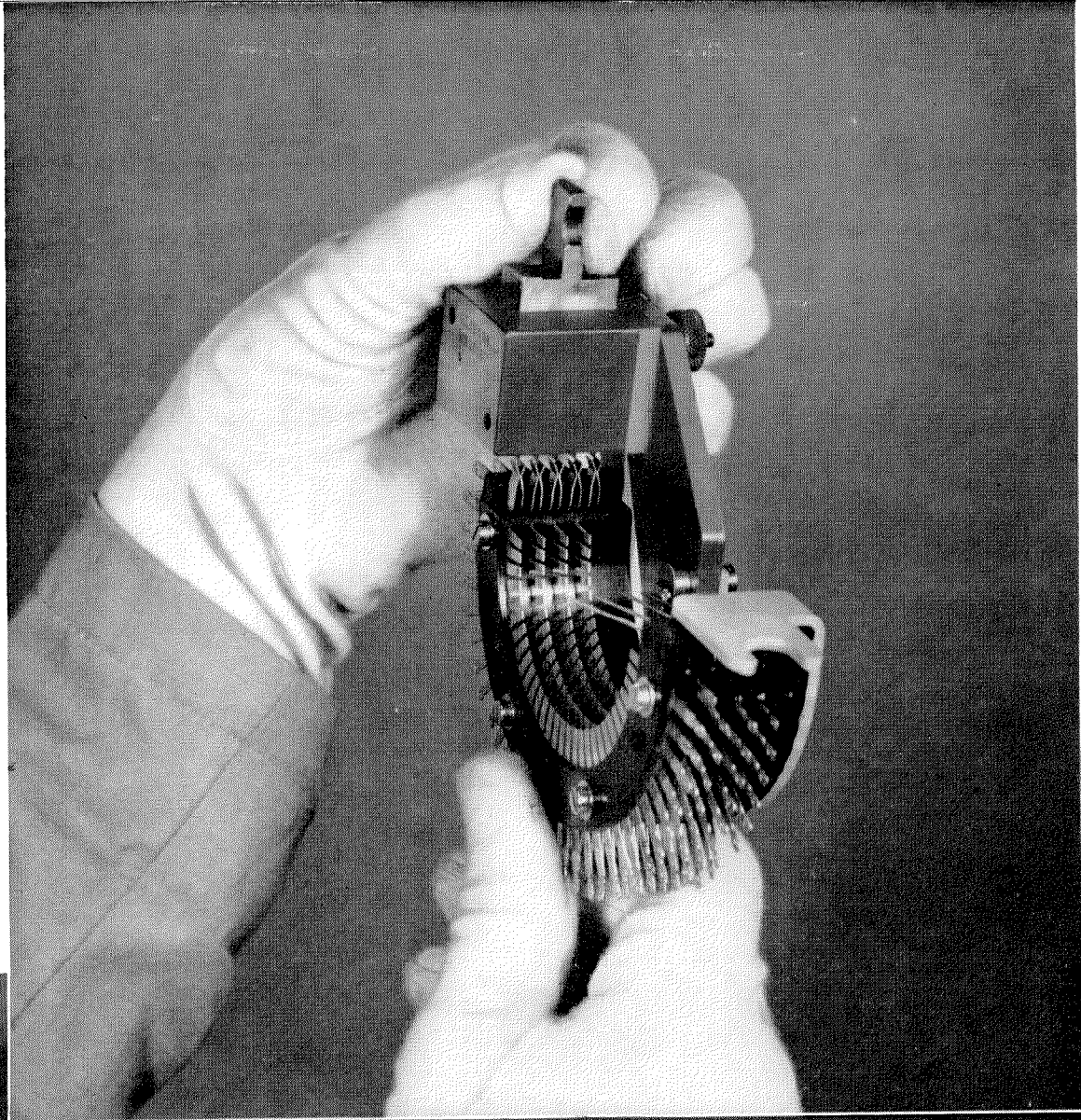


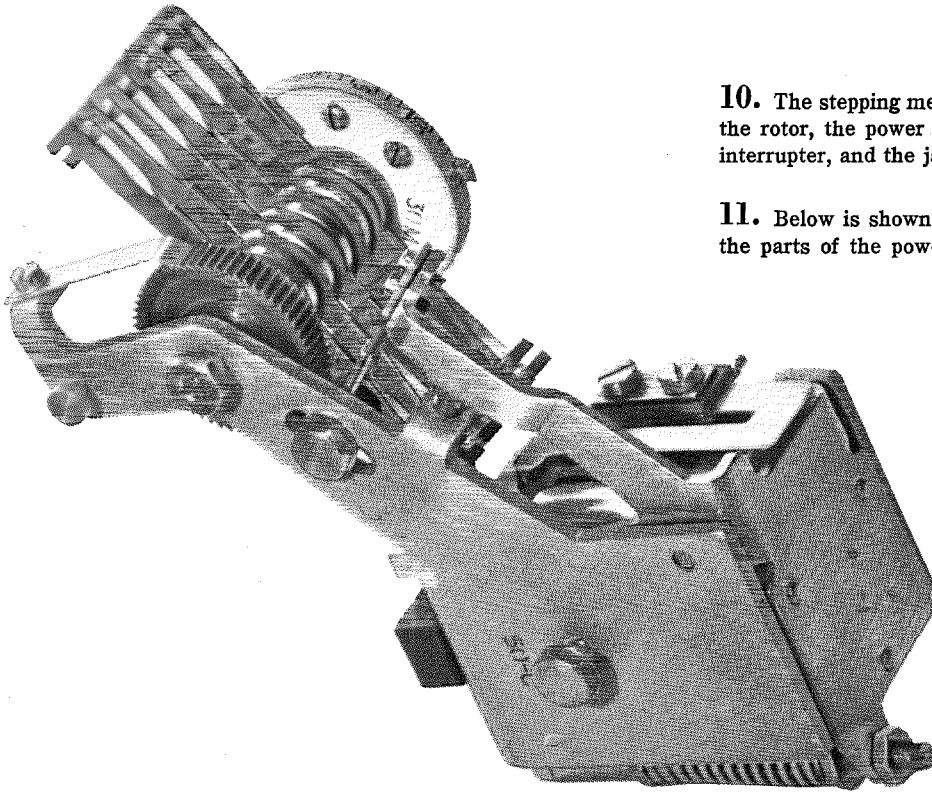
8. The contact banks are then checked for correct position and alignment of each level of terminals. This is performed in a single operation by the special gage shown in the photograph above.

9. The frame holding the contact bank is a simple flat punched-steel plate with mounting lugs bent over at 90 degrees. The frame is simply manufactured and is punched within very accurate limits with no supplementary treat-

ment afterward. It is dimensionally stable; deformations due to ageing are eliminated.

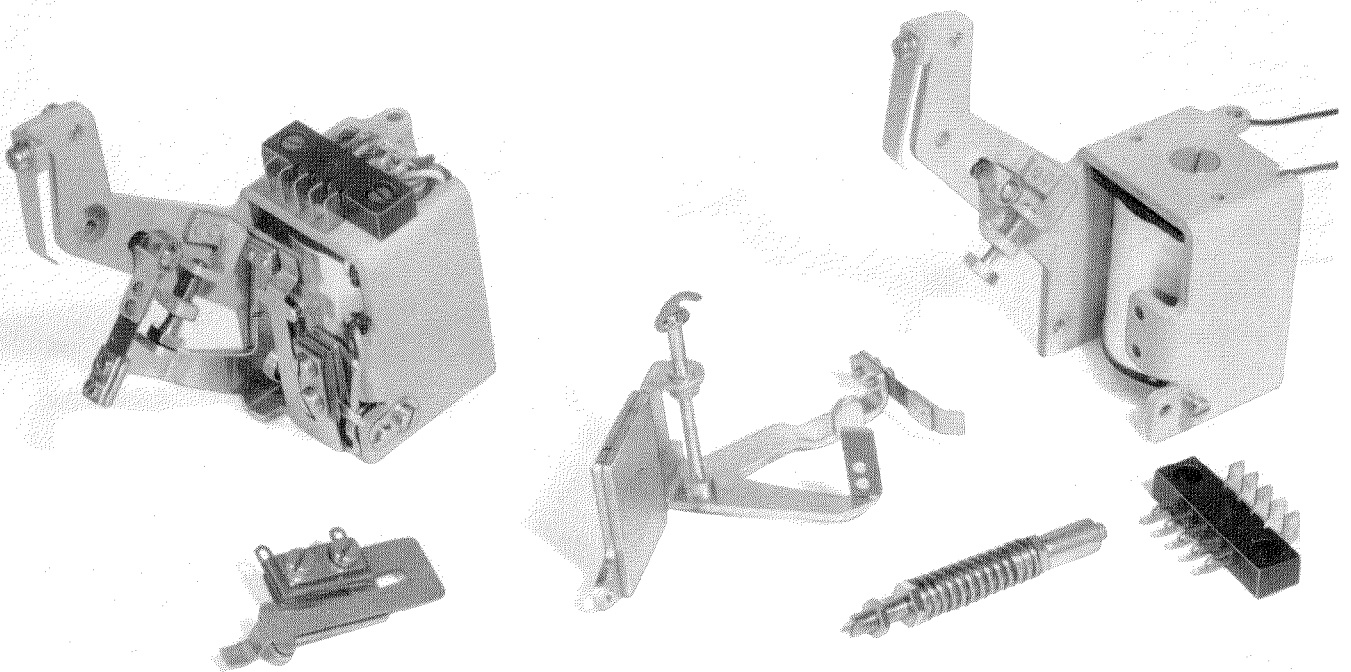
The contact bank is now assembled into the frame and the collector springs and jack-in springs are mounted on the contact bank. The springs are located by slots in the molding into which they are inserted and fixed by clipping. This permits the springs to be easily removed from the arc for possible replacement. The facing page shows how the collector and jack-in springs are checked and adjusted.





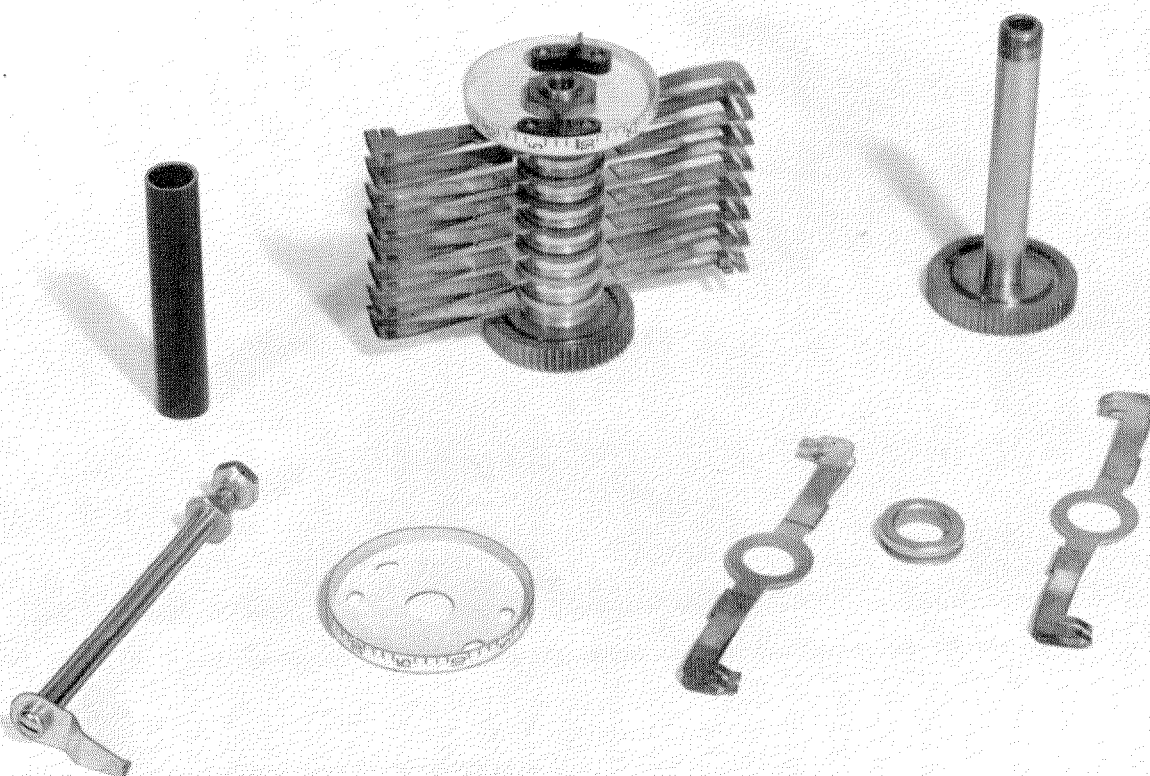
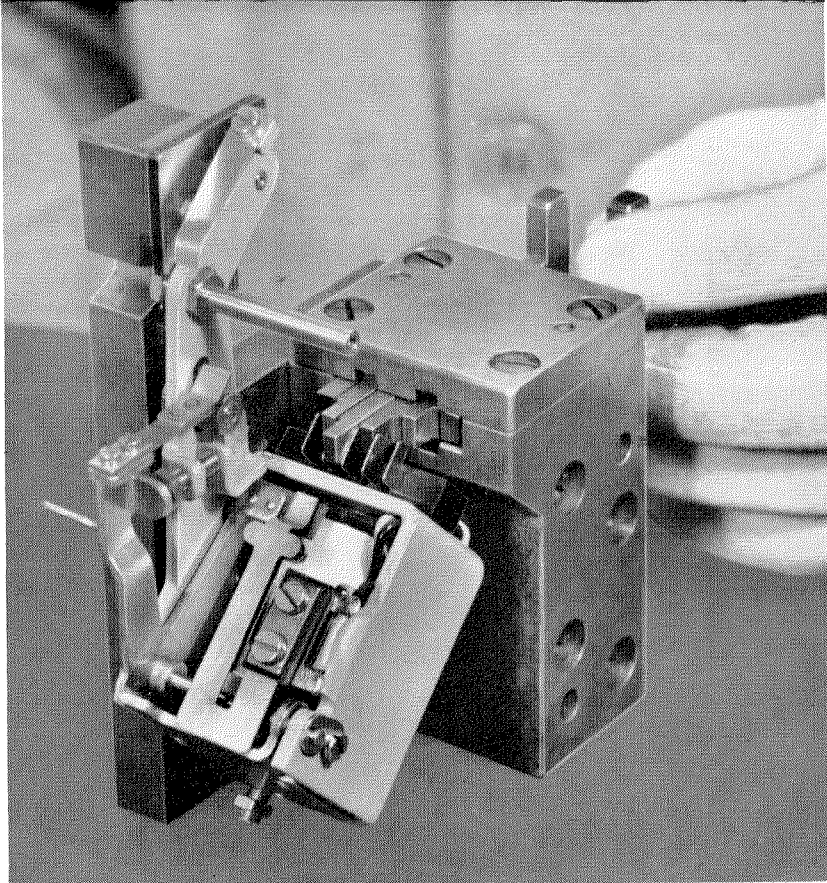
10. The stepping mechanism shown at the left comprises the rotor, the power magnet with driving pawl, the self-interrupter, and the jack-in contacts.

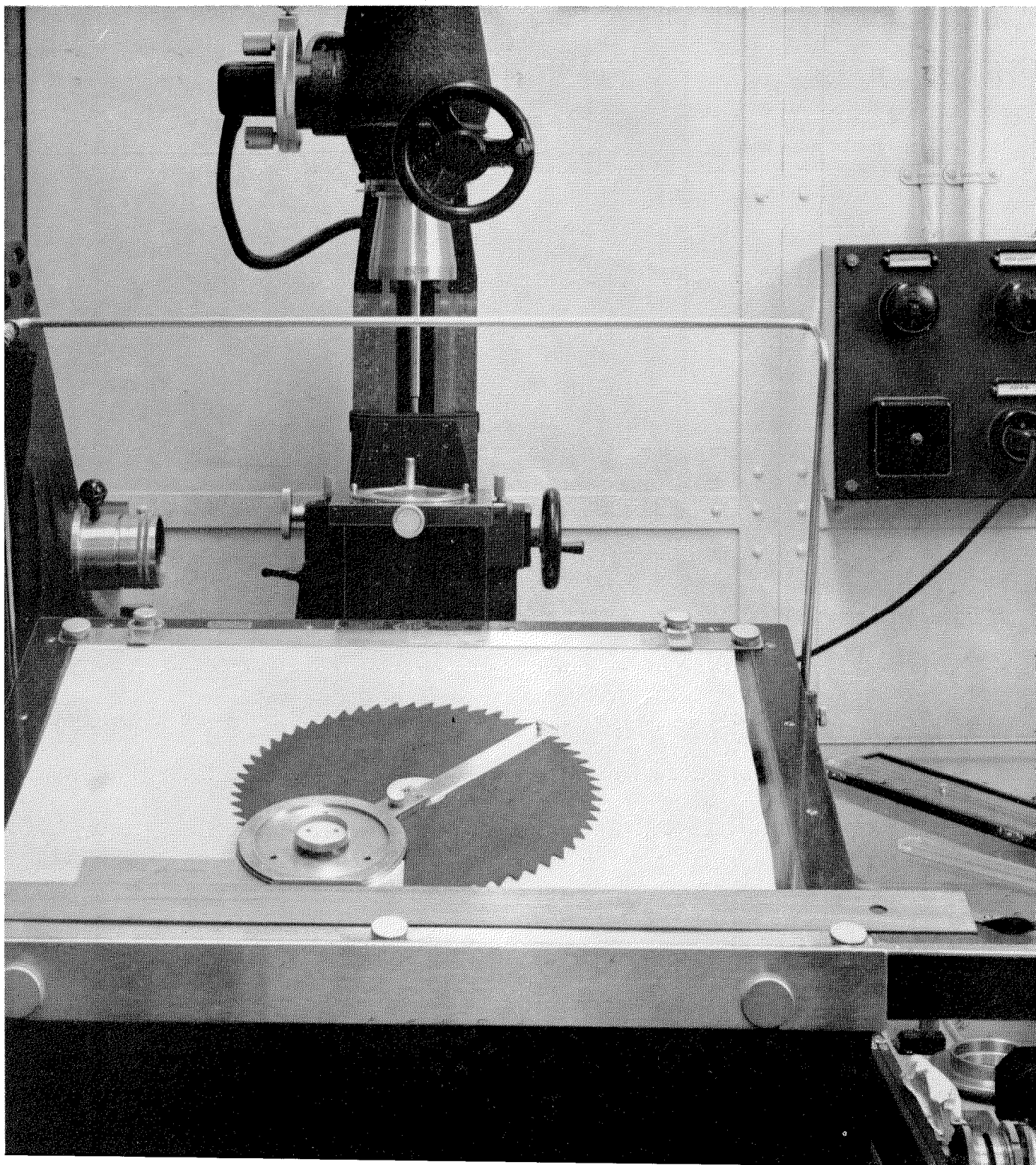
11. Below is shown an assembled view (upper left) and the parts of the power magnet with self-interrupter and jack-in contacts. The armature may easily be removed without dismantling any other part of the switch. The main spring has therefore been made of the push type and is provided with a self-locking screw for accurate adjustment of armature tension.



12. At the right is shown the gage for checking the correct position of the jack-in contacts.

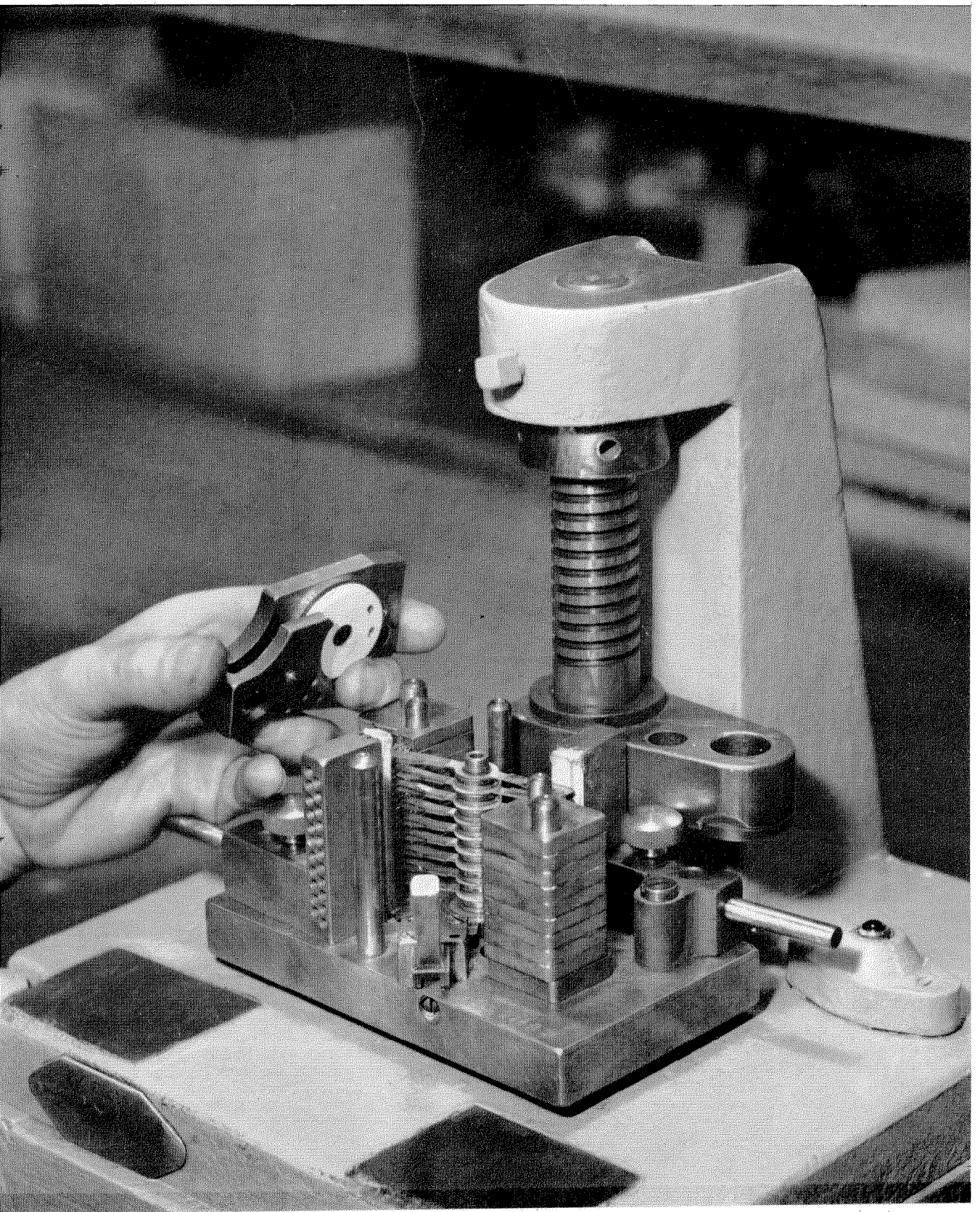
13. Below, the assembled rotor and its parts, consisting of a ratchet wheel with hub on which the contact brushes, collector rings, and number wheel are mounted.

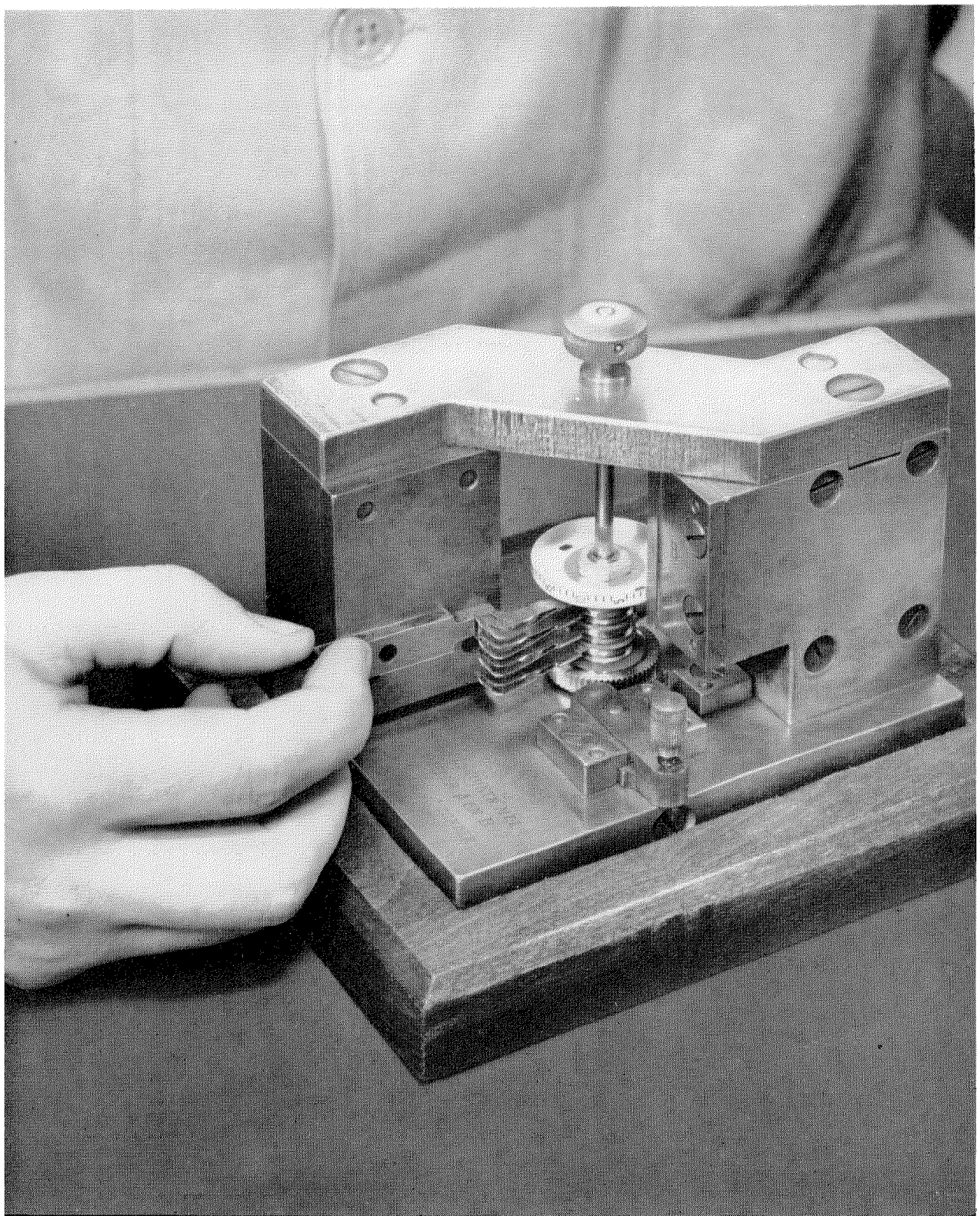




14. The teeth of the ratchet are checked on an optical comparator to ensure correct operation.

15. The rotor is assembled by means of the fixture shown here and is afterwards squeezed under a weight of 200 kilograms (4400 pounds) for firm fixation of the parts.

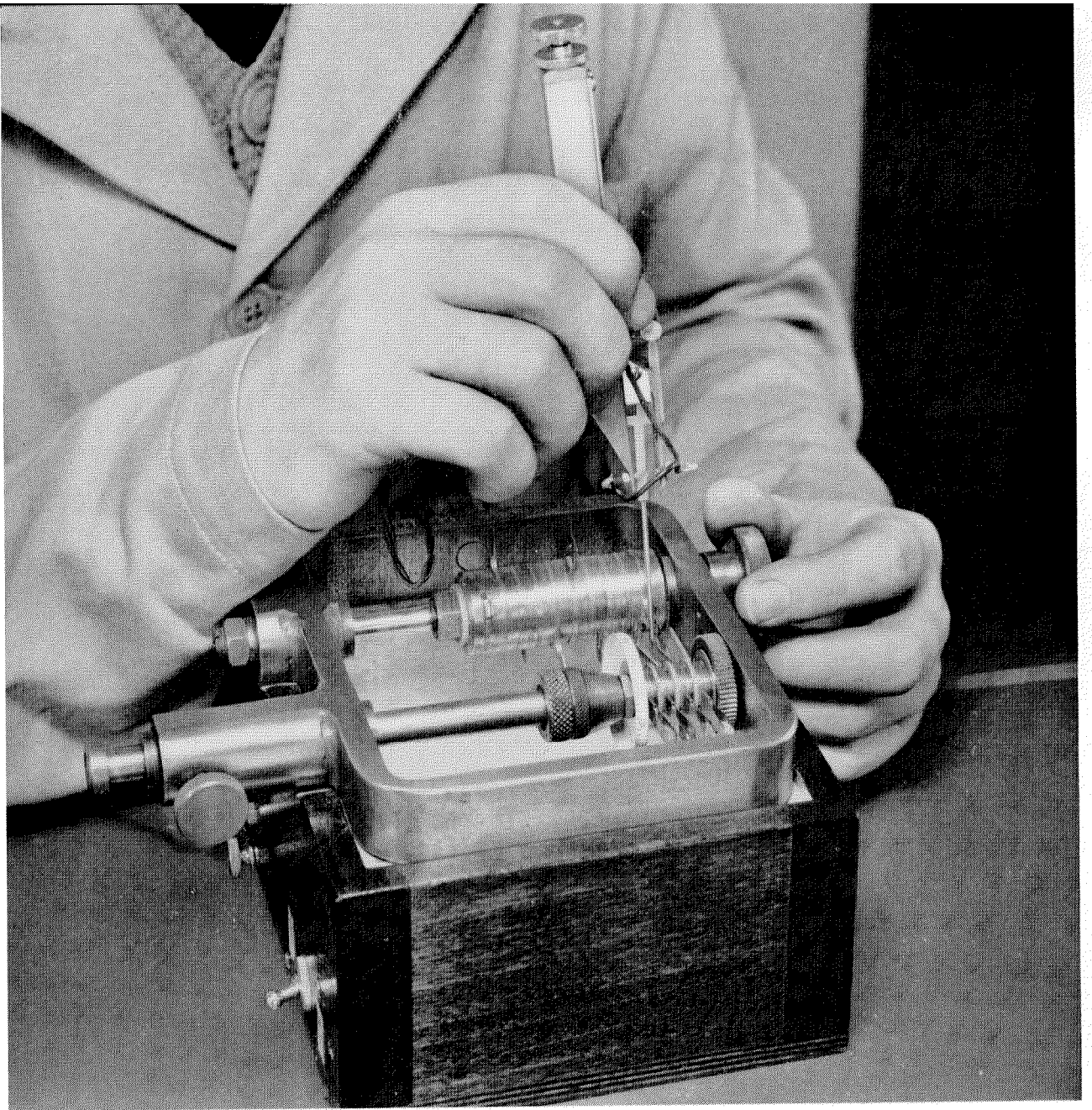




16. The contact brushes are of a special construction and are made in such a way that they may be adjusted to run properly over the terminals without affecting the contact pressure of the springs. Both springs of the same brush are therefore welded to a brass brush member that is fixed in the pile-up of

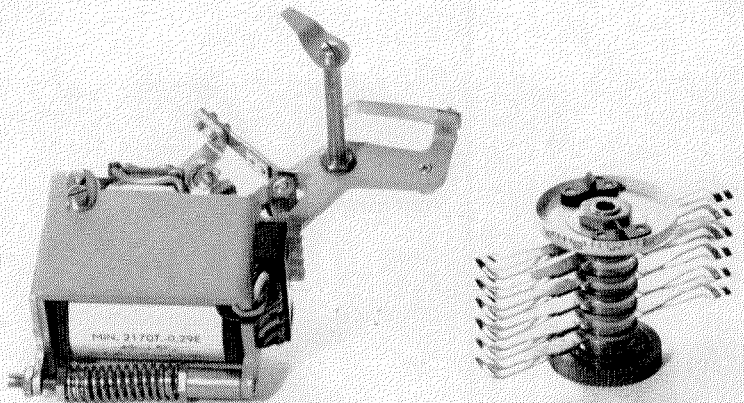
the rotor and may be adjusted to obtain the correct brush alignment.

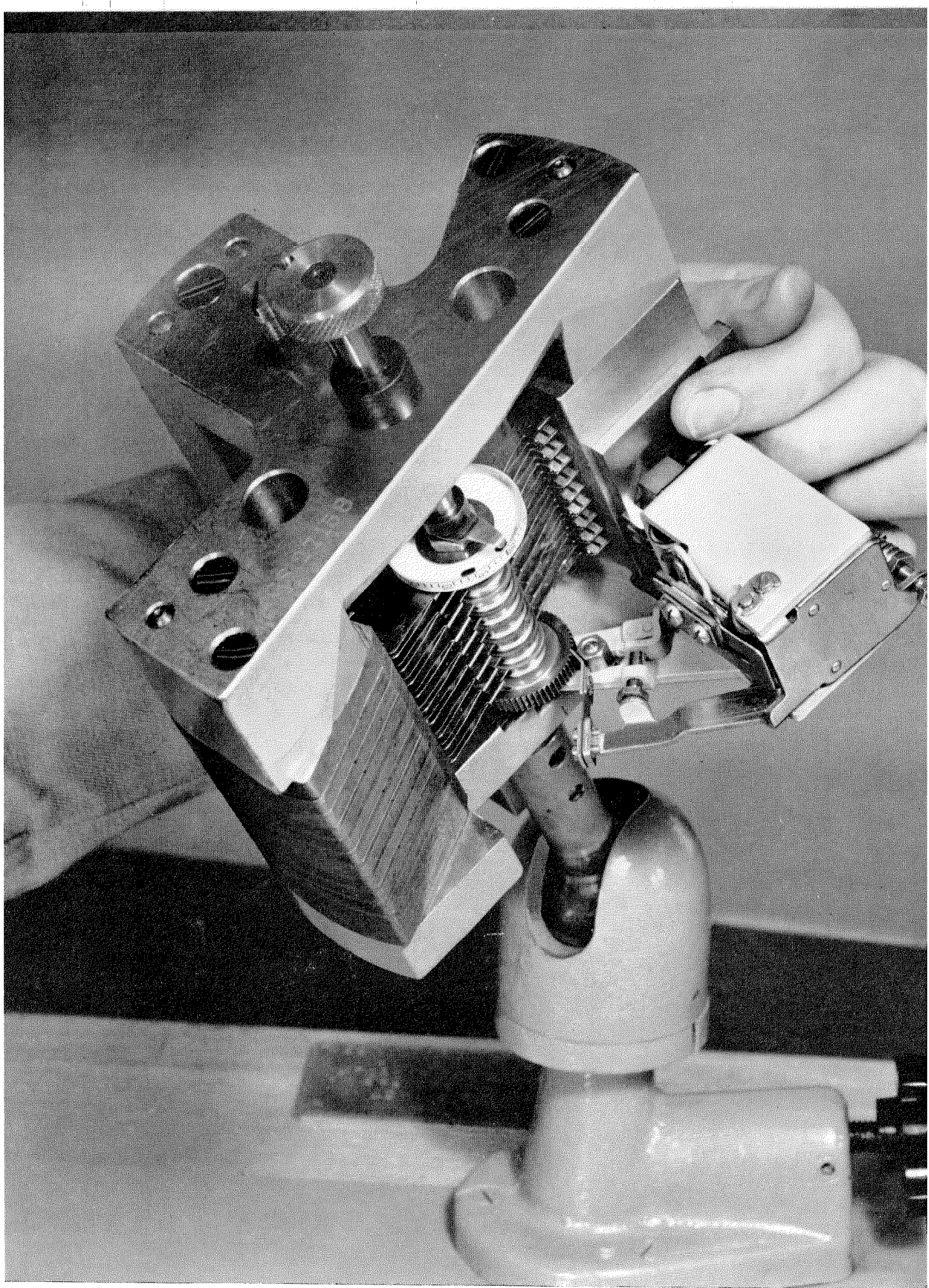
After assembly of the rotor, the correct alignment of the spring tips of the different brushes of a set is checked on a gage as shown here. The position of the number wheel is checked at the same time.



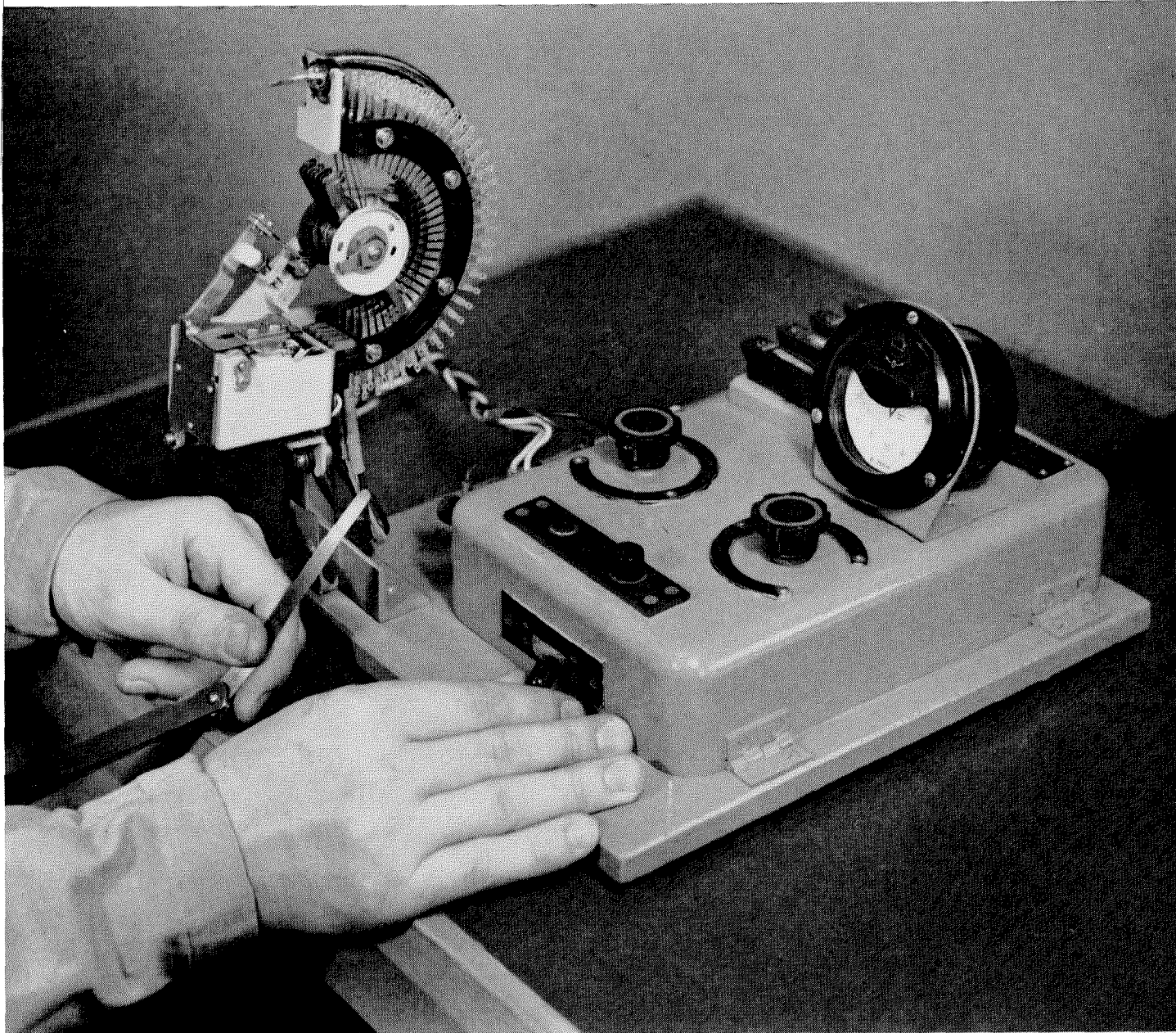
17. Above, the rotor is adjusted for the contact-brush position and for contact pressure on a gage. After this, the rotor may be mounted on any switch without further adjustment.

18. The photograph at the right shows the power magnet and assembled rotor.





19. On facing page, the magnet-and-rotor assembly is checked by a gage that represents a theoretical contact bank. This verification is of particular importance to obtain the interchangeability of the stepping mechanism to any contact bank of regular production.



20. When the contact bank is assembled with the stepping mechanism, an electrical test is made. It consists of a current test for armature operation and for correct operation under self-interruption. Above is shown the test box and fixture for the step-by-step selector.

Characteristics and Applications of Thermally Sensitive Resistors, or Thermistors *

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BASIC properties of thermistors, the terms used to specify their characteristics, and some of their applications to electronic and telecommunication problems are described in this tutorial paper.

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1. Properties of Thermistors

Thermally sensitive resistive materials have been known to scientists for over a hundred years. As early as 1834, Michael Faraday reported measurements on the high negative temperature coefficient of resistance of silver sulphide and later experiments by English and German scientists, using magnesium-titanium oxides and the oxides of uranium, have been recorded. The temperature coefficients obtained in this early work were between -1.5 per cent and -2.5 per cent per degree centigrade at 20 degrees centigrade, but the results obtained varied from sample to sample and moreover the characteristics proved unstable in life, particularly after exposure to moderate temperatures or to the passage of electric current through the sample.

Modern development may be considered to date from about 1930, when the Bell Telephone Laboratories sought a device to compensate for the variation in resistance of telephone transmission lines with temperature and concentrated on the study of a few carefully selected materials. After several years of research and development, all of the previous disadvantages had been overcome, and semiconductor materials had been produced having extremely stable characteristics and long life. Laboratories in other countries were also active and finally a new circuit device, in the form of an extremely stable thermally sensitive resistor was made available to industry. Various trade names were adopted by different

manufacturers, one of the best known being *Thermistor*, which is a contraction of the words *thermally sensitive resistor*, and for convenience this name will be used in the present paper.

1.1 COMPOSITION AND MANUFACTURE

Thermistors are complex metallic-oxide compounds usually made from the oxides of manganese, nickel, copper, and cobalt, although some manufacturers use mixed crystals such as $MgAl_2O_4$ and $ZnTiO_4$. Uranium oxide is also used, but thermistors made from this material exhibit polarisation effects and can therefore only be used on alternating current. In all cases, the desired resistivity and temperature coefficient are controlled by suitably proportioning the mixture. During manufacture, the thermistor material is fired at a high temperature and is not affected by subsequent heating up to 500 degrees centigrade, although due to limitations imposed by soldered connections, glass-bulb enclosures, et cetera, the maximum usable temperature of the finished product is about 300 degrees centigrade for those sealed into a vacuum or gas-filled enclosure and about 120 degrees centigrade for those exposed to the air.

The resistance material may be formed into small bead-type elements, tubes, rods, blocks, disks, or flakes and the heating of the element may be achieved either by passing current through the resistive material itself (directly heated type), by an independent heater insulated electrically from the element (indirectly heated type), or by placing the element in a heated enclosure such as an oven.

1.2 TEMPERATURE COEFFICIENT¹

The resistance R decreases with increasing absolute temperature T (see Figure 1) according

* Reprinted from *Proceedings of the Institution of Radio Engineers, Australia*, volume 13, pages 123-131; May, 1952.

¹ G. Bosson, F. Gutmann, and L. M. Simmons, "Relationship Between Resistance and Temperature of Thermistors," *Journal of Applied Physics*, volume 21, pages 1267-1268; December, 1950.

to the approximate expression

$$R = a \exp[b/T],$$

where

R = resistance at any temperature T in degrees Kelvin

a, b = constants depending on composition of thermistor material.

The temperature coefficient is $-b/T^2$ and thus decreases with increasing temperature; its value is usually of the order of -3 to -4 per cent per degree centigrade at 20 degrees centigrade compared with a value of about $+0.35$ per cent per degree centigrade for platinum.

Though the above simple law is adequate for most purposes, Bosson, Gutmann, and Simmons¹

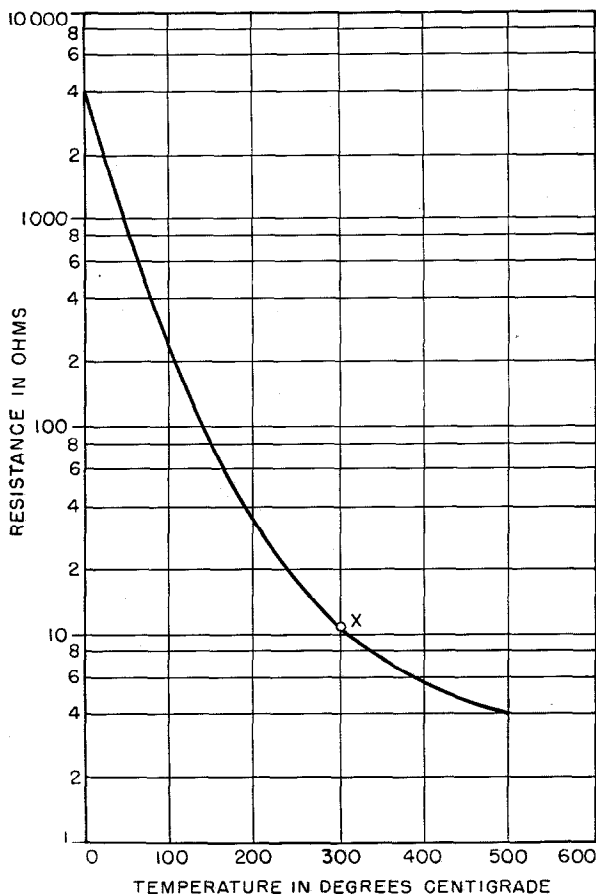


Figure 1—Characteristics of type-F2311/300 thermistor developed for thermometry. $R_c = 2000$ ohms at 20 degrees centigrade. Point X marks the maximum operating temperature.

have proposed an empirical law that gives a more accurate estimate of the resistance at any particular temperature provided that the characteristics of the material used are known.

1.3 CHARACTERISTICS

1.3.1 Resistance Value

Thermistors are available with almost any desired value of cold resistance R_c , with certain standardised values between the limits of 500 ohms and 500 000 ohms, and they may be directly or indirectly heated.

1.3.2 Heater-Coil Resistance

The heater-coil resistance of the indirectly heated type is generally $100 \text{ ohms} \pm 5$ per cent and the heater efficiency is defined as the ratio between the power dissipated in the bead and the power dissipated in the heater to reduce the resistance to 2 per cent of the cold value. The heater efficiency of typical indirectly heated thermistors is approximately 90 per cent.

1.3.3 Half-Temperature

There are several ways of specifying the manner in which the resistance of a thermistor element changes with heating power. One way is to specify the percentage change of resistance for a change of one degree centigrade and another is by a statement of *half-temperature* or a figure derived from this. By half-temperature is meant the temperature change in degrees centigrade necessary to change the value of the resistance to a value $R/2$ or to a value $2R$. For a complete identification of the value, it is necessary to state the approximate temperature at which this half-temperature is measured. The thermistor whose characteristic is shown in Figure 1 has a half-temperature of 18 degrees centigrade at 20 degrees centigrade, whereas for the same thermistor at 200 degrees centigrade, the half-temperature is 45 degrees centigrade. At the higher temperature, when the resistance is but a fraction of its original cold resistance, a change of 45 degrees centigrade is required to produce a further halving of the resistance value, and this factor sets one limit to the useful range of operation of the thermistor.

1.3.4 Power Sensitivity

The power sensitivity is defined as the rate of change of temperature with applied power input and is expressed in degrees centigrade per milliwatt at 20 degrees centigrade. This characteristic depends on a number of factors, including the mass of resistance material and the method of mounting the resistance element. The relation between power sensitivity and power input is approximately linear over the first 50-degree-centigrade rise from room temperature, and this fact is useful in determining the effect of ambient-temperature variations on the other characteristics of a thermistor. Table 1 shows the power

Hence to maintain the thermistor at a constant resistance, 1 milliwatt more must be dissipated for each 3.5-degree-centigrade fall in ambient temperature. If trouble-free service is desired the power input must not be allowed to exceed the maximum for continuous use; this maximum is specified by the maker. For periods up to a minute, it is usually quite safe to exceed this rating by 50 per cent. In the case of indirectly heated thermistors, the total power dissipation in the bead and heater must not exceed the maximum indicated. It should be noted that if the temperature is fixed, the thermistor obeys Ohm's law; that is, the relation between voltage and current is linear.

TABLE 1
CHARACTERISTICS OF VARIOUS TYPES OF THERMISTORS

Type	Description	Power Sensitivity Degrees Centigrade per Milliwatt	Dissipation Constant Milliwatts per Degree Centigrade	Maximum Voltage Factor <i>K</i>
<i>A</i>	Bead formed on two 0.020-inch (0.5-millimetre) platinum wires and sealed in an evacuated bulb	3.5	0.28	20
<i>B</i>	Indirectly heated bead sealed in an evacuated envelope. Heater resistance, 100 ohms. Maximum voltage between heater and thermistor, 25 volts	5.0	0.2	22
<i>C</i>	Rod with pigtail ends	0.066	15.1	2.3
<i>F</i>	Small bead fused into tip of evacuated glass bulb	1.6 (Air) 0.4 (Water)	0.62 (Air) 2.5 (Water)	—
<i>K</i>	Cylinder, 0.625 inch (16 millimetres) in diameter by 0.125 inch (3 millimetres) in length	0.065	15.4	2.5
<i>KB</i>	Block, mounted on metallic plate 1.25 by 1 inch (6.4 by 2.54 centimetres) with four mounting holes	0.04–0.01, According to Mounting	25–100 According to Mounting	1.2–2.2, According to Mounting
<i>L</i>	Indirectly heated bead in a gas-filled envelope. Heater resistance, 100 ohms	2.5	0.4	16
<i>M</i>	Bead, sealed in glass	1.0 Maximum, According to Mounting	1.0 Minimum, According to Mounting	—

sensitivity of thermistors made by one manufacturer. It will be noted that the block type (*K* and *KB*) are relatively insensitive to power input and this renders them particularly suitable for use as ambient-temperature compensators. When power is dissipated by the resistance element of a thermistor it becomes hotter and thus the resistance decreases with increase of power input; therefore, when the ambient temperature falls, a greater input is required to reduce the resistance of the thermistor to any particular value. For example, it can be seen from the table of power sensitivity that a power input of 1 milliwatt will increase the temperature of a type-*A* thermistor by 3.5 degrees centigrade.

1.3.5 Dissipation Constant

The dissipation constant is the reciprocal of the power sensitivity and is expressed as the power input necessary to raise the temperature of the element by 1 degree centigrade (see Table 1).

1.3.6 Voltage/Current Curves and Maximum Voltage Factor

The voltage and current characteristics are of considerable importance (see Figure 2). As the current through a thermistor is increased, the voltage at first rises rapidly and soon attains a

maximum value E_{max} . Thereafter, there is a rapid drop in voltage with further increase in current. As the current increases, the rate of fall of voltage decreases and may, under suitable con-

ditions, remain fairly constant over a wide range of current. It can be shown² that an approximate value of E_{max} for any thermistor can be calculated by means of the following formula:

$$E_{max} = (R_e)^{1/2}/K,$$

where R_e is the resistance at the ambient tem-

²J. A. Becker, C. B. Green, and G. L. Pearson, "Properties and Uses of Thermistors—Thermally Sensitive Resistors," *Electrical Engineering*, volume 66, pages 711-725; November, 1946.

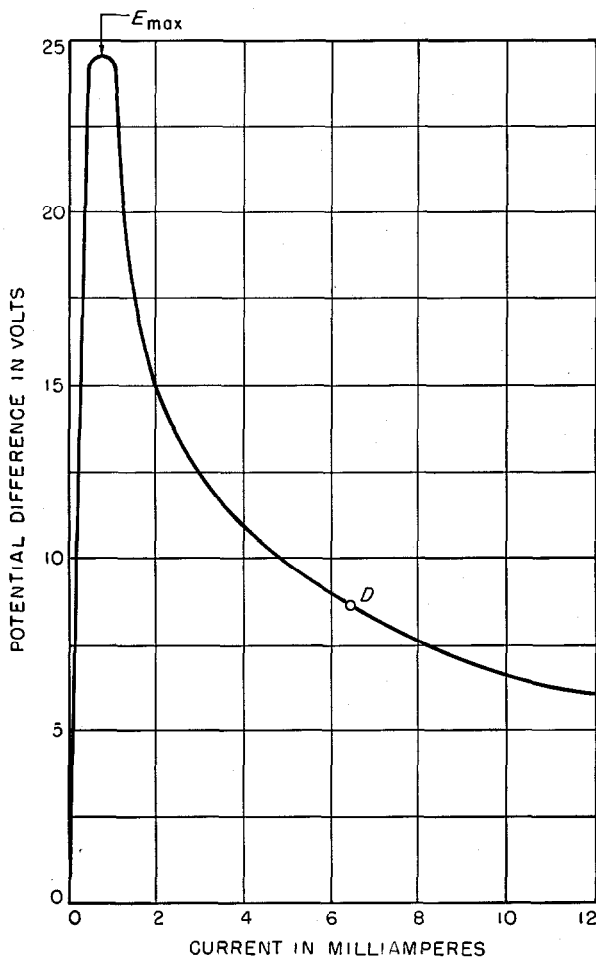


Figure 2—Typical thermistor (type A1522/100) voltage-current curve. $R_e = 100\ 000$ ohms at 20 degrees centigrade. Between E_{max} and point D there exists a negative-resistance portion that may be put to certain uses.

perature (that is, with no dissipation) and K is known as the maximum voltage factor and is dependent only on the ambient temperature and the type of thermistor. Its value for each type at an ambient temperature of 20 degrees centigrade, is given in Table 1.

$$(50\ 000)^{1/2}/20 = 11 \text{ volts, approximately.}$$

1.3.7 Time Constant

The time constant of an indirectly heated type is defined as the time taken for the logarithm of the resistance of the thermistor to change by $(e - 1)/e$ (that is, 63 per cent of the final change) when the heating power is suddenly applied or removed. It is usual to measure the time constant for a change of heating power which, if applied until equilibrium is reached, will halve (or double) the resistance value. It is found that in the high-vacuum types the time constant for a decrease of heating power is longer than for heating, whereas in the gas-filled types, the time constants (heating and cooling) are more nearly the same, and generally speaking, shorter than for the high-vacuum types. The average time constant of a type-B thermistor is of the order of 3 or 4 seconds, while that of the type L is of the order of 1 to 2 seconds. The time constant of a directly heated thermistor is measured by reducing the dissipation in the thermistor from a large value to zero because of the difficulty of maintaining a constant dissipation (other than zero) in an element whose resistance is changing. For the same reason, the transient behaviour of a directly heated thermistor will vary from circuit to circuit, depending on the manner in which the dissipation alters with the resistance of the thermistor.

1.4 STABILITY OF CHARACTERISTICS

After the thermistor has been subjected to the ageing treatment that is a normal part of the manufacturing process, there is every reason to believe that it retains its characteristics unchanged after long usage. Life tests have been

carried out both with continuous heating and also with the heating current switched on and off every 20 seconds; the original characteristics are maintained after many hundreds of hours of life test.

Experience has shown that the life of a thermistor depends upon its correct use in a circuit. The following simple rules have been drawn up as a guide to the operating conditions under which a thermistor will give good service.

A. The potential applied to the thermistor element must not exceed E_{max} when the circuit series resistance is less than 1.25 per cent of the thermistor cold resistance.

B. The potential applied to the thermistor element must not exceed twice E_{max} when the circuit series resistance is greater than 1.25 per cent of the thermistor cold resistance.

C. The maximum power that may be continuously dissipated is that indicated by the maker. For periods up to one minute, however, this figure may usually be exceeded by 50 per cent.

2. Some Applications of Thermistors

Thermistors have come into very wide use in recent years for a variety of different applications. Their small size, ruggedness, stability, and versatility make them extremely useful circuit elements, capable in many cases of fulfilling a function hardly possible by any other means.³

It is not necessary to list here all the possible applications of thermistors, but a very broad outline of the basic uses will be given accompanied in many cases by a fundamental circuit diagram.

2.1 TEMPERATURE MEASUREMENT

The use of thermistors for temperature measurement stems directly from their large temperature coefficient of resistance (approximately 10 times that of platinum), which makes possible very sensitive thermometers. In fact a thermistor thermometer has been described² that indicates changes of 0.0005 degree centigrade. The high resistance of a thermistor makes the effects of long leads negligible in most cases and the small

³ J. E. Tweeddale, "Thermistors—They Perform Countless Control Jobs," *Western Electric Oscillator*, pages 3-5, 34-37; December, 1945.

thermistor element can be very usefully employed for remote indication, particularly as the indicator can be a relatively inexpensive direct-current meter. Temperature measurements of

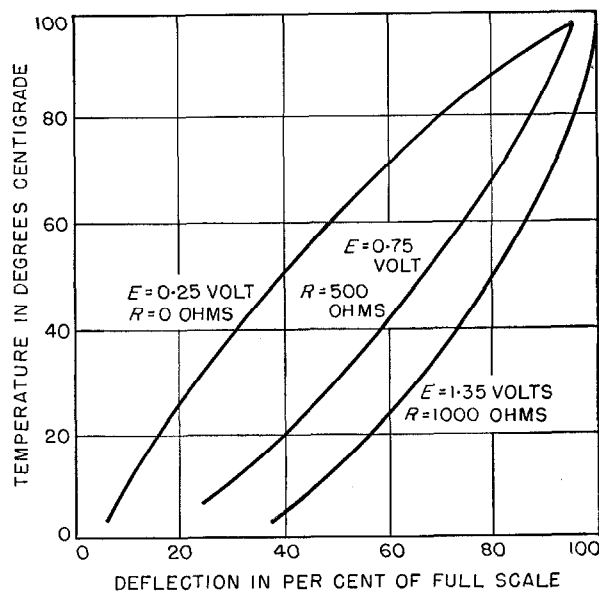
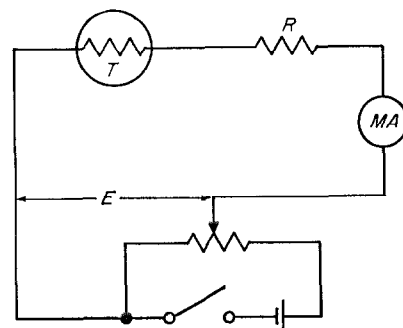


Figure 3—Thermistor thermometer. Type-A1311/100 thermistor is employed; $R_c = 1000$ ohms at 20 degrees centigrade. The 1-milliamper meter has a 100-ohm resistance. Characteristic curves are given below.

intravenous blood⁴ and supercharger rotors, and the insertion of tiny bead thermistors into journal boxes, gear housings, and power transformers during construction are good examples of this type of use. In radiosonde telemetering systems,⁵

⁴ L. F. Drummeter and W. G. Fastie, "A Simple Resistance Thermometer for Blood-Temperature Measurements," *Science*, volume 105, pages 73-75; January 17, 1947.

⁵ V. D. Houck, J. R. Cosby, and A. B. Dember, "Radio-sonde Telemetering Systems," *Electronics*, volume 10, pages 120-123; May, 1946.

the remote indication is carried out over a radio link. Mention should also be made here of a type of thermistor developed specially for thermometry (type *F* in Table 1), which has a very small mass of thermistor material fused into the glass tip of an evacuated bulb, so that it responds almost instantly to temperature changes.

Figure 3 shows a simple circuit using a thermistor for direct temperature indication. The calibration curves show how the scale linearity and the possible range of temperature measurement can be controlled by changing the battery voltage and the value of the series resistance. For more precise work, a Wheatstone bridge circuit may be used with the thermistor as the unknown resistor. The temperature may be read either directly on the indicator, or by null balance of the bridge; this latter method giving greater accuracy.⁶ The fundamental circuit is given in Figure 4. R_1 and R_2 are the ratio arms, T is the thermistor, and R_3 is a known resistor, either fixed or variable depending on whether the direct-reading or null-balance method is to be used.

A principal factor to consider in the design is the selection of circuit components and voltage so that the power dissipated in the thermistor

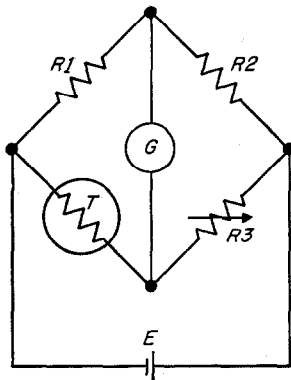


Figure 4—Wheatstone bridge circuit for temperature measurement.

itself is too small to produce sufficient self-heating to raise its temperature appreciably above that of its surroundings. The permissible dissipation can be determined from the dissipation constant. The slope of the resistance/temperature char-

⁶ E. L. Deeter, "Null Temperature Bridge," *Electronics*, volume 21, pages 180-186; May, 1948.

acteristic decreases with increasing temperature, so that it is advisable to dissipate the smallest possible amount of power in the thermistor in order to obtain maximum sensitivity. If substantial self-heating is present, the thermistor operates on a non-linear portion of its characteristic (see Figure 1).

When measuring very small temperature differences, it is sometimes desirable to compensate for ambient temperature variations. R_1 or R_2 may be replaced with a suitable thermistor the characteristic of which is well matched to that of T . Any change in ambient temperature then causes an equal change in resistance of both thermistors and the bridge remains in balance.

An object may be ordinarily inaccessible to direct temperature-measuring devices, either through its physical position or because it is in motion, or it may be too hot for contact thermometry. In such cases, its temperature may be determined by permitting radiation from the object to be focussed on a suitable thermistor, as in the noteworthy case of the infra-red bolometer developed during the war to detect the radiation from submarine engine exhausts. This utilised a flake thermistor so sensitive that it would detect the warmth of a human being a quarter of a mile away.⁷ Another example is the method used by a manufacturer of tin cans to measure the temperature of the soldered seams while the cans were moving past on a conveyor belt at a rate as high as 350 cans per minute.⁸ As it was impossible to make physical contact with the surface being measured because of its rapid motion and inaccessibility, radiant energy was the only energy available as a means of temperature measurement. An optical system was arranged to focus radiant energy from the cans on a flake thermistor and the variation in the amount of energy radiated as the cans moved past the line of sight of the optical system was sensed by the thermistor. The resulting signal voltage, suitably amplified, was presented as a pattern on a cathode-ray oscilloscope.

⁷ "Thermal Detectors," *Electronic Industries*, volume 5, pages 87, 116-117; September, 1946.

⁸ K. P. Dowell, "Thermistors as Components Open Product Design Horizons," *Electrical Manufacturing*, volume 42, pages 84-91, 190, 192, 194 . . . 216; August, 1948.

2.2 TEMPERATURE CONTROL

The same basic requirements apply to temperature control as for temperature measurement, in that the sensitivity should be high and the stability good. In addition, the resistance of the thermistor at the control temperature should be of a value suitable for the circuit used. The circuit shown in Figure 4 may be used with the indicator replaced by a relay that controls the heating or cooling device so that it is switched on or off, according to the requirement, when a predetermined temperature is reached. Normally a valve amplifier is required and the circuit shown in Figure 5 is suitable. The allowable

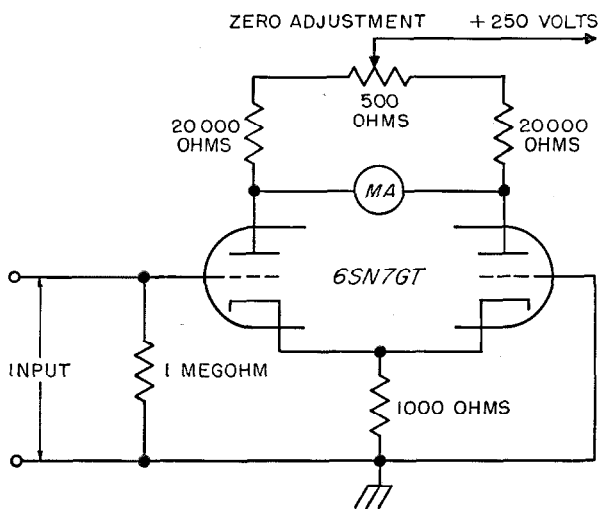


Figure 5—Valve voltmeter for use as indicator with Wheatstone bridge. A 1-milliamper meter is used.

power dissipation in the thermistor elements will be determined by their dissipation constants as for thermistor thermometers. Thermistor bridges have been used for the control of domestic heating systems, refrigeration, and food processing.

To maintain a constant temperature inside a room, a heating system need only balance the heat loss through walls and windows et cetera. There is a time lag in conventional control methods between the actual loss and the increased supply from the heat source to balance the loss. A simple thermistor bridge circuit provides an effective control by making use of two thermistors, either in series or in opposite

arms of the bridge, one being mounted close to the heat source and the other outside the house. The bridge is initially balanced to provide the desired room temperature and under these conditions the heat supplied exactly balances the heat losses to the exterior of the house. When the outside temperature falls, the resistance of the outside thermistor rises, unbalancing the bridge. The out-of-balance voltage is used to control the supply of heat, raising the temperature of the heat source and of the second thermistor, until the bridge balance is restored.⁸

Figure 6 shows a simple arrangement that has been used in a manufacturing process to control the temperature of a heating oven to within $\pm \frac{1}{2}$ degree centigrade of any desired temperature in the region of 200 degrees centigrade. The variable resistor is a vernier adjustment to set the grid bias of the thyatron and hence the temperature at which control is to be effected; it may be fitted with a dial calibrated in degrees. A battery is used to supply the thermistor circuit to avoid the effects in the grid circuit of an unstable mains voltage.

A similar arrangement, Figure 7, but completely alternating-current operated, has been used⁹ to control the temperature of a domestic refrigerator to within ± 1 degree centigrade. At this lower temperature, the thermistor is more sensitive and the effects of mains variation are not serious. The arrangement is very effective, the refrigerator motor starting immediately the door is opened.

⁸ A. H. Taylor, "Thermistor Thermostats," *Electronics*, volume 23, pages 154-159; July, 1950.

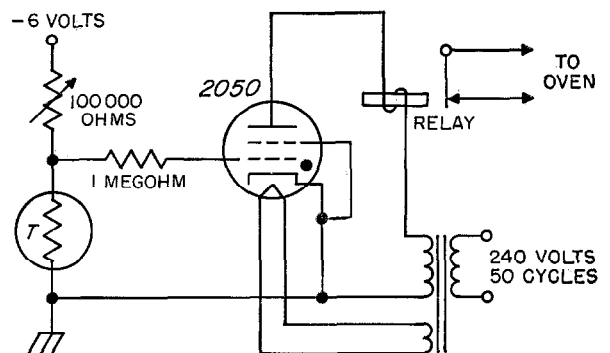


Figure 6—Temperature-control circuit. Type-F1512/300 thermistor is used; $R_e = 100\,000$ ohms at 20 degrees centigrade.

A more-elaborate circuit has been used¹⁰ in biological research to maintain the temperature of a metal chamber 8 by 8 by 4 inches (20 by 20 by 10 centimetres) at 38 ± 0.1 degrees centigrade while the conduction rate of a small mammalian nerve was being measured. The thermistor is used in a battery-powered bridge circuit followed by a two-stage valve amplifier. A portion of the oven heating element is connected in the plate circuit of the second amplifier stage and variations in plate current due to temperature changes sensed by the thermistor automatically keep the temperature within the prescribed limits. An interesting feature of this arrangement is the continuous variation of current through the heater element in contrast to the on-off switching employed in most control applications.

2.3 TEMPERATURE COMPENSATION

The negative temperature coefficient of a thermistor may be used to compensate for the positive temperature coefficient of metallic conductors such as the copper wire in coils or transformers et cetera. Over a small range of temperature the coefficient may be considered linear but in fact it decreases with increasing temperature, as shown in Figure 1.

In its simplest form, the temperature-compensation circuit consists of a metallic resistor and a thermistor in series; any increase in the value of the resistor due to a rise in ambient temperature is compensated for by the thermistor, which is also subject to the same ambient temperature. The temperature coefficient of copper is $+0.4$ percent per degree centigrade and that of a typical thermistor may be -4 percent per degree centigrade, so that if a thermistor having one tenth the resistance of the copper wire is placed in series with it, compensation will be obtained over a narrow range; but such a simple arrangement is not entirely satisfactory as only in rare cases does the law of the thermistor exactly fit that of the resistor. However, by shunting the thermistor with a further resistor, the desired characteristic may be obtained and good compensation may in most cases be ob-

tained over a reasonable temperature range. The resistor used in parallel with the thermistor should have a low temperature coefficient if it is subject to the same ambient temperature change.

Typical applications are the compensation for resistance changes due to temperature rise in the windings of transformers and the temperature compensation of electrical meters.

2.4 TRIGGER CIRCUITS

The negative-resistance portion of the voltage/current characteristic shown in Figure 2 can be used to provide a trigger action. If sufficient current is passed through the thermistor to make the voltage across it equal to E_{max} , then any small increase of the current or the ambient temperature reduces the thermistor resistance, which increases the current still further, and the resulting increase in current is very considerable, hence the term, trigger. The increase in current may be used to operate a control relay.

This principle has been applied to fire detectors and to sprinkler systems controlled by thermistors. The simplest arrangement is voltage

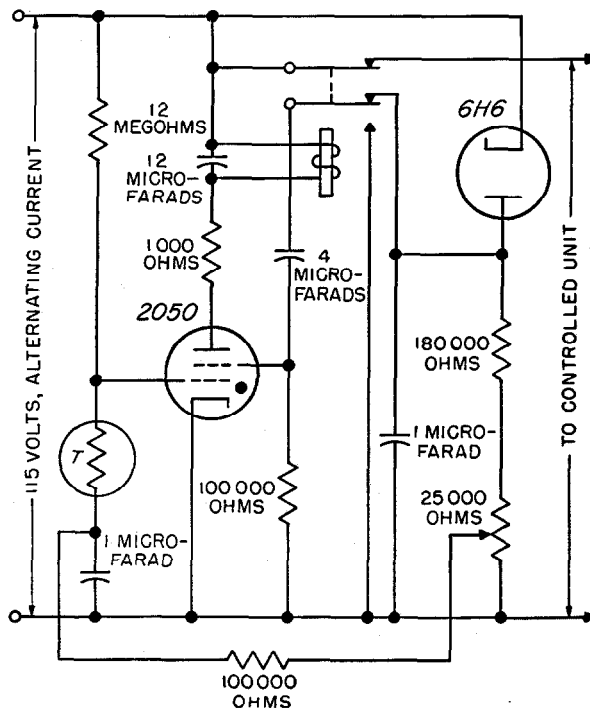


Figure 7—Temperature control circuit.

¹⁰ C. J. Dickinson, "Thermistor Continuous Temperature Control for Biological Research," *Electronic Engineering*, volume 21, pages 408-409; November, 1949.

source, thermistor, and warning light or bell all connected in series. When the ambient temperature rises beyond normal limits as a result of a fire, the thermistor triggers and operates the warning device. It is noteworthy that prevention of losses and reduction in fire-insurance rates where such devices are used may well repay the cost of installation many times over.

The same principle may be used to provide a control system operated by radio-frequency signals by first adjusting the circuit to the critical point and using the rectified radio-frequency signal to trigger the circuit. By careful adjustment of the circuit, fractions of a milliwatt may be detected and used to operate a relay.

2.5 TIMING DEVICES

The trigger characteristic previously referred to may be used to provide a time delay switch, and delay times from a few milliseconds to many minutes are possible.

A thermistor is wired in series with a relay and, until the thermistor reaches the critical temperature, the current rises slowly. At the temperature corresponding to E_{max} , a large increase in current occurs, which operates the relay. A certain amount of adjustment of this delay time may be obtained by means of a variable resistor in series with the circuit, as well as by choice of the time constant of the thermistor.

In the same way, a relay may be made to open a certain time after closing by connecting a thermistor in parallel with it, and this application may be useful for operations requiring controlled time such as, for example, photographic exposure, although it should not be used where very precise

limits on the timing interval are required unless compensation for ambient temperature changes is introduced.

By combining the series and parallel arrangement, a continuously operating timing switch may be constructed. In Figure 8, the relay $K/3$ is supplied by the resistor R_2 and the thermistor T_2 ; by suitable adjustment of R_2 , the relay may be made to operate after a predetermined time. When the relay pulls up, K_2 and K_3 operate, removing T_2 from the circuit and placing T_1 and R_1 in parallel; the hold-in current of the relay is maintained by R_3 . When T_1 reaches the critical temperature, its resistance rapidly falls, shunting the relay and allowing it to de-energize, and the supply is connected again via T_2 and R_2 . Thermistor T_2 has cooled in the meantime, and so the cycle repeats.

2.6 SURGE SUPPRESSION

Thermistors may be used for surge suppression by connecting them in series with the device to be protected; their high initial resistance prevents any short-duration surge from operating the device, whereas the operating voltage, although of much lower amplitude, is applied long enough to cause heating of the thermistor and consequent reduction of resistance to the point where the device operates. An example is the suppression of the surge of heater current in an alternating-or-direct-current radio receiver when it is switched on. The valve heaters in a receiver of this type are connected in series and supplied from the power mains through a voltage-dropping resistor. The cold resistance of the heaters is as low as a quarter of the hot resistance, and the large switching surge is apt to cause short valve life. A dial lamp is often used in series with the heaters and, due to the fact that it operates at a higher temperature than the heaters, the effect of the surge is even more serious in this case. The inclusion of a thermistor of 2000 to 3000 ohms cold resistance in series with the circuit reduces the surge to less than 10 per cent overload, and when it is hot, the voltage drop across the thermistor will be quite low. The same principles can be applied to reducing the starting current of small electric motors.

A further possibility is the protection of tungsten-filament projection lamps and lamps

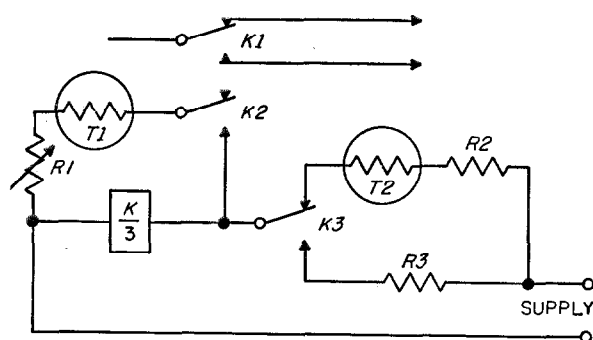


Figure 8—Switching circuit for timing device.

used for lighting in photographic studios, which have an inherent short life mainly due to the switch-on surge; suitable types of thermistors are being developed for this purpose.

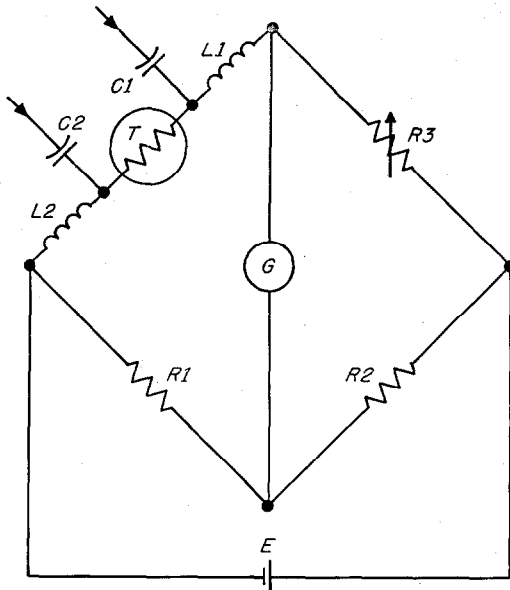


Figure 9—Power measurement circuit.

2.7 POWER MEASUREMENT

The resistance of a particular thermistor depends only on its temperature and, if the ambient temperature remains constant, its resistance is a function solely of the power dissipated in it. A thermistor thus provides a valuable means of power measurement, applicable to any waveform and any frequency up to 10 000 megacycles per second with types at present available commercially. The particular type chosen depends on the frequency.

For relatively low frequencies such as the measurement of the random noise power from an amplifier, an indirectly heated type is the most convenient and can be previously calibrated on direct current or 50-cycle-per-second alternating current. The heater coil is usually about 100 ohms and presents a constant load impedance; the resistance of the bead is then a measure of the root-mean-square power in the heater coil. A convenient method is to heat the thermistor first with direct current to a certain temperature and thus to a certain resistance. When the energy to

be measured is then applied to it, its temperature rises and its resistance drops. The direct-current energy is then reduced until the resistance and temperature regain their original values and since the heat dissipation depends only on the temperature, the reduction of the direct-current energy will be equal to the alternating-current energy supplied.

At high frequencies, the indirectly heated types cannot be used on account of the heater inductance, and it is more convenient to use the directly heated types in a bolometer bridge as shown in Figure 9. The thermistor is included in a normal bridge circuit but is isolated for radio frequencies by two radio-frequency chokes L_1 and L_2 . The thermistor is calibrated first on direct current and a curve drawn of input power versus resistance. Radio-frequency power is then applied to the thermistor by means of R_3 . The resistance value thus measured may be transferred to a power measurement from the curve.

2.8 AMPLITUDE STABILIZATION

For an oscillator to have good frequency stability and waveform it is necessary to provide some means of amplitude stabilization other than that of the curvature of the characteristics of the oscillator valves: Two methods are in general use; namely, the use of a diode rectifier feeding a variable bias voltage to one or more of the oscillator valves or a tungsten-filament lamp connected in the feedback circuit. The use of a thermistor leads to a simpler circuit than the diode rectifier, and it consumes much less power than the tungsten-filament lamp, besides working at a more convenient impedance level.¹¹ A typical method is shown in Figure 10, where any increase in oscillator output causes an increase in power dissipation in the thermistor, thus lowering its resistance, which reduces the input to the grid of $V1$ and tends to restore conditions to normal.

Thermistors have been extensively used to stabilize the levels of carrier- and audio-frequency amplifiers in wire-communication systems.¹² Negative-feedback-type amplifiers are

¹¹ L. A. Meacham, "Bridge Stabilized Oscillator," *Proceedings of the IRE*, volume 26, pages 1278-1294; October, 1938.

¹² J. H. Bollman, "Pilot-Channel Regulator for the K1 Carrier System," *Bell Laboratories Record*, volume 20, pages 258-262; June, 1942.

particularly suitable for this application and a simplified circuit is shown in Figure 11. An increase in output level causes more power to be dissipated in the thermistor, its resistance decreases and thus the amount of negative feedback increases until the level is restored to the original value. An indirectly heated thermistor may also be used in much the same way, the heater being connected to the amplifier output.

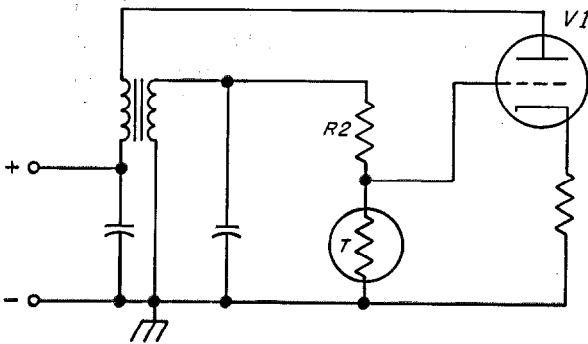


Figure 10—Oscillator amplitude-stabilization circuit.

2.9 LOW-FREQUENCY OSCILLATORS

An oscillator capable of generating small outputs at low frequencies may be constructed making use of the negative-resistance characteristic shown in Figure 2 between E_{max} and point D . By suitably biasing the thermistor so that it is operating in this region, the negative resistance may be utilized to compensate for the positive resistance of the inductors and if sufficiently negative will permit oscillation, provided the frequency is low enough to allow the thermistor to follow the cycles thermally. Experimental thermistors of extremely short time constant have been developed in the laboratory to cover the audible range and future developments may even extend the range, although these types are not yet commercially available.

It is also possible to use a thermistor as a relaxation oscillator in a resistance-capacitance circuit similar to a thyatron sawtooth oscillator by using the trigger properties mentioned previously.

2.10 VARIABLE RESISTORS

Indirectly heated thermistors are very suitable for use as variable resistors of low wattage

rating, and a smooth control over a range of 100 to 1 is readily obtainable.¹³ There is a slight time lag due to the thermal inertia of the thermistor but this may not always be of importance. The heater and the thermistor are isolated from each other electrically and therefore can be connected into separate circuits. Useful possibilities are suggested when it is considered that this provides a resistor whose value is proportional to a voltage or current without the need for an intermediate electronic/magnetic transfer device.

This principle may also be used for the remote control of resistors from a considerable distance. For example, an indirectly heated thermistor with a heater of 100 ohms resistance may have an element of several hundred thousand ohms and this may be varied by controlling the current in the 100-ohm heater circuit, which provides no problems over long distances.

2.11 MEASUREMENT OF VELOCITY OF FLUIDS

The velocity of a stream of fluid may be measured¹⁴ by its cooling effect on thermistors over a

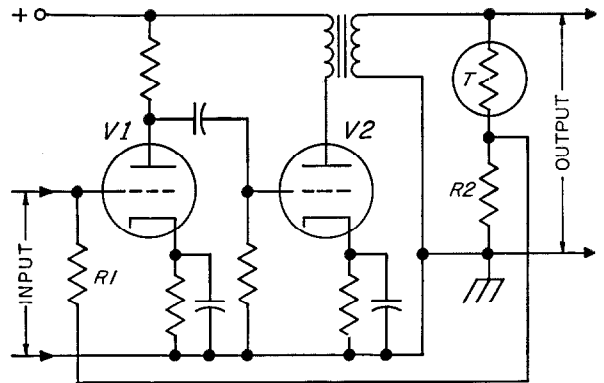


Figure 11—Negative-feedback circuit.

range of approximately 0.4 to 4 feet per minute. A bridge circuit is used with a compensating thermistor exposed to the same ambient temperature as the measuring thermistor, but screened from the cooling blast, in the same way

¹³ C. C. Roloff, "Thermo-Variable Resistors; Glass-Bulb Urdox Resistor," *Electrical Review*, volume 140, pages 315-316; February 14, 1947.

¹⁴ "Use of Thermistors to Measure Temperature and Wind Velocity," Office of Scientific Research and Development Report 4699, Department of Commerce, Washington 25, District of Columbia; Bureau of Publication Report PB 13102, Appendix C.

as the pitot and static tubes are used on airspeed indicators to compensate for changes in atmospheric pressure. A known power is dissipated in the thermistor and its resistance measured, after which it is exposed to the cooling blast. The greater the rate of flow, the greater the cooling and the higher will be the resistance. The power dissipated is then increased until the resistance is the same as it was without the cooling blast and this difference in power is converted to a temperature change and used to calculate the rate of flow by using the formulae below. These are reasonably accurate for air and water at an ambient temperature of 20 degrees centigrade, provided there is good thermal contact between the thermistor and the medium.

For air cooling,

$$\text{Cubic feet per minute} = \frac{1.75 \times W}{T},$$

for water cooling,

$$\text{Gallons per minute} = \frac{3.15 \times W}{\Delta T \times 10^3},$$

where W is the power difference in watts and ΔT is the temperature change in degrees centigrade.

Micro-meteorology, or the study of the variation of air temperature, humidity, and velocity over very small distances of the order of a foot or so, is an ideal field for the use of thermistors in this manner.¹⁵

2.12 VACUUM MANOMETERS

The thermal conductance of the surrounding medium has an effect on the relation between the power dissipated by a thermistor and its temperature, and this fact enables it to be used for vacuum measurement and control purposes. In these cases, the thermistor is really being used as a thermal conductance meter,¹⁶ because in a vacuum system the thermal conductivity of the gas is a function of pressure for any particular gas or mixture of gases. When a constant power is dissipated in the thermistor its temperature and hence its resistance is a measure of the gas

¹⁵ H. L. Penman and I. Long, "A Portable Thermistor Bridge for Micro-Meteorology Among Growing Crops," *Journal of Scientific Instruments*, volume 26, pages 77-80; March, 1949.

¹⁶ S. Dushman, "Manometers for Low Pressures," *Instruments*, volume 20, pages 234-239; March, 1947; also, *Product Engineering*, volume 18, pages 127-128; March, 1947.

pressure. The best results are obtained by using a bridge circuit and a measuring range of 10^{-5} to 10 millimetres of mercury is practicable.

Thermistor manometers are particularly useful where the gas used is capable of being decomposed thermally, and the rise in temperature of the thermistor can be limited, if necessary, to no more than 30 degrees centigrade above ambient temperature.

Thermistors have been used¹⁷ in this way to control the temperature of a vacuum furnace used for the manufacture of radio valves. A relay and an indicating meter, energized by thermistor manometers, provide automatic switching-on of tungsten heaters at a predetermined vacuum level and turns them off again if that level is later exceeded, thus safeguarding the heaters against burn-out due to excessive gas pressure.

3. Acknowledgments

The material for this paper has been compiled from a variety of sources, and particular thanks are due to Standard Telephones and Cables Pty. Limited, Sydney, for making available many laboratory reports on the subject.

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¹⁷ F. F. Davis, "Vacuum Furnace Control," *Electronics*, volume 21, page 81; May, 1948.

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Recent Telecommunication Development

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Germanium Positive-Gap Diode: New Tool for Pulse Techniques*

By A. H. REEVES and R. B. W. COOKE

Standard Telecommunication Laboratories, Limited; London, England

CERTAIN electroforming processes produce contact-type germanium diodes that exhibit negative-resistance properties. Such diodes may be effectively used to generate pulses corresponding to 20 megacycles per second with rise and fall times not greater than 0.01 microsecond. They may also be utilized as aperiodic double-stability devices, self-restoring trigger circuits, and binary counters.

. . .

1. History

The origin of the germanium positive-gap diode lies in a series of experimental researches conducted during 1950 and 1951 into the properties of normal germanium diodes. During electroforming tests that were part of these studies, there was occasionally observed a diode having an unusual property; a negative-slope resistance in a certain part of the forward (conducting) region of the diode characteristic. The possible practical importance of this discovery was at once realized; further studies were undertaken to find the best way of making such unusual diodes reproducibly and with the most desirable characteristics.

The probable importance of such a diode can be seen from the following considerations. For certain telephone circuits of both long and short distances and also for electronic switching in the telephone field, many engineers now consider it to be economically desirable to use up to 1000 speech channels or more on a single transmission path (metallic or radio) with one of the now-existing constant-amplitude pulse systems such as pulse-time or pulse-code modulation. To achieve this on pulse-time modulation would require a pulse rate of at least 6 megacycles per second and on 7-digit pulse-code modulation, 42 megacycles per second. To obtain the usual ad-

vantages of freedom from noise and crosstalk with either pulse-time or pulse-code modulation, the rise and fall times of the pulses would both have to be not more than about 0.01 microsecond.

The terminal equipment would have to produce and receive such very narrow pulses at the above high repetition rates and cheap repeaters would also have to be made to handle them. With pulse-code modulation, the terminal must also be able to code and decode pulses of this waveform. For switching, suitable distributor devices or networks would also have to be designed. The basic tools for nearly all these purposes are: (A) a nearly aperiodic pulse producer working at a repetition frequency up to about 50 megacycles per second and giving rates of rise and fall of the order of 0.01 microsecond; and (B) a similarly aperiodic double-stability device, or binary counter, working to limits of this same order. Owing to their relatively large time-constants, these requirements cannot be met by any existing hard-vacuum or gas-filled valves; some new tool altogether is needed.

A germanium diode with a negative-slope resistance in the conducting direction, however, introduced a completely new range of possibilities, for it was known that the self-capacitances of such diodes could be reduced to at least the same order as that of vacuum valves; while in the conducting direction the diode resistances could be much less than that of the usual valves. The negative-resistance slope, of course, gave the possibility of triggering action.

Tests on samples of positive-gap germanium diodes showed immediate promise of greatly reduced triggering time-constants as compared with valves. It was not long before some diodes were produced on a laboratory scale that nearly met the needed limits of pulse repetition frequency and completely met the needed rates of pulse rise and fall that have been quoted.

* Reprinted from *L'Onde Electrique*, volume 34, pages 32-37; January, 1954.

2. Construction and Characteristics

A whisker, preferably of silver containing a small quantity of *N*-type impurity such as arsenic, is placed in contact with a germanium

Figure 1 shows some typical characteristics, obtained in this case from a constant-current feed derived from a pentode valve, as indicated in Figure 2.

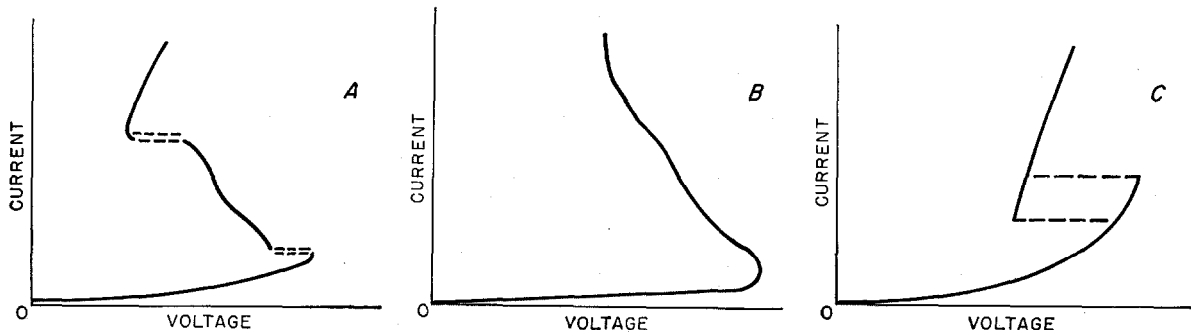


Figure 1—Several types of current-voltage characteristics that may be observed after the electroforming process. The characteristic at *C* indicates a usable positive-gap diode.

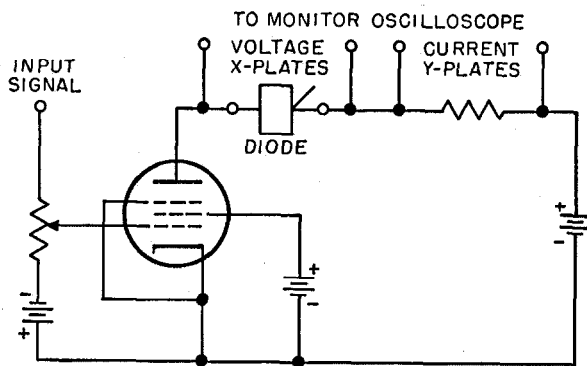


Figure 2—Constant-current source for observing current-voltage characteristics.

surface of *N*-type bulk material prepared by usual methods. Electroforming is then carried out with an unusually high current of the order of 1 to 5 amperes, normally in several bursts of a few seconds each, until the desired voltage-current characteristic is observed on a cathode-ray-tube monitor with a 50-cycle-per-second sweep. The electroforming current may be either alternating or direct current.

The desired shape of the forward (conducting) characteristic explains the name positive gap, for after correct forming there is a gap in the monitor picture, indicating at a certain voltage a double stability in the conducting direction due to a negative-resistance slope. The voltage difference on triggering from one stable position to the other is usually of the order of several volts. The corresponding current change can be 40 milliamperes or more.

Figure 3 shows the conditions for double stability of a positive-gap diode having the typical characteristic *OABC*, as traced at constant current in a simple series circuit with a battery *E* and a load resistance *R*. Double stability is obtained only if the load line cuts the curve *OABC* three times, as does the straight line *FG*. (The third intersection at *K* represents, of course, an unstable equilibrium point between the two stable points *L* and *M*.) The criterion for this is that the angle θ shall be greater than α , where $\cot \theta$ gives the load resistance *R* and *HJ*, the tangent to curve *OABC* at point *K*, must therefore not exceed a limiting value. It is also

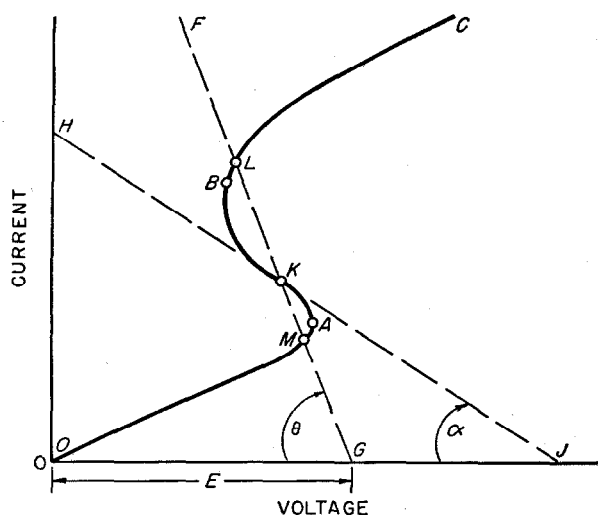


Figure 3—Double-stability characteristic.

necessary that there shall be a suitable value of E for the three intersections to be obtained.

Another typical trace is shown in Figure 4. In this case the trace line is missing between points A and B , showing that even at substantially infinite load slope resistance (from the pentode feed), double stability is still present, the gap AB representing a sharp triggering dur-

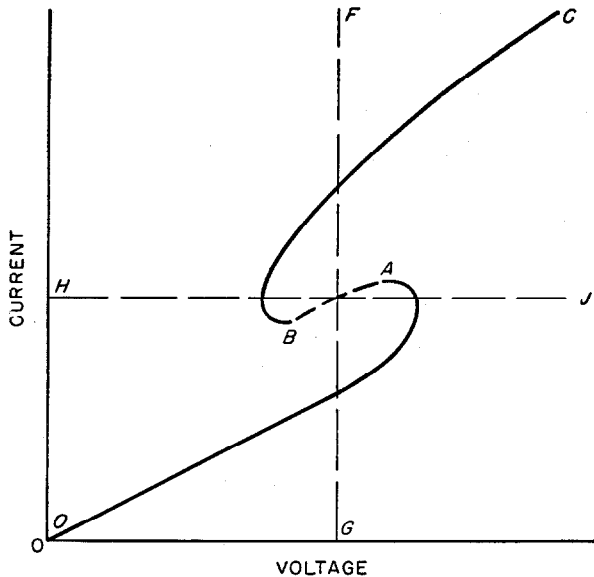


Figure 4—Double stability with zero-to-infinite load-resistance variation.

ing the 50-cycle-per-second sweep. It is to be noted that the current value at B is lower than that at A . It will be evident from the two load lines FG and HJ that this type of characteristic can show double stability at all values of load resistance from zero to infinity.

A characteristic of the shape $OABC$ (Figure 4) can be explained by assuming that there are at least two paths in parallel from the whisker to the bulk germanium, one showing a voltage-current characteristic similar to that of Figure 3 and the other having substantially a pure positive-slope resistance of a suitable value.

3. Diode as Narrow-Pulse Generator

Although by special devices pulses have been obtained for laboratory or test purposes of duration times of 0.001 microsecond or less, the apparatus involved is usually somewhat complex. A very simple and cheap method of obtaining pulses at up to, say, 20 megacycles per second

repetition frequency with a rise or fall time not exceeding about 0.01 microsecond is clearly of some practical interest for many applications. The positive-gap diode provides a convenient solution to this problem.

One very simple circuit approximately fulfilling these requirements is shown in Figure 5. The whisker of the positive-gap diode D is connected to ground via the small resistor $R1$, the battery, and the secondary winding of the transformer. The base of the diode is grounded via a load impedance shown here as an inductance L shunted by its self-capacitance and the resistance $R2$. The output pulses pass through capacitor $C2$. Capacitor $C1$ is very small so as to by-pass only the very-high-frequency components.

Assuming that pulses are required at a steady repetition frequency, a sine wave is applied at this required frequency to the input terminals. The battery voltage and resistor $R1$ are then adjusted so that the diode is swept through its triggering point twice in each period of the applied sine wave—once with rising current and once with falling current. With critical damping, two very sharp pulses are thus obtained across the load impedance for each sine-wave period, one positive and one negative. With a damping less than critical, a damped train will result at the output terminal—an adjustment that is sometimes useful. (See for example section 6, describing the use of this diode as a binary counter.)

With the circuit of Figure 5, the pulse rise and fall times could be estimated fairly accurately by the amount of shock excitation produced at various different natural frequencies of the tuned

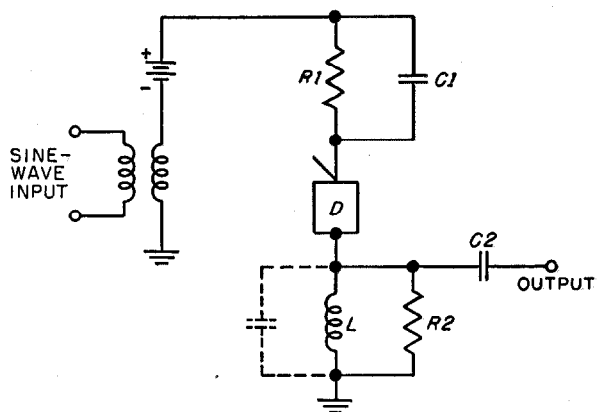


Figure 5—Repetitive-pulse generator.

load circuit. On some samples of diode the time of pulse rise was about 0.0065 microsecond and the time of fall about 0.016 microsecond. This difference in rise and fall time, put to a practical

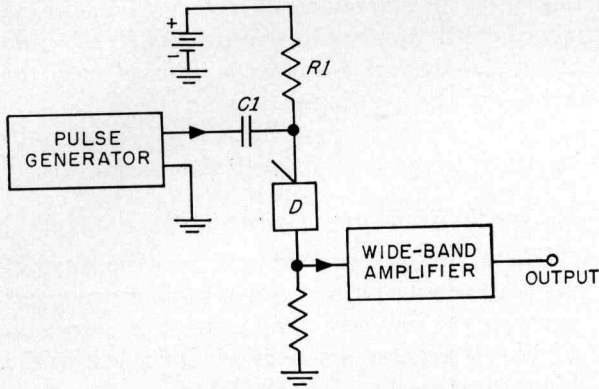


Figure 6—Aperiodic double-stability circuit. *

use in the binary counter circuit of section 6, can be explained in terms of the higher internal impedance of the diode in the off stability position as compared with that in the on position. A pulse-repetition frequency of up to 20 megacycles per second was obtained from a number of diode samples. A fairly typical peak output was found to be several volts at about 20 milliamperes, corresponding to a pulse-generator internal impedance of the order of 100 ohms.

Pulse generators based on positive-gap germanium diodes in the circuit of Figure 5 are now in use at Standard Telecommunication Laboratories; they have been found to be very practical and simple tools for experimental work.

4. Aperiodic Double-Stability Device

Figure 6 shows the first circuit using the positive-gap diode as an aperiodic double-stability device. The diode *D* was connected to ground at the whisker end via the small impedance *R1* and battery and at the base end via a small resistance of 80 ohms. A pulse generator produces alternating positive and negative pulses at repetition frequencies up to 20 megacycles per second. This function was performed by distorting a sine wave. The generator was coupled to the whisker of the diode via the

small capacitor *C1* and the output from the diode was observed on a cathode-ray tube at the output terminal, after raising the level sufficiently in the wide-band amplifier.

On some diode samples the results showed substantially aperiodic triggering backwards and forwards by the input pulses at frequencies up to 20 megacycles per second, the limiting value of the test equipment. On other diode samples, the triggering ceased at somewhat lower repetition frequencies.

To test the diodes at frequencies higher than 20 megacycles per second, the arrangement shown in Figure 7 was used. The base of the diode *D* was connected directly to ground, while the whisker was grounded via the resistor *R1* and the battery as before. A sine-wave oscillator, adjustable up to about 50 megacycles per second, was arranged to apply several peak volts across the diode via a transmission line and impedance-matching device. The output of the positive-gap diode was observed on a special high-frequency type of cathode-ray tube at the output terminal. A wide-band amplifier was not needed because the diode was chosen to have a large enough voltage difference between the two stability points to make further amplification unnecessary.

When the sine-wave voltage applied to the diode was insufficient to cause triggering, the output trace on the cathode-ray tube was nearly sinusoidal, as shown in Figure 8A. When diode triggering occurred, the kinks in Figure 8B were observed. The tuned circuit associated with the diode was too highly damped to suppress the harmonics represented by these kinks.

By this means two diode samples showed substantially aperiodic triggering, backward and

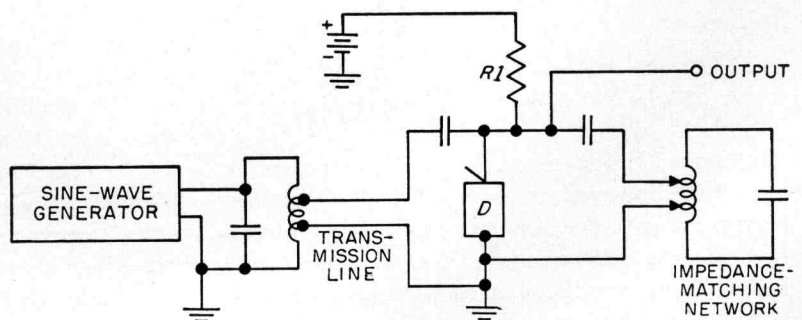


Figure 7—Double-stability testing circuit for use up to 50 megacycles per second.

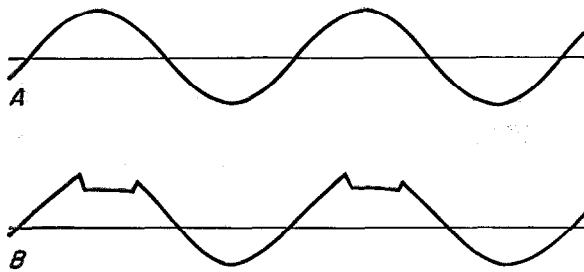


Figure 8—Waveforms on monitor for Figure 7. Kinks indicate double-stability conditions.

forward, at repetition rates of up to 43 megacycles per second.

5. Self-Restoring Trigger Circuit

As shown in Figure 9, the positive-gap diode with grounded base is connected to ground at the whisker end via resistor $R1$ and the battery. Uni-directional input pulses are applied across the diode in some suitable manner; for example by means of a pulse generator and transformer. If capacitor $C1$ be added across $R1$, suitable values of $C1$ and $R1$ will cause a relaxation oscillation to be produced—providing that the battery and resistance $R1$ are such as to bring the diode into a suitable negative-resistance region on its characteristic. If the voltage across the diode is changed slightly to bring the latter just outside the negative-resistance portion of the curve, os-

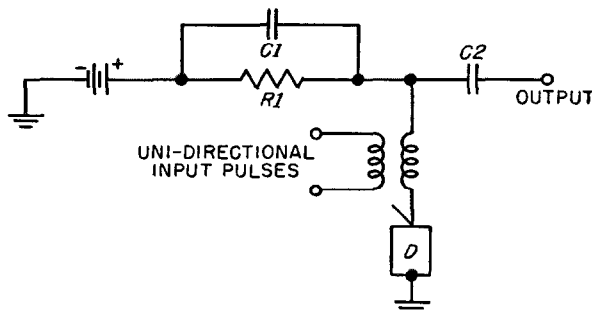


Figure 9—Self-restoring trigger circuit.

cillation will then cease. If the input pulses are arranged, however, to sweep the diode voltage once more into the negative-resistance region at their peaks, half-periods of the relaxation oscillation can be produced at each such input pulse, alternate oscillation half-periods taking place independently of input pulses and restoring the initial circuit conditions after an adjustable time

interval depending chiefly on the value of the time-constant of $C1-R1$.

There is thus a half-period of free relaxation oscillation timed by each input pulse; this is followed by the start of a self-restoring half-period, timed chiefly by the time-constant of $R1-C1$; and after this there is a quiescent interval until the arrival of the next input pulse.

In this way a convenient and stable self-restoring trigger circuit has been constructed.

6. Aperiodic Binary Counter

There are, of course, several possible ways of using the positive-gap diode as a binary counter; perhaps the simplest arrangement is shown in Figure 10, where only one diode is needed for the binary feature.

The first positive-gap diode $D1$ shown in the figure is used to convert the original signal wave to be counted (one having a rounded waveform) into the series of sharp pulses that is needed for the binary input; it is substantially the circuit already shown in Figure 5. Positive-gap diode $D1$ has its base grounded via a resistor $R2$ (100 to about 400 ohms, in typical cases). The whisker of $D1$ is connected to ground via the input transformer, resistor $R1$, and the battery. The small capacitor $C1$ is used as a by-pass for the sharp pulses produced by triggering. As in the case of Figure 5, therefore, the (assumed) rounded input waveforms from the input transformer, which are the signals to be counted, are converted into very narrow pulses at the output of capacitor $C2$. These narrow pulses then shock-excite the tuned circuit formed by inductance L and its effective self-capacitance shown in broken lines; the resonant frequency being in typical examples in the region of 100 megacycles per second. The circuit of diode $D2$ is completed by resistor $R3$ and the battery, shunted in the 100 megacycles-per-second region and above by the very small capacitor $C3$. The binary-counter output is obtained at the output terminals via capacitor $C4$.

The operation of the circuit will now be explained. In Figure 11, curve A is the assumed input waveform. Curve B represents the current waveform through diode $D1$. Curve C is the derivative of B . It will be noted that the negative-going pulses are smaller than the positive pulses; due to the fact that in a positive-gap diode the internal time-constant on triggering from "current-

low" to "current-high" is normally considerably less than the time-constant in the reverse sequence—a phenomenon shown in curve *B* as a steeper slope in the rise portion than in the fall portion. By virtue of small capacitor *C2*, the derived curve shown in *C* shock-excites the tuned circuit *L* and its distributed capacitance, giving a waveform represented as curve *D*. There are two damped trains; an initial train *a* due to shock excitation by the (larger) positive pulse and a following train *b* of relatively small amplitude due to the negative pulse, which is of much smaller peak value than the positive pulse.

The circuit constants of diode *D2* are adjusted so that triggering occurs on the first *negative* peak of the train in curve *D*. (Under the assumed initial conditions, triggering cannot occur on the larger positive peak because this peak is of the wrong sign for that purpose.) Re-triggering to the off position on the second positive peak is eliminated by the internal time-constant of the diode; if the resonant frequency of the tuned load circuit is 100 megacycles per second, for example, the interval between the first negative and the second positive peaks is only 0.005 microsecond, much too short for the diode to be triggered back to the current-low condition. The negative peak at the start of the second following train is too small in amplitude to cause triggering of diode *D2*; therefore this diode will remain in the current-high condition until the arrival of the positive peak at the beginning of the next damped train. Then, diode *D2* will be triggered back to its initial current-low condition.

Neglecting ripples due to pulses, the current through diode *D2* will therefore follow curve *E*.

It is thus clear that substantially aperiodic binary-counting action can be obtained in this way.

Counting has been achieved on some diode samples at input repetition frequencies up to 15

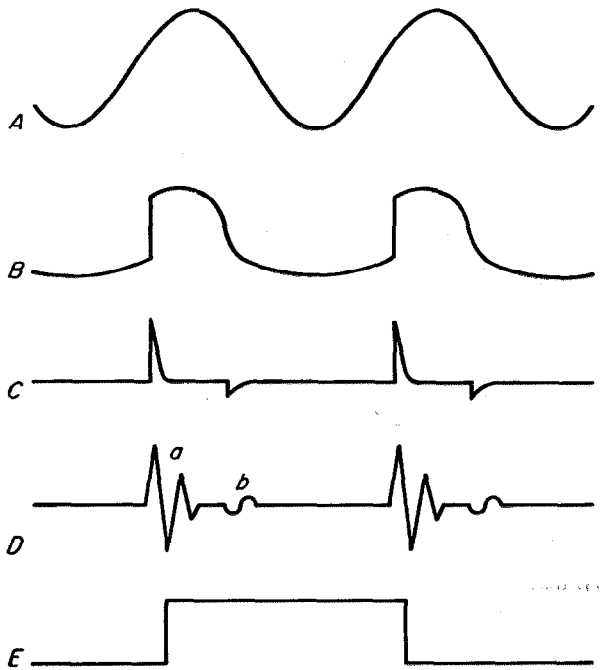


Figure 11—Waveforms for circuit of Figure 10.

megacycles per second. A tool such as this, in addition to its uses in telecommunication, should prove valuable in certain types of high-speed electronic calculators, et cetera.

7. Further Uses

This new form of diode can also be used as a high-frequency oscillator in an *L-C* circuit; a limit of about 300 megacycles per second was obtained from some samples at several milliwatts output power. The diode has also proved to be a successful tool in a very compact super-regenerative receiver in this same frequency range. It should be understood that there is, in fact, a family of such diodes having a range of characteristics and that differing types can be made and employed for various needs. As the purpose of this paper, however, is to discuss only pulse techniques, these further uses will not be described in detail.

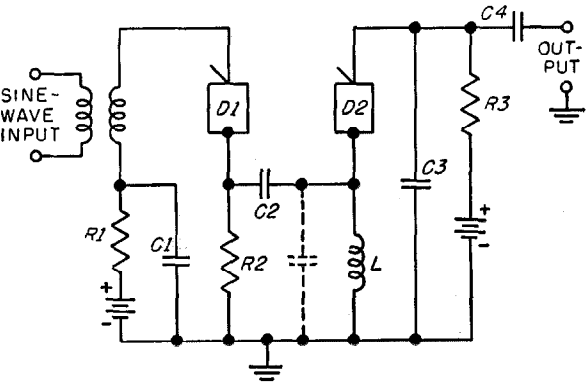


Figure 10—Aperiodic binary counter.

Broad-Band Reflectometers at High Frequencies

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A USEFUL instrument in any work with radio frequencies is a device for measuring the phasor sum or difference of two voltages, one being proportional to the current through a load and the other being proportional to the voltage across the load. Such an instrument will measure either the incident or the reflected power in a transmission line. These data may be combined to indicate the standing-wave ratio on the transmission line or the net power being transmitted. Other combinations of the basic "voltage" and "current" voltages can be used to measure magnitude and phase angle of the load impedance or to indicate load resistance or conductance independently of the reactive component.

Well-known forms of this instrument are the reflectometer (or directional coupler) and standing-wave indicator. The variation described here differs primarily in the use of resistive rather than reactive coupling elements, thus eliminating the dependence of sensitivity and insertion loss on frequency. Improved techniques for low radio frequencies permit broad-band use at constant sensitivity from 300 kilocycles per second to 30 megacycles. Applications are extended to provide the sign as well as magnitude of error for servo control; to indicate separately the phase, magnitude, resistance, or conductance of the load; or to measure directly the standing-wave ratio or net delivered power.

• • •

1. Reflectometer Theory

The voltage and current at any point in a transmission line may be expressed as the phasor sum of two component waves (see Figure 1) traveling in opposite directions, each having the instantaneous voltage-current relation $e = Z_0 i$, but the currents being taken in opposite directions. The incident wave that is traveling from the source toward the load, therefore, may be expressed at any point by $e_i = Z_0 i_i$ and the reflected wave moving from the load toward the source by $e_r = -Z_0 i_r$.

The phasor relations on which the reflectometer and related wave detectors are based may be obtained from Figure 1.

$$e_t = e_i + e_r, \quad (1)$$

$$Z_0 i_t = e_i - e_r, \quad (2)$$

where

e_t = voltage to ground from a given point on the transmission line

i_t = current at same point on the transmission line

e_i = voltage of incident wave at same point

e_r = voltage of reflected wave at same point

Z_0 = characteristic impedance of the same transmission line.

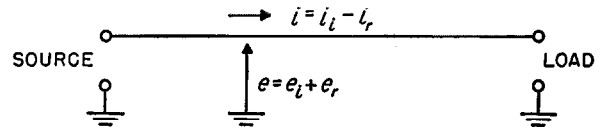


Figure 1—Voltage and current along a transmission line.

From the sum and difference of (1) and (2) may be derived

$$e_i = (e_t + Z_0 i_t)/2, \quad (3)$$

$$e_r = (e_t - Z_0 i_t)/2. \quad (4)$$

The magnitudes of e_i and e_r are constant (except for attenuation) at all points on the line, the voltage standing-wave ratio being

$$\frac{|e_i| + |e_r|}{|e_i| - |e_r|} = \frac{|e_t + Z_0 i_t| + |e_t - Z_0 i_t|}{|e_t + Z_0 i_t| - |e_t - Z_0 i_t|}. \quad (5)$$

Accordingly, if suitably adjusted voltages proportional to line current and voltage are added (or subtracted) vectorially, the rectified result will yield by (3) and (4) a direct voltage proportional to the incident (or reflected) voltage on the line. This is evident from Figures 2 and 3.

1.1 STANDING-WAVE-RATIO MONITOR

By arranging the detecting rectifiers to produce one polarity of direct current for incident-

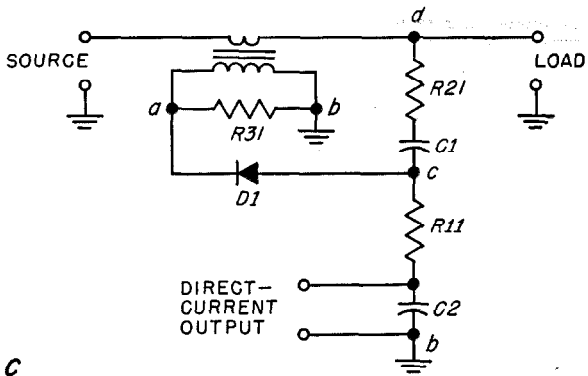
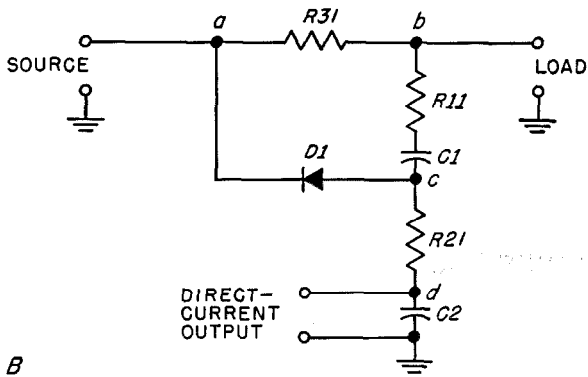
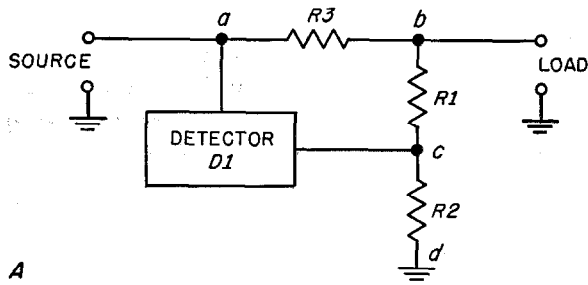


Figure 2—Detection of an incident wave. The rectifiers $D1$ are suitable for radio frequencies and the blocking capacitors $C1$ and $C2$ present a low impedance at the operating frequency. A is the simplified schematic circuit, B is the current-resistor form, and C is the current-transformer type of circuit.

wave output and the opposite polarity for the reflected-wave output, a potentiometer connected between the two can be adjusted to produce a null for any chosen ratio of incident-to-reflected output. This is shown in Figure 4. Such a null circuit gives a definite polarity indication of whether the voltage standing-wave ratio on a line is greater or less than a preselected

value and may be used as a limit-bridge type of monitor to accept or reject a given load adjustment.

1.2 POWER METER

If square-law detectors are provided, the difference between the squared values of the incident-wave output and the reflected-wave output of the detectors is a direct measure of total power delivered to the load.

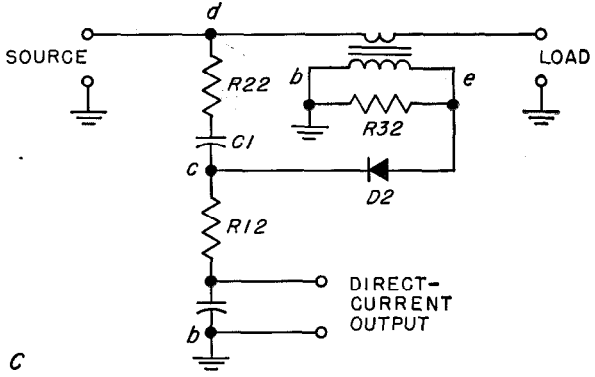
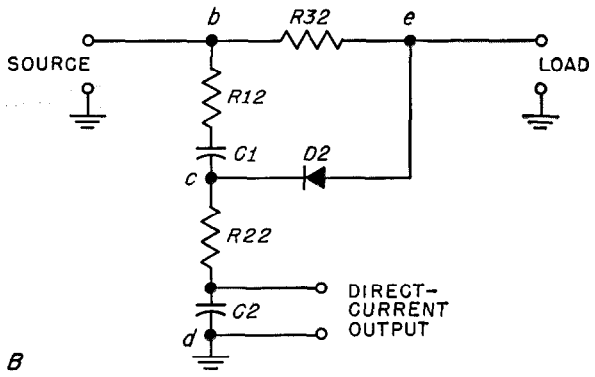
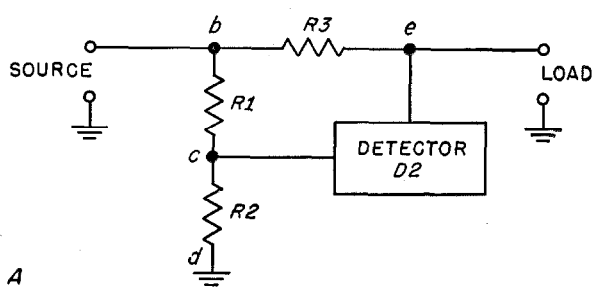


Figure 3—Detection of a reflected wave using a detector and blocking capacitors suited to the operating frequency. A is the schematic circuit, B is for a current-resistor source of voltage, and C is for a current-transformer arrangement.

For the more-usual case of linear detectors (where I and R are obtained, rather than I^2 and R^2), it is also possible to obtain a direct indication of power by using a linear combination of incident and reflected outputs to operate a direct-current wattmeter as is shown in Figure 5.

Since the wattmeter indication is proportional

to the product of the currents in the two coils, $I + R$ is supplied to one coil and $I - R$ to the other. The product is obviously $I^2 - R^2$ or the net power delivered to the load.

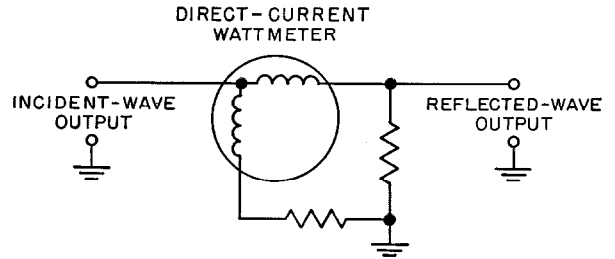
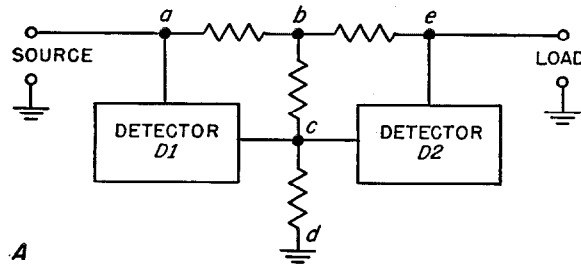


Figure 5—Method of obtaining a direct indication of power.

2. Measurement of Load Resistance, Conductance, or Phase Angle

2.1 RESISTANCE BRIDGE

When the conditions (A), (B), (C), and (D) given below are imposed on the reflected-wave detector, the circuit of Figure 6 is obtained. It

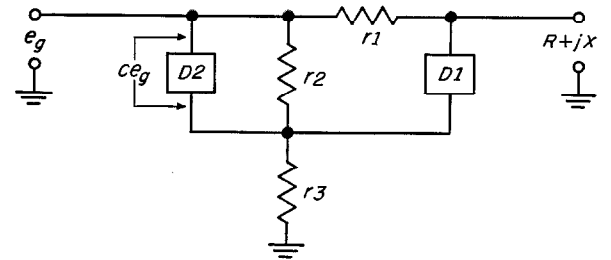
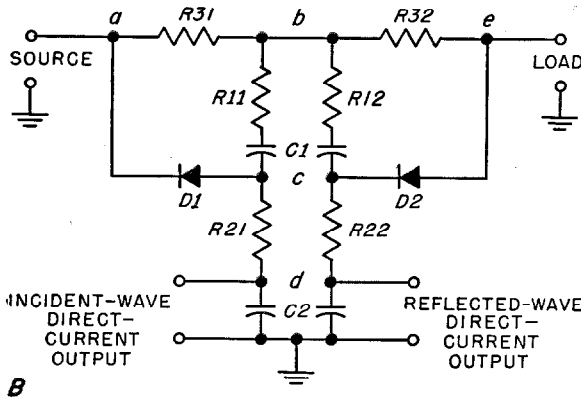


Figure 6—Resistance bridge.

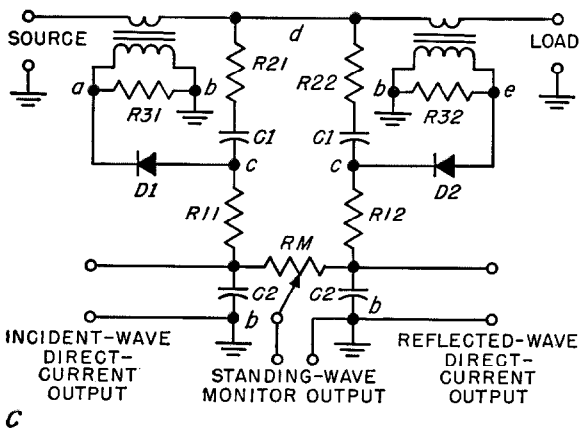


Figure 4—Combined reflectometers or standing-wave monitor; A being the simplified; B, the current-resistor; and C, the transformer arrangements.

responds only to variations in the equivalent series resistance of the load, the null obtained being unaffected by the series reactance. The polarity of the output indicates whether the load series resistive component is greater or less than a chosen value. This circuit is of particular interest if the resistance and reactance of the load are to be adjusted independently.

$$e_g = (R + r_1 + jX)i$$

$$c = \frac{1}{2} \frac{r_1}{Z_0 + r_1} \quad (A)$$

$$D_1 = |ce_g - r_1 i| \quad (B)$$

$$D_2 = |ce_g| \quad (C)$$

$$r_2 + r_3 \gg R + r_1. \quad (D)$$

$$D_1 = \left| \frac{1}{2} \frac{R + r_1 + jX}{Z_0 + r_1} r_1 i - r_1 i \right|$$

$$D_2 = \left| \frac{1}{2} \frac{R + r_1 + jX}{Z_0 + r_1} r_1 i \right|$$

$$D_1 = \frac{r_1 i}{2(Z_0 + r_1)} \left| R + r_1 - 2Z_0 - 2r_1 + jX \right|$$

$$D_2 = \frac{r_1 i}{2(Z_0 + r_1)} \left| R + r_1 + jX \right|$$

$$D_1 - D_2 = \frac{r_1 i}{2(Z_0 + r_1)} (|R - r_1 - 2Z_0 + jX| - |R + r_1 + jX|).$$

For $R = Z_0$,

$$|Z_0 - r_1 - 2Z_0 + jX| = |Z_0 + r_1 + jX|.$$

$$D_1 - D_2 = 0 \text{ for all values of } X.$$

For $R > Z_0$,

$$|Z_0 + \Delta - r_1 - 2Z_0 + jX| < |Z_0 + \Delta + r_1 + jX|,$$

$$D_1 - D_2 < 0 \text{ for all values of } X.$$

For $R < Z_0$,

$$D_1 - D_2 > 0 \text{ for all values of } X.$$

2.2 CONDUCTANCE BRIDGE

A similar modification, shown in Figure 7, permits determination of the load conductance or shunt resistance independent of parallel susceptance or reactance.

$$D_2 = \frac{i}{G_1} = \frac{i_L + G_2 e_L}{G_1} = \frac{e_L}{G_1} (G + G_2 + jB).$$

$$\text{Let } c = 2 \frac{G_0 + G_2}{G_1},$$

$$\begin{aligned} D_1 &= ce_L - \frac{i}{G_1} \\ &= \frac{2G_0 + G_2}{G_1} e_L - \frac{G + G_2 + jB}{G_1} e_L \\ &= \frac{e_L}{G_1} (2G_0 - G + G_2 - jB). \end{aligned}$$

$$D_1 - D_2 = \frac{e_L}{G_1} (|2G_0 - G + G_2 - jB| - |G + G_2 + jB|)$$

for $G = G_0$.

$$D_1 - D_2 = \frac{e_L}{G_1} (|G_0 + G_2 - jB| - |G_0 + G_2 + jB|) = 0$$

for all values of B .

For $G > G_0$,

$$\begin{aligned} D_1 - D_2 &= \frac{e_L}{G_1} (|G_0 + G_2 - \Delta - jB| \\ &\quad - |G_0 + G_2 + \Delta + jB|) \\ &< 0 \text{ for all values of } B. \end{aligned}$$

For $G < G_0$,

$$D_1 - D_2 > 0 \text{ for all values of } B.$$

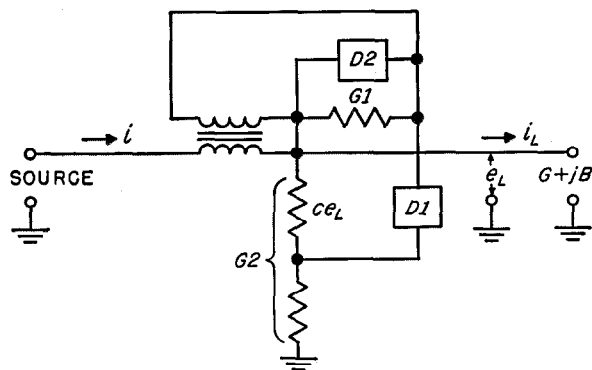


Figure 7—Conductance bridge.

2.3 PHASE BRIDGE

By the addition of suitable phase-shifting networks, shown in Figure 8, a circuit responsive only to the phase angle of the load impedance is obtained. The balance conditions for this bridge are independent of frequency and impedance magnitude as shown by the balance equations. The sensitivity is necessarily somewhat affected, but by proper choice of parameters a practical frequency range of 60 to 1 can be covered with a sensitivity variation of about 4 to 1. For minimum variation with frequency, corresponding reactance and resistance values should be numerically equal (45-degree phase relation) at the geometric mean frequency (that is $X_3 = R_3$, et cetera) and the voltage arm X_5 - R_5 should be tapped so as to make $e_{ab} = e_{cf} = e_{de}$ for a matched load.

$$\frac{X_3}{R_3} = \frac{X_4}{R_4} = \frac{X_5}{R_5} = \frac{X'_6}{R'_6} = \tan \alpha$$

$$R_1 = R_2, X_3 = X_4, \text{ and } R_3 = R_4.$$

The following approximations are assumed in the discussion.

$$e_g = e_i, i_g = i, R_3 \gg R_1, \text{ and } R_5 \gg R_1.$$

$$e = e_L = e_g$$

$$\begin{aligned}
i &= i_L = i_g \\
e &= i(M \exp [jB]) \\
e_{ac} &= iR_1 = -e_{ec} = -iR_2 \\
e_{cg} &= e \\
e_{bc} &= (e_{ac} \cos \alpha) \exp [j\alpha] \\
e_{dc} &= (e_{ec} \cos \alpha) \exp [j\alpha] \\
e_{cf} &= (e_{cg} \sin \alpha) \exp [j(90^\circ - \alpha)] \\
e_{bc} &= (iR_1 \cos \alpha) \exp [j\alpha] \\
e_{dc} &= (iR_1 \cos \alpha) \exp [j\alpha] \\
e_{cf} &= i(M \sin \alpha) \exp [j(90^\circ + B - \alpha)] \\
D_1 &= |e_{bf}| = |e_{bc} + e_{cf}| = |i(R_1 \cos \alpha) \exp [j\alpha] \\
&\quad + i(M \sin \alpha) \exp [j(90^\circ - \alpha)]| \\
D_2 &= |e_{df}| = |e_{dc} + e_{cf}| = |-i(R_1 \cos \alpha) \exp [j\alpha] \\
&\quad + i(M \sin \alpha) \exp [j(90^\circ + B - \alpha)]| \\
D_1 &= |i \exp [j\alpha]| |R_1 \cos \alpha + jM \sin \alpha \exp [jB]| \\
D_2 &= |i \exp [j\alpha]| |R_1 \cos \alpha - jM \sin \alpha \exp [jB]| \\
\exp [jB] &= \cos B + j \sin B \\
D_1 &= |i \exp [j\alpha]| |R_1 \cos \alpha - M \sin B \sin \alpha \\
&\quad + jM \cos B \sin \alpha| \\
D_2 &= |i \exp [j\alpha]| |R_1 \cos \alpha + M \sin B \sin \alpha \\
&\quad - jM \cos B \sin \alpha|
\end{aligned}$$

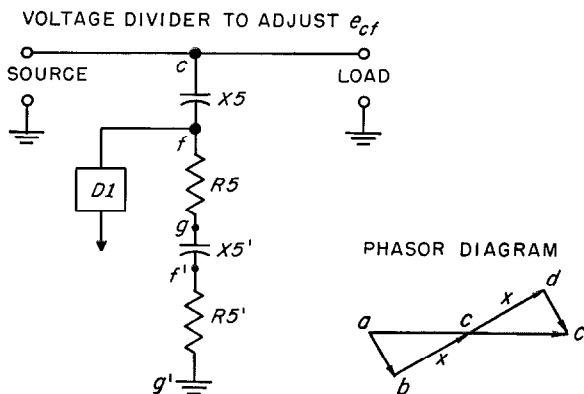
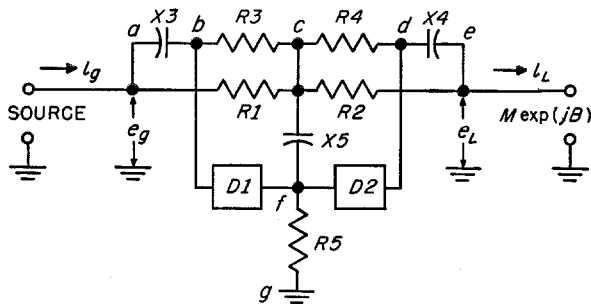


Figure 8—Phase bridge.

For $B = 0$ (load resistive), $\sin B = 0$, and $D_1 - D_2 = 0$ for all values of M .

For $0 < B < 90$ degrees (inductive), $0 < \sin B < 1$, and $D_1 - D_2 < 0$ for all values of M .

For $90 < B < 180$ degrees (capacitive), $0 > \sin B > -1$, and $D_1 - D_2 > 0$ for all values of M .

As a companion to the phase bridge, a simple modification of Figure 2 with D_1 replaced by separate (oppositely poled) detectors, one across ab and the other across bc , provides an indication of magnitude of load impedance that is independent of phase angle.

3. Construction

The following methods may be used to obtain the required results at frequencies up to 30 megacycles. At higher frequencies, some developmental work may be necessary.

Resistors appear in all the circuits above and (since phase is measured) must remain resistive over the required frequency range. With average carbon resistors, only values of resistance below 150 ohms can be used in the case of the 2-watt size and values below 250 ohms in the 0.5 watt size. To obtain large values of resistance, it is necessary to use these smaller values in series. Some cancellation of the capacitances occurs because of the lead inductance and in some cases resistors as large as 1000 ohms may be used as building elements. Extreme care is necessary in the latter case because long leads and metallic terminal posts act as capacitances to ground. In general, the leads should be as short as possible with no use of terminal posts except with resistors of less than 100 ohms. If a resistor of 20 000 ohms is needed, the problem is acute and a configuration must be found in which there is a proper balance of (A) capacitance from the radio-frequency line to the resistors in the top of the string, (B) capacitance among the resistors in the string, (C) capacitance between the bottom resistors and ground, and (D) distributed capacitance of the string to ground. One successful method of doing this is an arrangement with 20 resistors each of 1000-ohm 0.5-watt rating soldered together in series as shown in Figure 9. The capacitance from the radio-frequency side may be adjusted by varying the position of a wire attached to the radio-frequency line. The capacitance among resistors

may be varied by the spacing among them, and the capacitance to ground may be varied by moving the assembly with respect to ground.

There is a lower limit to the value of resistance that may be used as well as an upper limit. This

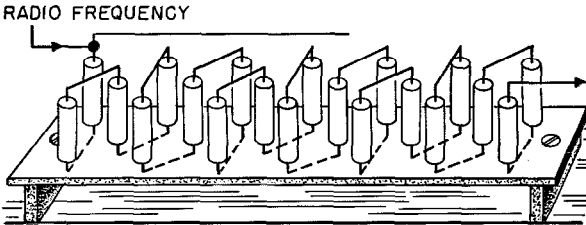


Figure 9—Disposition of 20 resistors to reduce capacitive coupling to a minimum value. A final balance is obtained by dressing with upper rigid lead that is connected to the radio-frequency line.

is imposed by the series inductance due to the physical length of a unit between points of attachment and includes lead length. Thus to obtain 1 ohm, it is necessary to use 10-ohm resistors in parallel mounted on metallic sheets to minimize leads.

In those cases in which power is not a problem, there are some carbon-film resistors of comparatively high value that are satisfactory. This situation is not a usual one. Instead, there occurs a need for components that will handle more than the 2-watt capacity of the carbon type. The precious-metal film-type 5-watt size in resistance from 60 ohms to 200 ohms provides for such a requirement. If the power dissipation approaches 50 watts, then the noninductive wire-wound type with calibrated frequency characteristics can be used. The latter may be adjusted conveniently for a pure resistance value by sliding a metallic tube from the radio-frequency end over the outside of the resistor.

The method of determining whether a resistor or resistor assembly is adjusted correctly for little or no phase angle is that of final performance in a circuit. Before being placed in a circuit, the resistors and sections of resistors should be measured on a radio-frequency bridge to insure that the reactive component is less than 2 percent of the magnitude of the resistor. The effect of the final ground configuration should be included in the measurements. When capacitance from the radio-frequency lead to the top of the resistor is

used, care should be taken to deduct the added direct capacitance to ground from the measurement.

Final performance is checked by using a flat resistive load and observing the outputs of the detectors for a fixed radio-frequency input power over a frequency range. The output should remain constant and should not vary continuously with frequency. If variations occur, the fault usually may be traced to the divider.

In all the circuits above, one or more series resistors are used to measure current. An extremely convenient substitution for each of these elements is a current transformer in which the secondary is isolated from the primary radio-frequency power and yet produces an output voltage proportional to the current in the load. The circuits of Figures 2C, 3C, 4C, and 7 are then practical. The advantages are increased sensitivity and the convenience of having all direct-current outputs at radio-frequency ground potential. The detectors are in a more-favorable physical position and are less likely to pick up radiated radio-frequency power that causes confusing unbalances.

The construction of the current transformers deserves some description. The best cores for broad-band use are the toroids made up of 1-mil (0.3-millimeter) superpermeability tape. The secondary winding¹ consists of a relatively few turns of flat wire put on by hand. Each turn is placed on top of the previous turn to form a spiral. The insulation is kept at a minimum to reduce leakage. In general, as many secondary turns are used as will produce self-resonance at the center frequency of the range desired. The primary consists of the radio-frequency line passing through the center of the core. A shield

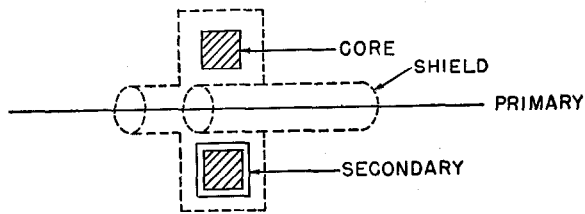


Figure 10—Shielded current transformer.

¹ D. Maurice and R. H. Minns, "Very-Wide Band Radio-Frequency Transformers," *Wireless Engineer*, volume 24, pages 168-177, 209-216; June and July, 1947.

is used as shown in Figure 10 to reduce leakage and to eliminate electric-field coupling.

The size of the core depends on the power it must dissipate. An average design allows about 1 percent of the primary radio-frequency power to be lost in the core. For 100 watts of primary power, a core of 0.25-inch (6-millimeter) square cross section and 1-inch (25-millimeter) diameter will handle 1 watt. The total open-circuit impedance of a winding is independent of the core size. Large cores require fewer turns than small cores, and the smaller windings produce the same open-circuit impedance because of the decreased magnetic-path length. This, plus the increased efficiency of smaller cores due to a more-equal distribution of core and copper losses, encourages economy of design up to the point where physical size and *leakage reactance* become limitations.

The open-circuit impedances at the two ends of the desired frequency range determine the resistive loading that may be used across the secondary. At the low end, the winding impedance is inductive and at the high end it is capacitive. Resistances across the secondary winding of 10 to 15 percent of the total open-circuit impedance will provide voltages that are constant with frequency and have very little phase shift referred to the current in the primary.

The leakage reactance referred to above may be defined as the flux around any one wire or turn that does not link nearby wires or turns. It is comparable to a series inductance and can usually be kept to a small fraction of its maximum value (the self-inductance of a wire, turn, or winding by itself). Where leakage is large, self-resonance with stray capacitance occurs at a low frequency. The results are a severe drop in open-circuit impedance of the winding at the high-frequency end (the magnitude of $R^2 + X^2$ is about the same on both sides of the center frequency), a drop in output, and undesirable phase shift. These conditions are to be avoided. Where leakage is small, the phase shift produced is sometimes evident, but in most cases can be compensated for in the final adjustment of the voltage divider.

Some practical values of resistances to be used in the above circuits can be given. In Figures 2B and 3B, e_{ab} may be obtained from 1 ohm constructed as previously mentioned. For 100 watts of radio-frequency power, the current

into 50 ohms is 1.41 amperes and the voltage across 50 ohms is 70.7 volts making $e_{ab} = 1.41$ volts. To get $e_{bc} = e_{ab} = 1.41$ volts, the ratio of bc/bd is $1.41/70.7$ or $1/50$. If an 18-ohm resistor is used to obtain e_{bc} then $bd = 50 \times 18 = 900$ ohms. This divider can be made up of one 18-ohm resistor, one 130-ohm resistor, and the rest are 150-ohm units. The total divider is 898 ohms. With 70.7 volts across it, the wattage to be dissipated is $5000/900$ or about 55 so that each of the 150-ohm resistors can be 1 watt. The capacitors shown are all for direct-current blocking and radio-frequency bypassing. Any convenient size serving these functions is satisfactory. The detectors may be crystals of the germanium type. A 1-milliamperer meter may be used as an indicator with 1000 or 2000 ohms in series. The direct-current return is through the load or generator or through a radio-frequency choke if both the former are open direct-current circuits.

Figures 4B and 4C use two dividers and the resistance of each is double that for Figures 2 and 3 to prevent loading. The direct-current potentiometer at the ground level may be 5000 or 10 000 ohms. In this case, all the current used to generate e_{ab} does not go through the R_{32} resistor because of the shunting effect of the dividers bd . Therefore, e_{ab} does not equal e_{bc} exactly and some correction is necessary if exact answers are required.

In the current-transformer form, the resistors used to obtain e_{ab} and e_{bc} depend on the properties of the cores. A convenient and inexpensive core for the 2- to 26-megacycle range is made of Crolite 70, a ferrite, and is 1.1 inch (28 millimeters) in outside diameter, 0.6 inch (15 millimeters) in inside diameter, and 0.25 inch (6 millimeters) in thickness. With a 20-turn secondary and a 1-turn primary, the resistance across the secondary may be as large as 300 ohms. This resistance divided by the square of the turns ratio is the resistance reflected in the primary and is 0.75 ohm. For 1.41 amperes again, the voltage is 1.06 in the primary and 20 times that or 21 volts across the secondary. The wattage rating is $(21)^2/300$ or about 1.5 watts so that two 150-ohm 1-watt resistors are satisfactory.

The divider ratio is now $21.1/70.7$ or ap-

proximately 2/7. If the losses in the core are considered, the ratio becomes almost exactly 2/7. The divider can be made up of seven 150-ohm 1-watt resistors with the tap at the 300-ohm junction as measured from ground.

If a current transformer is used in place of the series resistors, the voltage across the detectors is about 24 decibels greater and the sensitivity is increased by the same amount. The insertion loss of the circuits is not increased so that continuous monitoring is still possible. It may be necessary however to replace the crystals with tubes to prevent burnout.

The performance of the above circuits has been proved successful over as large a frequency range as 80 to 1. As a standing-wave indicator, the accuracy is within a few percent. There is sufficient power output for stability and excellent

control of automatic tuning devices. Because of the low insertion loss, there is no need for the switching in and out of radio-frequency circuits. In fact, several variations for various purposes may all be left permanently attached. The chief annoyance of the circuits is the care and familiarity required in lining up a new layout where effects due to stray capacitances of as little as 1 micromicrofarad ($-j1500$ ohms at 30 megacycles) may defy all sensible attempts at correction. A helpful guide in such cases is the use of a small variable capacitance (3 to 30 micromicrofarads) that may be placed by hand across various parts while the circuit is operating to indicate whether parallel or series reactances are needed. A completed design however may be easily specified with standard tolerances so as not to require further adjustments of this sort.

Pulse Generator for the Mechanoelectronic Telephone Switching System

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IN EACH mechanoelectronic telephone switching exchange, a special group of circuits delivers the pulses that are necessary to perform the switching operations that were described in a previous article.¹

It is obvious that these pulses must be present all the time, for any interruption or fault in pulse production will have serious and immediate consequences in the proper operation of the entire exchange. A principal objective in the development of this generator was, therefore, its reliability.

1. Functions

The generator must be able to deliver 5 group cycles of principal pulses designated *a*, *b*, *c*, *d*, and *e*; having the lengths and recurrence frequencies indicated in Table 1. It must be realized

TABLE 1
GROUPS OF PRINCIPAL PULSES

Group	Number of Sources	Length of Pulses in Microseconds
<i>a</i>	5	200
<i>b</i>	5	1000
<i>c</i>	4	200
<i>d</i>	11	200
<i>e</i>	3	200

that group *a* is developed by 5 separate sources, each successively producing a pulse of 200-microsecond duration in a cycle that is 1000 microseconds long. Similarly, *b* has 5 sources that produce 1000-microsecond pulses to fill a 5000-microsecond cycle.

Due to differences among the circuits driven by these pulses, 3 types are required as shown in Table 2.

¹J. Kruihof and M. den Hertog, "Mechanoelectronic Telephone Switching System," *Electrical Communication*, volume 31, pages 107-150; June, 1954.

There are 2 auxiliary pulses in each 200-microsecond interval. One starts at the beginning and the other terminates at the end of each time unit. They are concerned with impulse regeneration and have the characteristics given in Table 3.

TABLE 2
CHARACTERISTICS OF PULSES

Type	Group	Base Circuit Voltage Without Pulse	Voltage Amplitude of Pulse	Top Circuit Voltage With Pulse	Rise Time in Microseconds
<i>N</i>	<i>a, b, c, e</i>	-40	+16	-24	20
<i>N</i>	<i>d</i>	-40	+24	-16	20
<i>P</i>	<i>a, b, c, e</i>	0	+16	+16	20
<i>P</i>	<i>d</i>	0	+24	+24	20
<i>R</i>	<i>a, b, c, d, e</i>	-110	+60	-50	10

2. Reliability

Several schemes may be used to ensure a reliable source of pulses.

Two generators may be provided with a suitable switching system that will replace a faulty generator with a good one.

Two disadvantages of such a system are that some time is necessary to effect the cutover, during which the pulses are either distorted or absent, and power is wasted in the standby for it must be kept under current to minimize cutover time.

TABLE 3
AUXILIARY PULSES

Type	Circuit Voltage Without Pulse	Circuit Voltage With Pulse	Pulse Position in Microseconds in 200-Microsecond Time Unit		Rise Time in Microseconds	Decay Time in Microseconds
			Start	Stop		
<i>d₃</i>	-48	-8	0	25	10	20
<i>d₂</i>	0	-24	175	200	10	20



These two of the three bays house the pilots, distributors, amplifiers, and control circuits.

A second scheme is to have two generators working in parallel. Their output signals go to the same points at the same time. Both generators are controlled by another circuit that is able to detect a fault in either of the two generators and remove it immediately from operation. The advantage over the previous scheme is that the switching system is protected against faulty pulses since the blocking of the improperly working circuit can be done almost instantaneously by electronic means. Furthermore, when working in parallel, the circuits share the load and no power is wasted.

Even more preferable is a scheme that would eliminate blocking. In it, the parallel working circuits are connected so that if one fails, it does not interfere with the proper functioning of the other.

In all these cases, the output of each circuit must be verified against a standard waveform to permit any discrepancy in the output signal to be detected. If it is accepted that the probability of two parallel-working circuits failing at the same time is very small, the signal of one circuit can be used to control the other, and vice versa. The probability of failure becomes smaller with simpler circuits. Therefore, if one subdivides a complete circuit into subcircuits, each of which is paralleled by a companion and associated with

a fault-detecting and blocking system, the reliability of the whole system will be improved. If one of the subunits is out of order, only a small part of the complete circuit is working without parallel equipment and the chance that another

A simplified block diagram of the generator is given in Figure 1.

All three schemes of reliability described above have been applied, their choice depending on the kind of circuits to be protected.

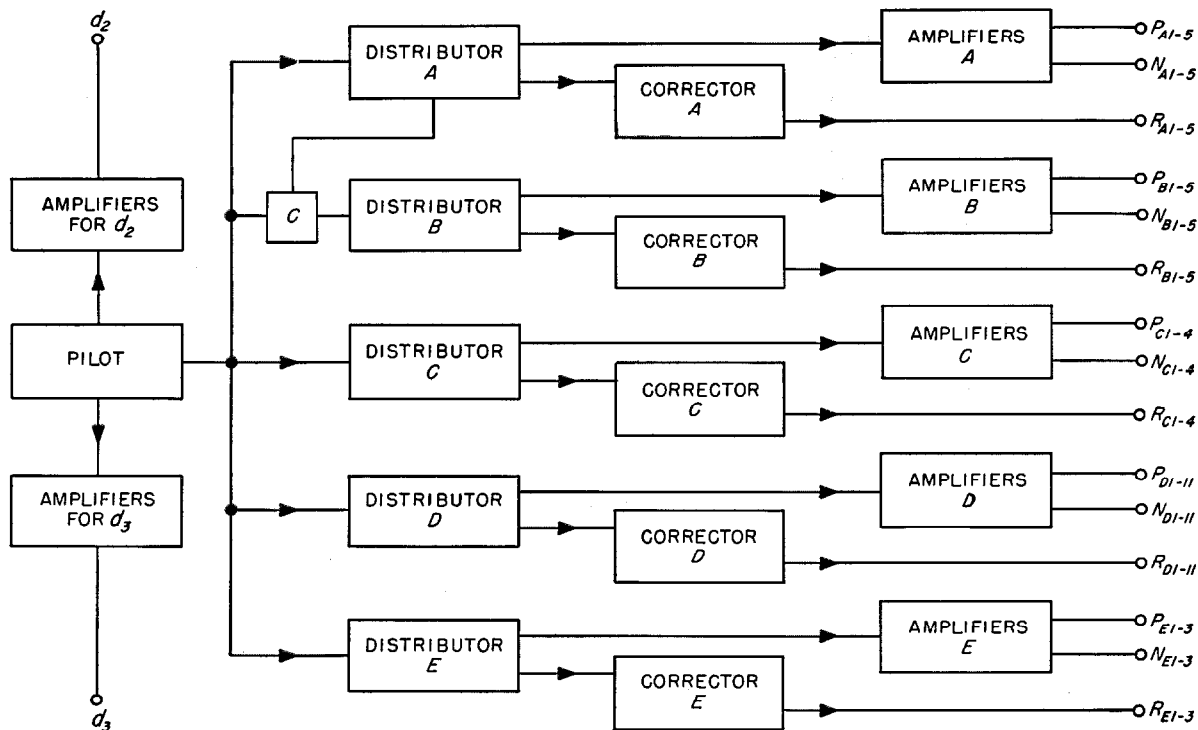


Figure 1—Block diagram of pulse generator.

fault will occur in this place is very small indeed.

Considering this fact, together with arguments of ease of construction and assembly, the pulse generator has been subdivided into 4 major parts.

- A. Pilot circuit, which establishes the time base of the system.
- B. Distributor or pulse-counting circuits, which produce the 5 groups of pulses mentioned in Table 1.
- C. Correctors, which correct the rise and decay times of a special group of pulses that must have short rise and fall times.
- D. Amplifiers, which deliver pulses with the right amount of power to the exchange.

3. Description of Circuits

3.1 PILOT CIRCUIT

The pilot circuit has 4 functions.

- A. To establish the basic recurrence frequency.
- B. To furnish signals to the amplifiers for pulses d_2 and d_3 .
- C. To furnish pulses that are used to trigger the distributor circuits.
- D. To furnish pulses that monitor the operation of the pilot circuits.

To ensure reliability of operation, two pilots are operated in parallel with a common control

that electronically blocks either circuit if it does not deliver a proper signal.

The circuits make use of conventional multivibrators, flip-flops, amplifiers, et cetera so that an explanation of their functions will be based on block diagrams only.

Referring to Figure 2, each pilot is composed of the following parts.

A. Generator Gd_2 delivers pulses of 25-microsecond duration at intervals of 200 microseconds. This is a self-starting multivibrator controlled by a loosely coupled resonant circuit that is adjusted to produce a signal of 5000 cycles per second.

B. Separating or isolating tube S_1 that delivers

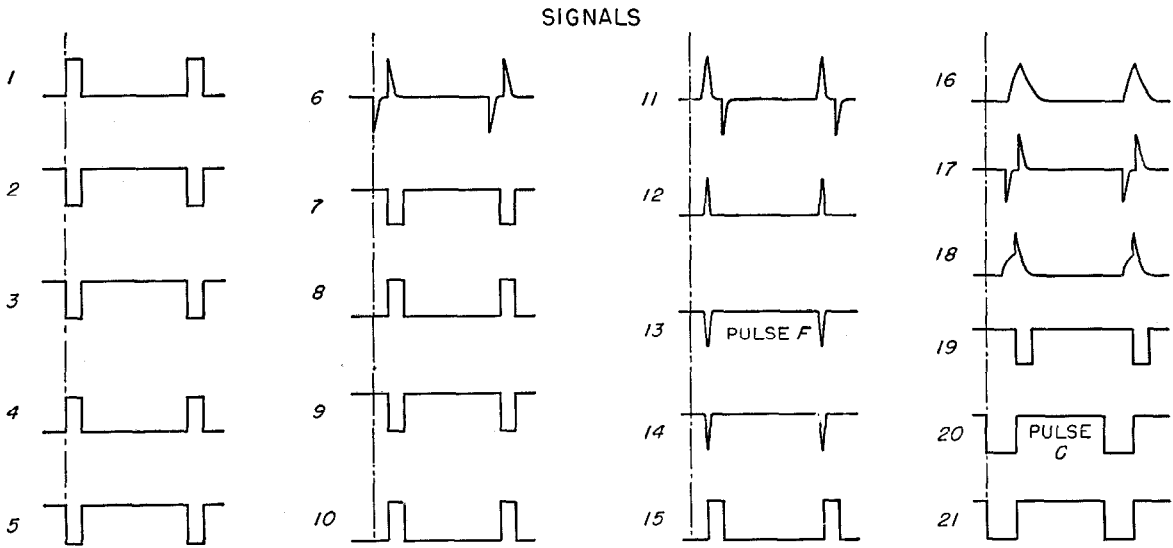
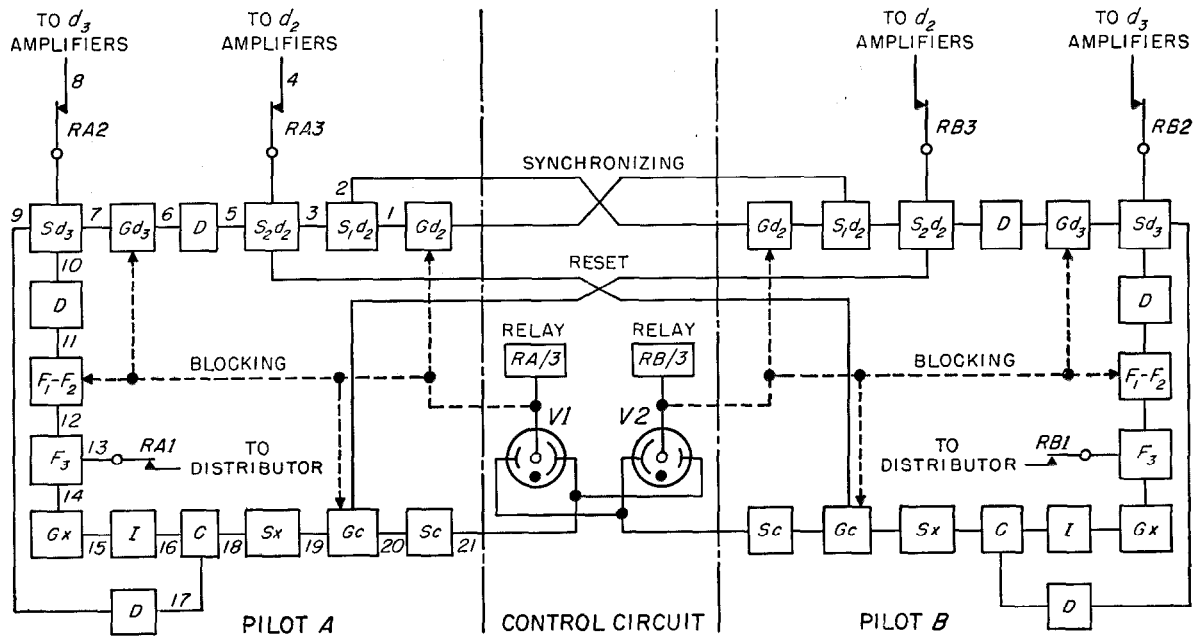


Figure 2—Pilot and pilot-control circuits. C = combination, D = differentiation, $F-F$ = flip-flop pulse generator, G = generation of indicated series of pulses, I = integration, and S = separation or isolation circuit.

negative synchronizing pulses to the other parallel-working pilot.

C. Separating tube S_2 delivering positive pulses to the d_2 amplifiers and a reset signal to the parallel-working pilot for control purposes.

C. Generator Gd_2 is triggered by the differentiated signal coming from S_2 and D ; it sends a negative pulse to a separating tube.

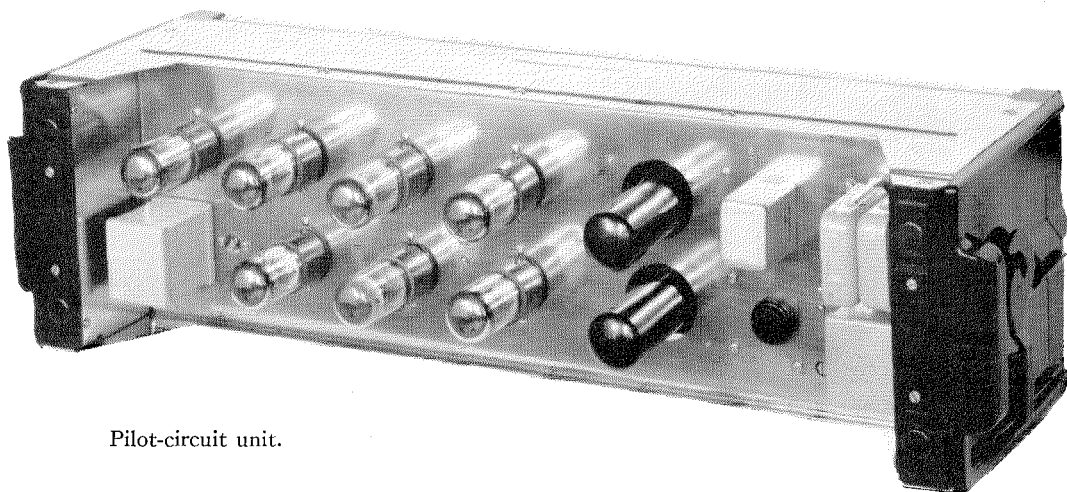
E. Separator Sd_2 sends pulses to the d_2 amplifiers.

F. Generator $F_1 - F_2$ is triggered by a differentiated signal coming from Sd_2 and sends a short signal to the distributor circuits through amplifier F_3 .

seconds. However, it is forced to restore after this time lapse by a signal coming from the S_2d_2 tube of the other pilot and is triggered by the signal coming from tube Sx of its own circuit. The reason for applying forced restoration in Gc will become clear after the description of the control circuit.

J. Tube Sc serves as a separator tube between generator Gc and the control circuit.

This control circuit consists mainly of two cold-cathode tubes, each associated with a relay that is operated when the corresponding tube is fired. The function of the relay is to open all wires connecting a faulty pilot to the remaining part



Pilot-circuit unit.

G. Generator Gx is triggered by F_3 and furnishes pulses that are integrated and combined with differentiated pulses coming from Sd_2 .

H. To open separator tube Sx , it is necessary that both the integrated and the differentiated signals be present at its input and that a proper time relation exist between these signals. This means that if the trailing edge of d_2 comes too early or too late, due to malfunctioning of Sd_2 , the differentiated signal does not ride high enough on top of the integrated pulse and the combined signal is unable to open tube Sx .

I. Generator Gc is a monostable multivibrator with a restoring time greater than 150 micro-

of the equipment so that it may be removed without disturbance and to give an alarm.

Two pulses are applied to each cold-cathode tube. The C pulse coming from pilot A is applied to the cathode of tube $V1$ and to the trigger of tube $V2$, while the C pulse coming from pilot B is applied to the cathode of tube $V2$ and to the trigger of tube $V1$.

The result is that if both pilots are working under normal conditions, the potential difference between cathode and trigger remains insufficient to fire either of the tubes, although both electrodes are regularly making potential excursions. Only if one of the control pulses is absent will the conditions in the corresponding cold-cathode tube be favorable to firing. This also explains why forced restoration has been applied on Gc .

If, for instance, this tube were a one-shot multivibrator, it would be possible due to small changes in its components for the control pulse delivered by it to be shorter or longer than that delivered by the other pilot. If this difference becomes large enough, one of the cold-cathode tubes would be actuated. It could be the one corresponding either to the faulty pilot or the good pilot, since there is no telling which control pulse is too long or too short.

It is interesting to examine some of the major faults that could occur.

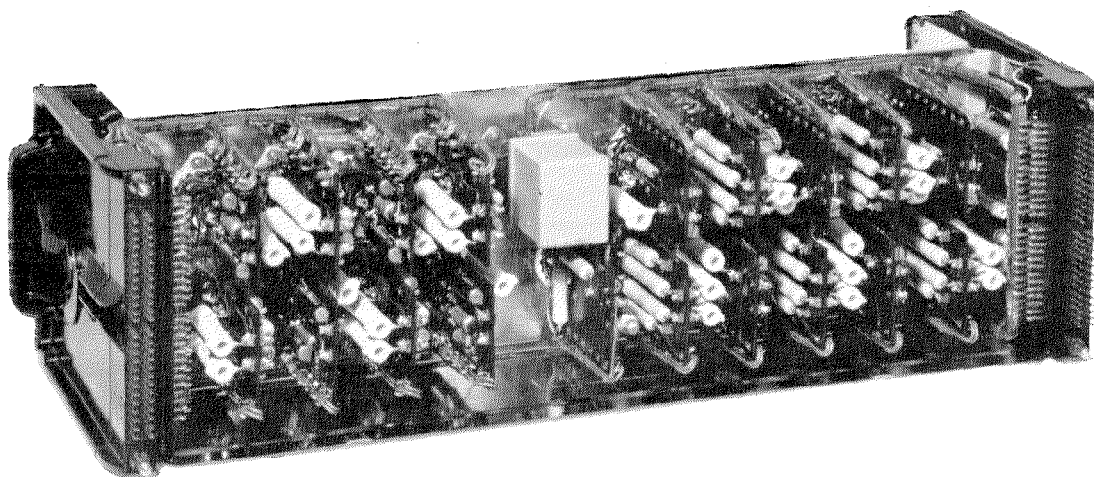
A. Generator $F_1 - F_2$ stops working. Signal 12 is not produced and thus signal 16 is not available to combiner C. Since Sx cannot be opened by

Blocking is exerted immediately on all crucial points of the defective pilot, so that no faulty pulses may be sent to the following circuits.

The electronic blocking system makes use of a point in the control circuit where an important voltage drop is observed when the control is operated. This low voltage is applied through rectifiers on the grid circuit of the most important tubes in the pilot, so that these tubes are unable to emit signals.

A system of push-buttons permits the intentional firing of a control, as well as the extinguishing of the tube after a faulty unit has been repaired and replaced.

The circuit arrangement is such that the tubes cannot deionize and the operated relays cannot



Rear view of a b -group distributor. The plug-in contacts may be seen at each end.

signal 17 alone, the cold-cathode tube in the corresponding control is operated.

B. Length of pulse d_s is shortened or lengthened. Signal 17 comes too early or too late with respect to signal 16 and the combined signal 18 is not large enough to open tube Sx , with the same result as above.

C. Generator Gd_s stops working. Signal 7 is absent and again the corresponding cold-cathode tube is fired.

D. Generator Gd_s is stopped. None of the pulses is present and again the control is actuated.

In each case, the cold-cathode tube is fired in the same time unit that the failure occurs.

be released unless the circuit is restored to working order.

3.2 DISTRIBUTOR CIRCUITS

The purpose of the distributor circuits is to generate the 5 groups of pulses shown in Table 1. Each distributor has two identical units working in parallel under the supervision of a common control.

Each of the two units receives trigger pulses from both pilots, so that even if one pilot is out of order all of the distributor units continue to function.

A distributor is a circuit of the well-known ring counting chain type having as many flip-flops in a closed loop as there are sources in a cycle, for example, 5 for cycle a , 11 for cycle d , et cetera.

The four units delivering 200-microsecond pulses for cycles *a*, *c*, *d*, and *e*, are directly triggered by pulses coming from the pilot circuit through amplifier *F*₃ and have the waveform of signal 13 in Figure 2. Since these trigger pulses are produced every 200 microseconds, the output pulses are automatically of the correct length as each output pulse is derived from one of the flip-flop circuits during the time between adjacent trigger pulses. The counting chains are arranged to have the required stepping cycle to provide the number of sources required.

Cycle *b*, which has to deliver pulses of 1000-microsecond duration, is triggered by the same signal as the other chains. After the 1st trigger pulse has passed, a gating circuit controlled by an *a*-cycle pulse prevents the 4 following trigger pulses from actuating the flip-flop, which holds until the 6th pulse of the *a* cycle arrives. Since an *a*-cycle pulse is present only once every 5 time units, a trigger pulse actuates the *b*-cycle units only once every 1000 microseconds.

Each flip-flop is followed by two separator tubes, which are cathode-followers. The purpose

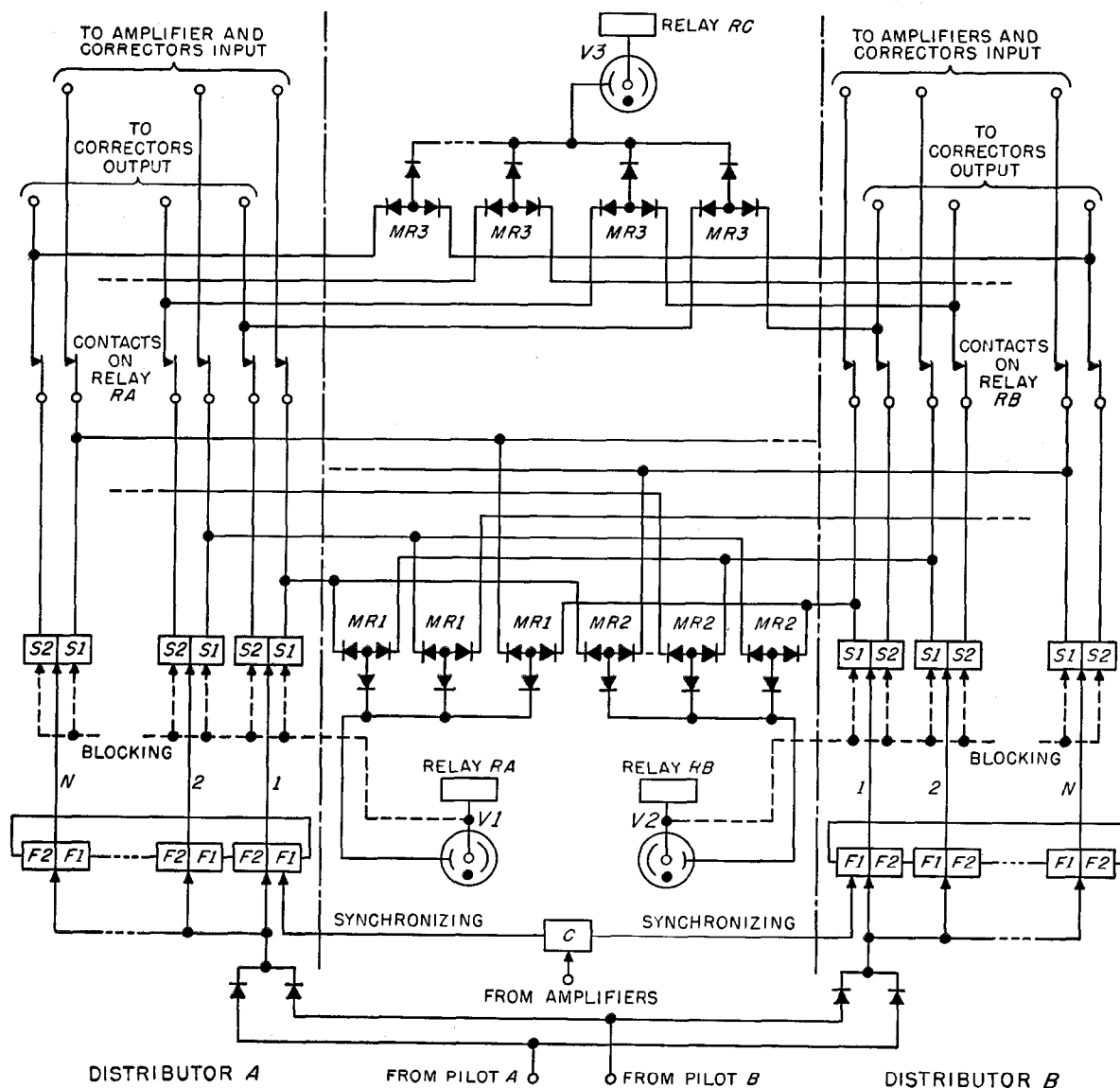


Figure 3—Two distributors are operated in parallel under the supervision of a common control circuit.

of these tubes is principally to act as buffers between the counting chains and the other circuits of the generator.

In Figure 3, two parallel-working distributor chains are represented with their control circuit. Two kinds of controls are introduced into the distributors; the first is applied to the counting chains proper and the second to the separator tubes.

A major fault that could occur in a distributor would be for the counting chain to stop stepping around. In that case, one of the flip-flops will be in the actuated state while all the others are in the nonactuated condition. This effect is inherent in the design of the counting chains. If, therefore, a chain does not continue to step, it may be accepted that one of the outlets will always be at a much higher potential than all others.

For control purposes, each pulse circuit of distributor *A* has been associated with the pulse circuit of following time rank produced by distributor *B*, and vice versa. The circuits are isolated from each other by the networks of rectifiers *MR1* and *MR2* that are connected to the triggers of the cold-cathode tubes *V1* and *V2* that can disable the distributors.

When the two chains are working under normal conditions, no signal appears on the triggers of the control tubes since one side of each rectifier group *MR1* and *MR2* is always at a low potential. For instance, when *S1* on point 1 of distributor *A* is high, the other side of *MR1* is low, since no pulse is delivered on point 2 of distributor *B* at that instant. It is only if one of the flip-flops in a chain does not turn over, that both sides of a rectifier group *MR1* or *MR2* will be high at the same time and the corresponding cold-cathode tube will be fired.

If, for instance, distributor *A* stops working when its outlet 1 is high, the control tube of this distributor will be triggered in the following time unit when the pulse from distributor *B* appears at the other branch of the rectifier combination *MR1*. This operates *V1* and disables the faulty distributor *A*. If it is distributor *B* that stops on one of its points, a pulse of following time rank from distributor *A* appearing on the second branch of *MR2* will cause the proper cold-cathode tube to light and disable distributor *B*.

Blocking is exerted on all separators of the

faulty chain, so that no signal will be emitted by this chain. At the same time, a relay is operated to open all outlets and give an alarm. Blocking is done by lowering the potential of the grids of the separators, so that they are brought below cutoff.

The control of the separator stages has been accomplished by grouping by pairs the outputs of tubes *S2* of the same rank and by strapping these groupings through separating rectifiers *MR3* to the trigger of the cold-cathode tube *V3*. In this way, the potential observed at this trigger will be equal to the top of the pulses since both sides of each grouping are always high at the same time. This is repeated for each grouping in turn so that there appears on the common point a signal that is equal to all pulses placed side by side. If, however, a separator fails to work, one side of a group of *MR3* will be low and a gap is produced in the signal on the common point. Tube *V3* is fired and relay *RC* is operated, giving an alarm. No blocking is exerted since a failing separator cannot influence the parallel-working circuits, all of which are equipped with separating rectifiers.

To keep both counting chains in step, a synchronizing signal is injected on the first flip-flop of each chain. This signal is obtained from the corresponding amplifier stages after proper transformation in circuit *C*.

Here again a push-button system permits the intentional blocking and unblocking of a distributor.

3.3 CORRECTOR CIRCUITS

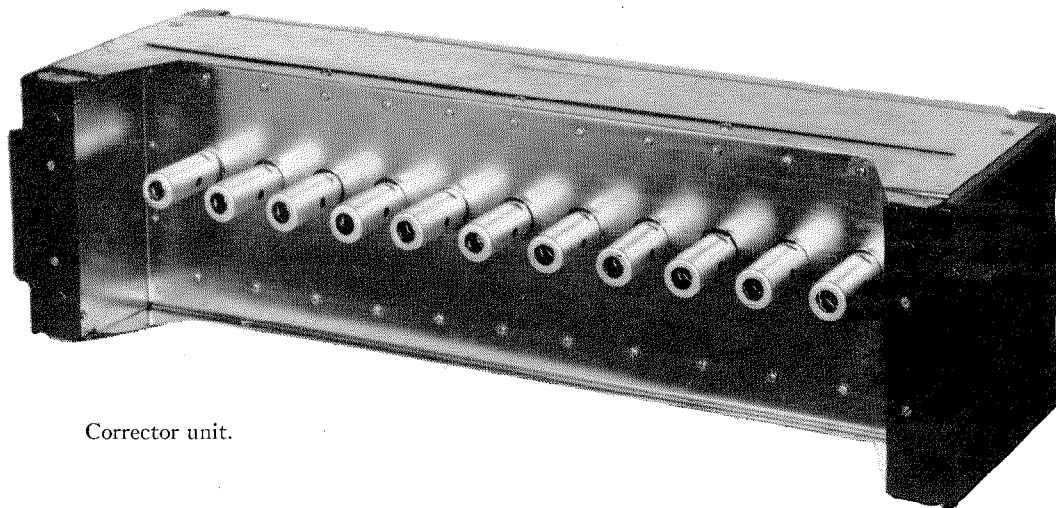
The purpose of these circuits is to produce the *R* type of pulses for which the requirements regarding rise time and decay are much more exacting than for the other pulses. The power with which these pulses should be delivered is negligible so that they are taken directly from the second separator tubes in the distributor circuits and sent to the different points in the exchange where they are required. This means long wires, which present the problem of distributed capacitance and the correction of the waveform distortion that it produces.

The principle on which the correction of these pulses is based consists in the forced charge of the wire capacitance at the beginning and the

forced discharge of this wire at the end of a pulse. To effect these charges and discharges, use is made of thyratrons that serve as ultrafast switches while capacitors are used to store the charges that have to be injected on the wires.

Figure 4 shows a schematic drawing of a 3-point corrector, without its control circuit.

resistor $R2$, tube $V2$, lead 2, and wire capacitance $C2$ to ground. Resistor $R2$ limits this discharge to a value suitable for the tube. The result is that the capacitance of wire 1 is discharged while that of wire 2 is now fully charged. The operation repeats at the end of the pulse on wire 2 in favor of wire 3, and continues around the chain.



Corrector unit.

The distributed capacitances of the wires connected to the pulse leads are represented by $C1$, $C2$, and $C3$, while the storage capacitors are $C1'$, $C2'$, and $C3'$.

The pulses coming simultaneously from distributors A and B are connected to the 3 R -pulse leads via separating rectifiers $S1$ and $S1'$.

The storage capacitors are connected between the pulse leads through thyratrons that are triggered by pulses coming from the distributors after they have passed through a pulse-forming circuit. The thyatron is thus triggered by a signal coming from the distributor at the moment another pulse from the distributor appears on the corresponding pulse lead.

Let us suppose that a pulse is present on output $VS2-1$. This lead is, therefore, at about -50 volts since this is the top value of the R pulses. Rectifier $S2$ is blocked on that point while all other leads are at -110 volts since their rectifiers $S2$ are passing current.

When the pulse on lead 1 stops, the next pulse starts on lead 2 and at the same time $V2$ is triggered. With the thyatron conducting, a discharge path is established from the wire capacitance $C1$ through lead 1, storage capacitor $C1'$,

The scheme used to insure reliable operation in the pilot and distributor circuits is not applicable here. The line capacitance is really a part of the circuit and this makes it impracticable to apply a parallel-working principle.

There are, therefore, two corrector units for each cycle, one of which works into the live load while the other is connected to a dummy load.

If the first circuit develops a fault, the live load is switched in place of the dummy load at once, while if the standby circuit misfunctions, only an alarm is given to call attention to the trouble.

The control circuit comprises a fault-detecting system consisting of rectifiers and cold-cathode tubes, and controls both the top and the bottom of the pulse train coming out of each circuit. Again push buttons are provided for intentional firing and deionizing of the control tubes.

3.4 POWER AMPLIFIERS

The P and N pulses must deliver a certain amount of power to the exchange. The entire equipment is designed to produce more power for each outlet than necessary so that in case of

failure of some apparatus the power delivered will not be less than the minimum amount required to drive the exchange.

To satisfy this requirement, an amplifier was constructed that could deliver a unit amount of power. This amplifier was kept as simple as possible, comprising only one tube, an output

amplifiers. The output transformers have two secondary windings to permit the base of the pulses to be placed at the correct value; that is, ground for P pulses and -40 volts for N pulses. Resistor $R3$ permits adjustment of the height of the pulses.

The control circuit for an amplifier comprises

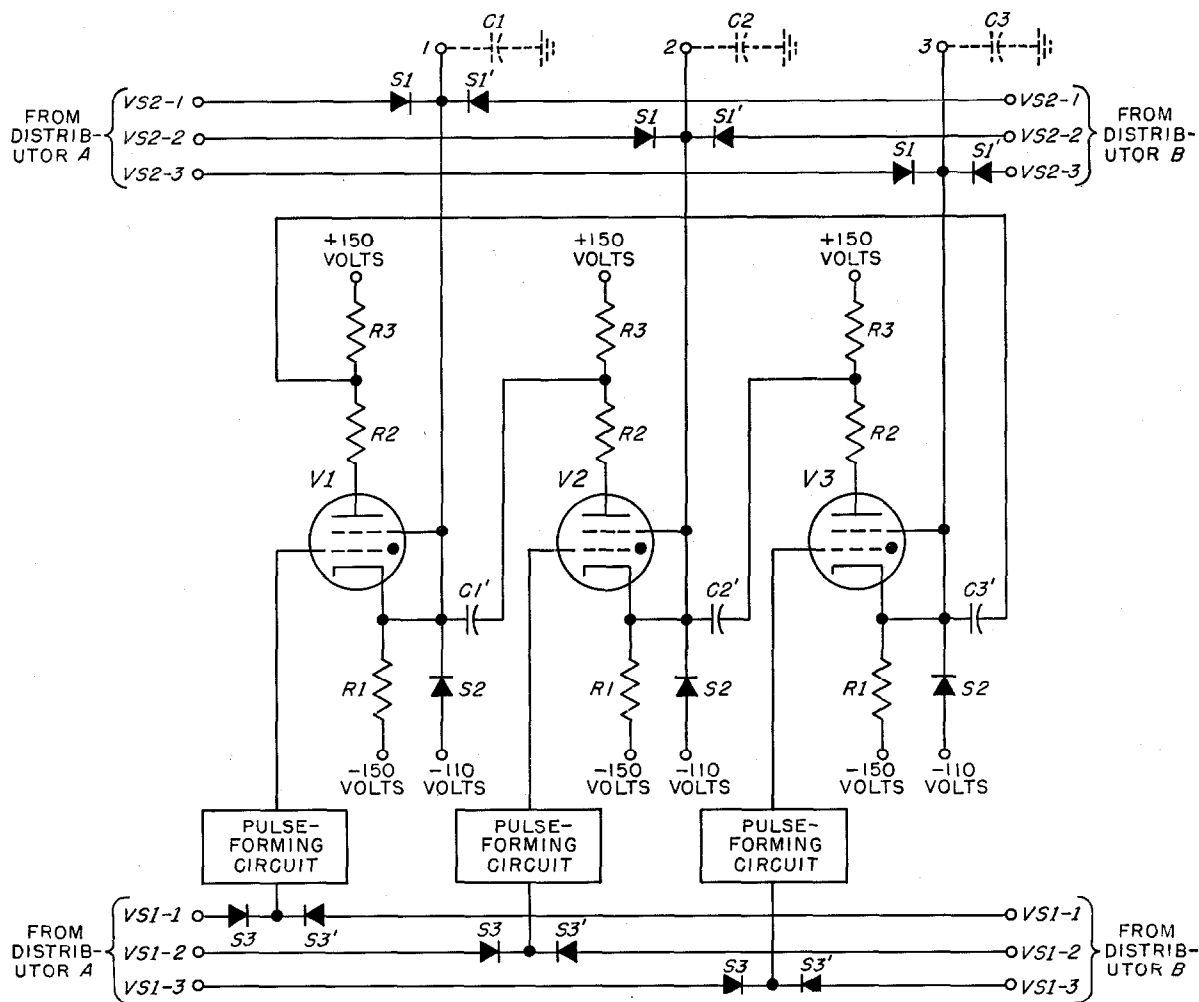


Figure 4—Three-point corrector.

transformer, and some associated small components.

On each output lead are placed as many of these unit amplifiers in parallel as are required for the amount of power to be delivered, plus one additional amplifier. In case of failure of one of the amplifiers, there will still be enough in working condition to carry the load.

As may be seen from Figure 5, the P and N pulses are obtained at the output of the same

rectifiers $S1$ and $S2$ and resistors $R1$ and $R2$. When no pulses are delivered, the output lead for N pulses is at -40 volts, while the output lead for P pulses is at ground potential. These potentials are impressed on control wires $C1$ and $C2$ through rectifiers $S1$ and $S2$, which are conducting.

When a pulse arrives, both control wires are raised by 16 volts as long as the pulse is present, returning to their original values at the end of the pulse. In other words, the control pulses are a

reproduction of the *P* and *N* pulses. All pulses from a cycle are brought to a common control circuit, by means of separating rectifiers *S5* and *S6*, so that the input of the control transformer *TC* sees a constant voltage that is equal to the top value of the pulses.

If, however, one of the amplifiers does not deliver signals on one of its two outputs, *N* or *P*, the corresponding control wires *C1* or *C2* are kept at -40 volts or ground by means of rectifiers *S1* or *S2*, so that the control pulses that are normally delivered by the parallel-working amplifiers do not appear on the control wires.

Here again each amplifier receives input pulses from two distributors so that even with one of these circuits out of order, pulses continue to be delivered normally to the exchange.

4. Construction of Pulse Generator

4.1 COMPONENTS

Reliability again has been the major factor in choosing the components of the pulse generators.

All resistors and capacitors have been selected to have large safety factors regarding power dissipation and working voltages. Possible faulty

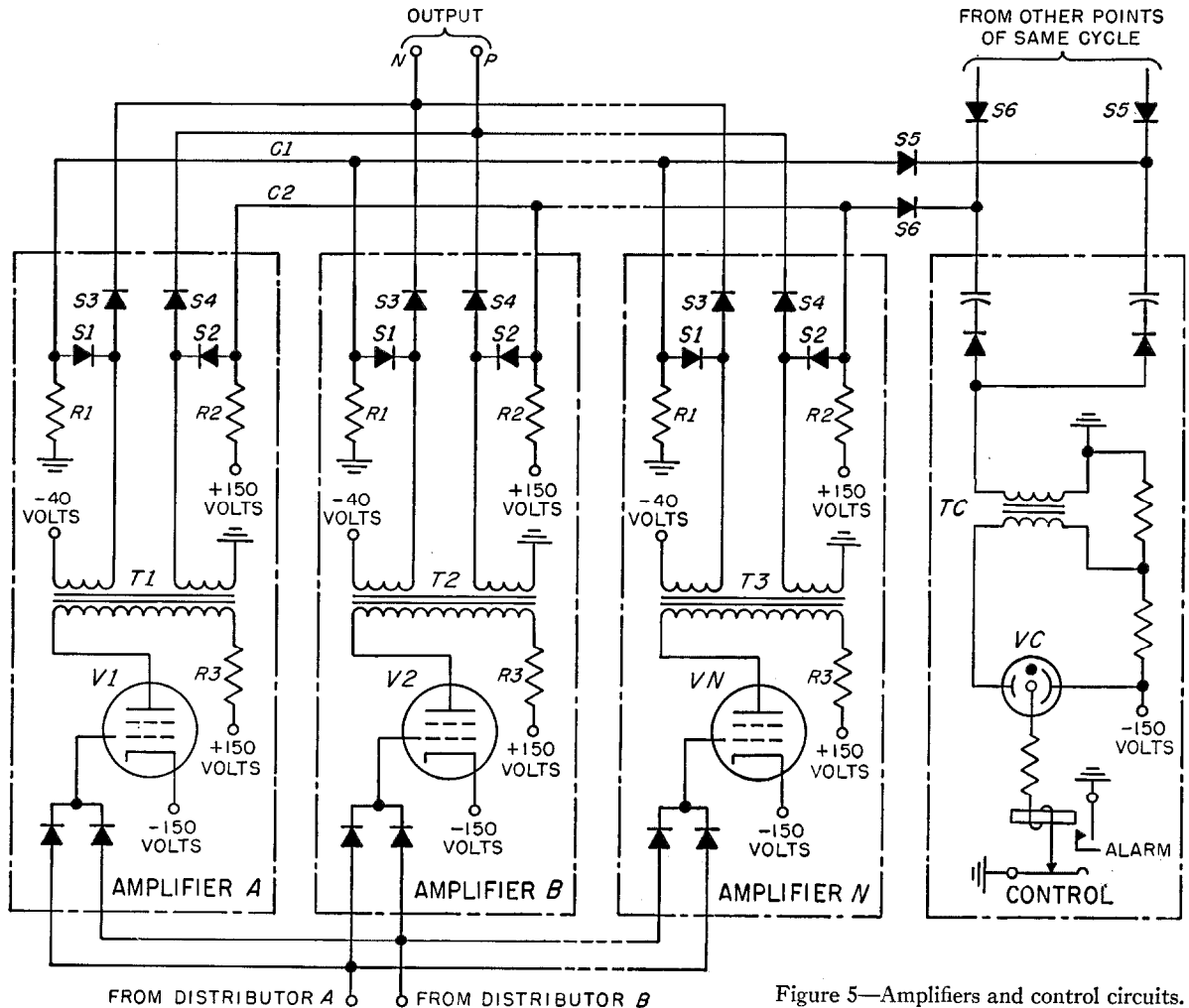


Figure 5—Amplifiers and control circuits.

The result is that one pulse of the train applied to transformer *TC* is missing, and tube *VC* fires. An alarm is given but no blocking is exerted since a defective unit is separated from the others by means of rectifiers *S3* and *S4*.

conditions have been considered in determining these values, so that failure of one component will not subject other parts to excessive loads and require their replacement also.

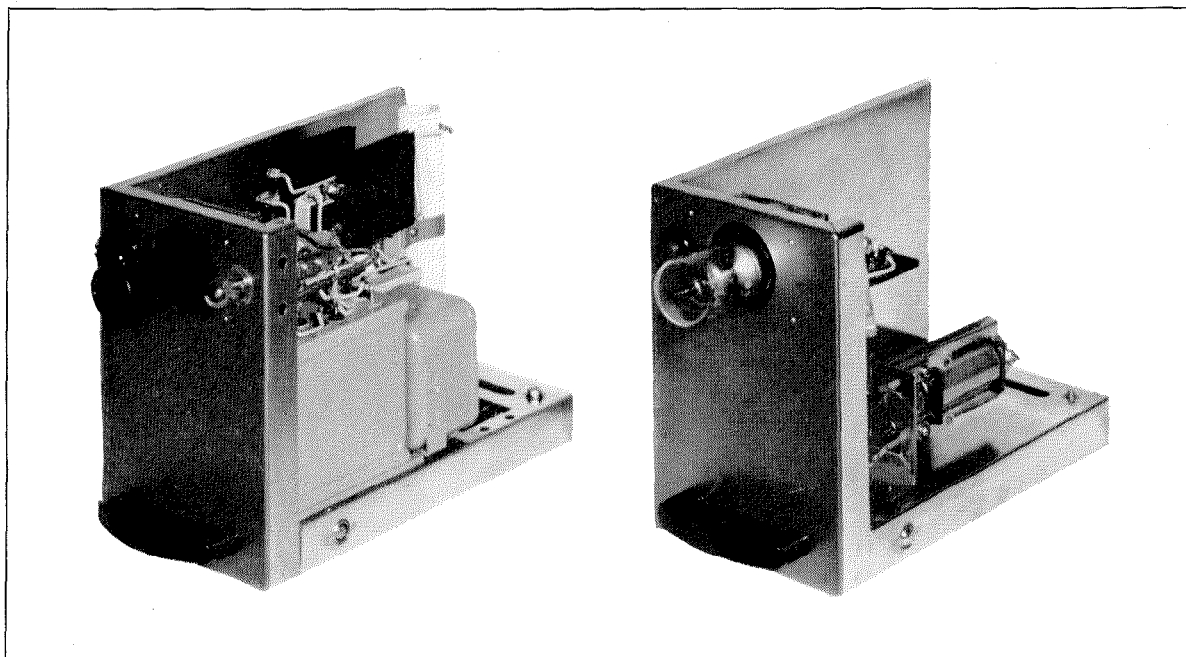
The number of germanium rectifiers has been

kept to a minimum because of the rather high price of this component. Most of the rectifiers are of the miniature selenium type and generous safety factors have been applied regarding the voltage each disk must withstand.

Four different types of tubes are used in the generator. The *13D1* Brimar tube manufactured by Standard Telephones and Cables was selected

plug-in units that are noted for having very reliable high-pressure contacts.

In each individual unit, all the tubes are placed on the front side with their sockets mounted on a plate that separates the tube envelopes from the components in the back; in this way heat transfer to these components is kept to a minimum.



Amplifier unit and its control system.

as the twin triode principally because with these tubes the contacts of the jack-in units do not have to carry heavy currents, since a 25-volt filament supply is required.

Ordinary *6V6* tubes were chosen for the power stages, while in the pilot circuit, a few *6L6*'s are used; both of these tubes are highly reliable when applied with proper care.

The thyratrons in the corrector circuits are of the *2D21* type. Although of miniature size, they are able to pass the required peak currents. All tubes are subjected to a short life test to eliminate early failures.

4.2 GENERAL LAYOUT

The complete pulse generator comprises three bays of equipment. Two of the bays are of the type used in the *7E* system and employ the same

The amplifier bay houses about 100 unit amplifiers. It is constructed along the same lines as the switch bays used elsewhere in the exchange.

The bays are enclosed by doors and air is forced to flow from the bottom to the top. This is necessary to keep the ambient temperature in which the components are working at a normal value. Cooling the vacuum tubes considerably lengthens their useful life.

The air intake is outside of the exchange building. A filter is included in the airflow system so that all dust particles are eliminated.

4.3 POWER SUPPLY

The voltage sources to the generator are under the control of timing circuits for the purpose of applying these voltages in the correct order.

To reduce the danger of cathode damage, high voltages are only connected when the filaments

of the tubes are warm. The control circuits are the last to be energized so that they are unable to work during the starting interval of the generator.

An interruption of the primary power supply causes all high potentials to be removed at once. They are again applied in the proper sequence after the supply is re-established. The time lapse for the applications of the different voltages is directly related to the duration of the interruption. If it is a very short one, the tubes have had practically no time to cool off and the high potentials are reconnected very rapidly whereas with longer interruptions the timings are automatically lengthened.

The filament transformers are so connected that even with one of them out of order, there is still at least one of each type of unit, namely:

pilot, distributor, and corrector, that continues to work normally. As for the amplifiers, the filaments are so distributed over the transformers that even with one out of order there are enough amplifiers left on each point to supply the exchange.

5. Conclusion

The pulse generator is a major part in the mechano-electronic exchange and the slightest fault in pulse delivery will have immediate consequences in the functioning of the exchange.

It is, therefore, understandable that some time and effort were devoted to the engineering and construction of this circuit, with absolute reliability of operation as a primary objective. It is felt that this goal has been achieved.

Recent Telecommunication Development

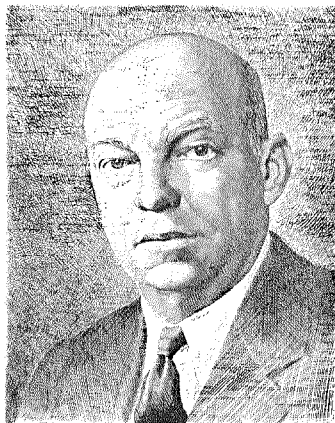
Etching of E. H. Armstrong

EDWIN HOWARD ARMSTRONG (1890-1954) is portrayed in the latest of the series of etchings published by the International Telecommunications Union. Born in the city of New York, he was educated at Columbia University with which he was closely associated until his death.

Major Armstrong contributed four outstanding inventions. His regenerative circuit of 1913 supplied a much-needed increase in sensitivity and selectivity to the early vacuum-tube detector. In 1918, he produced the super-heterodyne that is the basis of practically all modern radio receivers. It allows unlimited increases in selectivity and sensitivity without impracticable multiplicity of controls or instability. It makes the noises developed in the circuits preceding the amplifier the real limitation on over-all gain. Two years later, superregeneration promised to supplant all other methods of reception, particularly at the higher frequencies, but some inherent limitations

restricted its applications severely. In 1933, his method of wide-band frequency modulation provided the first practical system of eliminating the effects of static; it makes the acoustical output of thunder more damaging to reception of music during a local thunderstorm than the electrical disturbances of the lightning that produces the thunder.

The etching of Armstrong is the twentieth in the series that was started in 1935. On a good grade of paper measuring 9 by 6 $\frac{5}{8}$ inches (23 by 17 centimeters) including margins, these etchings are available at 3 Swiss francs each from Secrétariat général de l'Union internationale des Télécommunications, Palais Wilson, 52, rue des Pâquis, Genève, Suisse. The entire series is com-



prised of etchings of Ampere, Armstrong, Baudot, Bell, Erlang, Faraday, Ferrié, Gauss and Weber, Heaviside, Hertz, Hughes, Kelvin, Lorentz, Marconi, Maxwell, Morse, Popov, Pupin, Siemens, and Tesla.

United States Patents Issued to International Telephone and Telegraph System November, 1954 — January, 1955

UNITED STATES patents numbering 18 were issued between November 1, 1954 and January 31, 1955 to companies in the International System. The inventors, titles, and numbers of these patents are given below; summaries of several that are of more-than-usual interest are included.

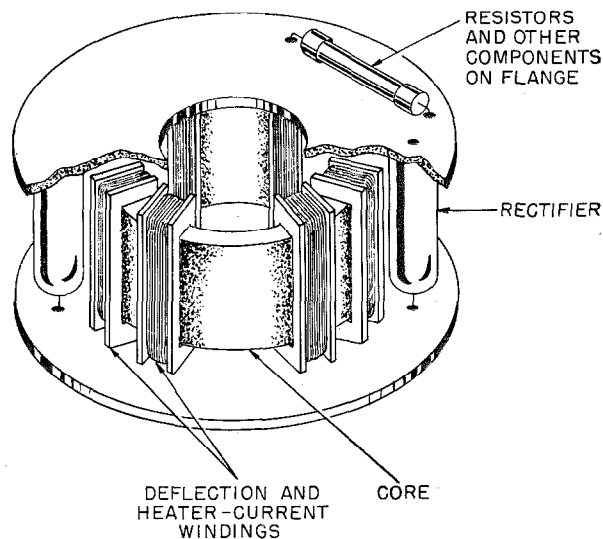
- W. A. Billings and Z. Brzozowski, Joining Wire, 2 696 227
- A. Brown and D. A. Romlin, Electric Switch, 2 696 553
- G. Deakin, Telephone System, 2 696 525
- M. den Hertog and C. de Zeeuw, Selection Systems for Electrical Circuits or Equipment, 2 694 751
- M. den Hertog, System of Identification of the Calling Line in Telephone Systems, 2 694 753
- M. J. DiToro, W. Graham, and B. M. Dwork, Fundamental-Pitch Detector System, 2 699 464
- J. H. Eisele and A. R. Geisselsoder, Dry-Disk Rectifier Assembled from Unperforated Rectifier Plates, 2 698 918
- R. Haberkorn, Pneumatic-Tube System, 2 698 722
- P. L. Hancock, Sideband Generator, 2 697 220
- J. V. Martens, Voice-Frequency Receiver, 2 698 878
- S. E. Mayer and H. F. Sterling, Electric-Current Rectifier, 2 695 380
- H. G. Miller and J. Stillman, Automatic Tuning Device, 2 699 504
- W. Reinhard and A. Rappold, Anode-Voltage Supply for Television Receivers, 2 694 784
- W. Sichak and H. A. Augenblick, Jr., Tunable Band-Pass Filter, 2 697 209
- F. H. Stieltjes, Balancing of Carrier Cables, 2 696 526

- V. J. Terry, R. H. Dunn, and P. G. Collier, Remote Control and Supervision System, 2 694 802
- E. P. G. Wright, D. S. Ridler, D. A. Weir, and J. Rice, Telecommunication Exchange, 2 694 752
- E. S. Wright and A. C. Robertson, Piezoelectric Crystal Plate, 2 698 909

Anode-Voltage Supply for Television Receivers

W. Reinhard and A. Rappold
2 694 784—November 16, 1954

A deflection yoke for television tubes that utilizes a toroidal magnetic core with the deflec-

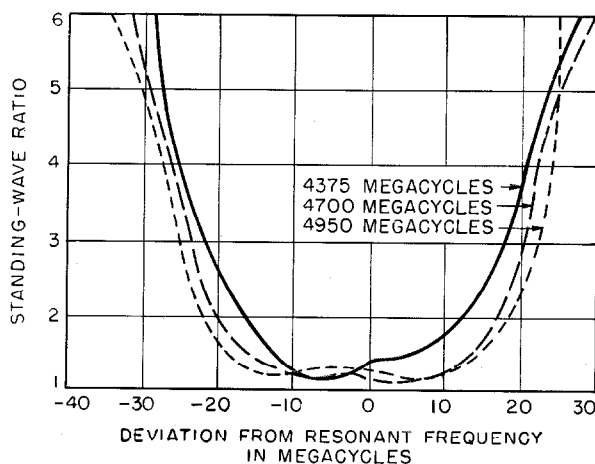
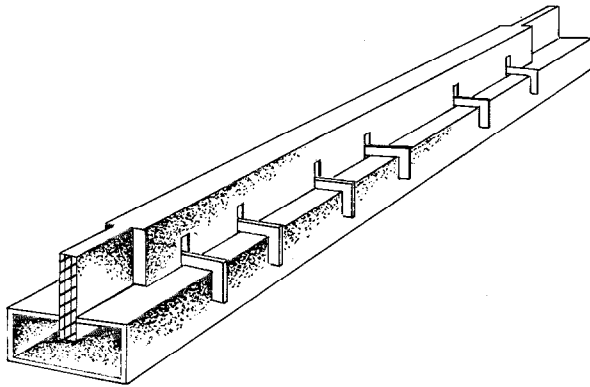


tion windings magnetically coupled to auxiliary coils also wound on the core for supplying the heating current for the rectifiers used to obtain the cathode-ray-tube high voltage from the deflection waveforms. These rectifier tubes are mounted in the same yoke structure and since the heating current is obtained at this high-voltage point, it is not necessary to provide special insulation for them.

Tunable Band-Pass Filter

W. Sichak and H. A. Augenblick, Jr.
2 697 209—December 14, 1954

A tunable waveguide-type microwave filter in which the filter is divided by irises into sections



and all of these sections are tuned simultaneously by the insertion of an insulating strip into the waveguide along its entire length.

Balancing of Carrier Cables

F. H. Stieltjes
2 696 526—December 7, 1954

A method of balancing high-frequency cables to compensate for phase disturbance causing

crosstalk in which the conductors of successive sections of the cable are systematically exchanged in groups, pairs of groups being transposed with each other on adjacent sections of the cable.

Electric-Current Rectifier

S. E. Mayer and H. F. Sterling
2 695 380—November 23, 1954

A rectifier formed of titanium dioxide of the conducting type having on its upper surface a thin layer of nonconducting titanium dioxide. The base-plate electrode is silver on the surface of the conductive titanium dioxide and the counterelectrode deposited on the nonconductive titanium dioxide may consist of lead foil, tellurium, or may incorporate thallic oxide.

Pneumatic-Tube System

R. Haberkorn
2 698 722—January 4, 1955

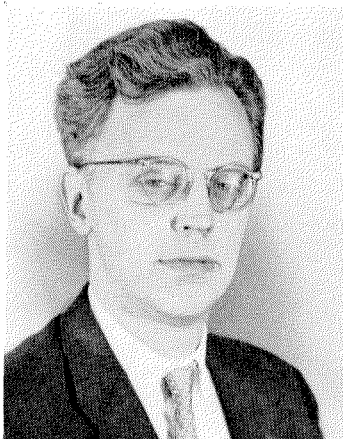
A deflection switch is described for pneumatic-tube carriers in which there is provided an arrangement for simultaneously shunting the carrier into a desired branch tube and also controlling the flow of air by means of a damper.

Fundamental-Pitch Detector System

M. J. DiToro, W. Graham, and B. M. Dwork
2 699 464—January 11, 1955

A system for producing pulses repeated in accordance with the fundamental frequency of various voice sounds. The voice sounds are rectified; amplitude discrimination is incorporated and band-pass filters determine the fundamental frequency of the sound. This is accomplished by various switching and storage systems and is followed by a differentiating circuit for producing pulses timed in accordance with the fundamental frequency.

Contributors to This Issue



ROBERT T. ADAMS

ROBERT T. ADAMS was born in Sparkill, New York, on January 6, 1915. He received his engineering education at Cornell University.

From 1935 to 1937, he did engineering work for the National Rayon Dyeing Company. For the next eight years, he was with the Western Electric Company.

Since 1946, Mr. Adams has been associated with Federal Telecommunication Laboratories and is now a department head. His work on air navigation, communication receivers,

microwave links, and radar has resulted in several patents. He is coauthor of a paper in this issue on reflectometers.

Mr. Adams is an Associate Member of the Institute of Radio Engineers.

• • •

REGINALD B. W. COOKE was born in London, England, on May 17, 1913. He joined Standard Telephones and Cables, Limited, in 1929 and until 1943 was a production-planning engineer.

After service with the Royal Corps of Signals, he joined Standard Telecommunication Laboratories in 1946 and worked with A. H. Reeves on a number of projects that included gas counter valves and transistor and other semi-conductor devices. He is co-author of the paper in this issue on a positive-gap diode.

• • •

J. L. DONCEEL was born in Antwerp, Belgium, on August 12, 1899.

He joined the apparatus division of Bell Telephone Manufacturing Company in 1920.

He was engaged successively in the inspection, investigation, and development of automatic switches.

In 1938, Mr. Donceel was placed in charge of the life-test department and in 1945 of the design of automatic apparatus. He was appointed head of the apparatus division in 1953. He reports here on a new step-by-step selector switch.

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ALEXANDER HORVATH was born in New York City on December 2, 1911. He received a bachelor's degree in



J. L. DONCEEL

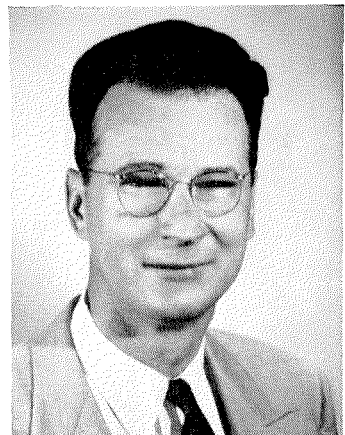
physics from Antioch College, where he participated in the spectroscopic analysis of the chlorophylls and their derivatives.

After a short period of instructing at Antioch and later as an instructor and field engineer with the United States Army Signal Corps, he joined Federal Telecommunication Laboratories in 1944. He is now in charge of a components department of the guided-missile laboratory and is a coauthor of a paper on reflectometers in this issue.

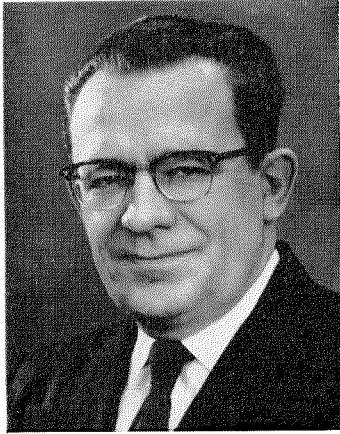
Mr. Horvath is a Senior Member of the Institute of Radio Engineers.



REGINALD B. W. COOKE



ALEXANDER HORVATH



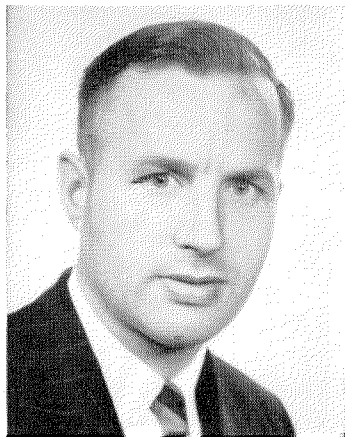
JOSEPH F. HOUDEK, JR.

JOSEPH F. HOUDEK, JR. was born on May 10, 1908 in Chicago, Illinois. He received a BS degree in physics and chemistry from Lewis Institute in 1934.

From 1934 to 1937 as a consulting engineer and orchestra leader, he was concerned with crystal microphones as orchestral pickup devices and the design of associated amplifiers. From 1937 to 1941, he was with the General Electric X-ray Corporation.

In 1941, Mr. Houdek joined the laboratories of Kellogg Switchboard and Supply Company, becoming chief of the general laboratory in 1954. In this issue, he describes the design and performance of capacitor microphones.

Mr. Houdek is a registered professional engineer in Illinois, a member of the American Institute of Electrical Engineers and active in its Chicago Section, a member and officer of the Chicago Acoustical and Audio Group,



JOHN W. HOWES

and holds memberships in the Armed Forces Communication and Electronics Association and in the Physics Club of Chicago. He has been granted several patents on wire-communication apparatus. In addition to his engineering work, he has been an active musician for over 30 years.

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JOHN WESLEY HOWES was born at Neutral Bay, Sydney, Australia in 1916. He attended Sydney Technical College.

After doing radio receiver service work he joined the Australian army in 1940, transferring to the Royal Australian Air Force and attaining the rank of Flight Lieutenant.

Late in 1945, he became a member of the engineering staff of the valve laboratory of Standard Telephones and Cables Pty., Limited, where he is concerned with the manufacture of transmitting valves. He reports in this issue on thermistors.

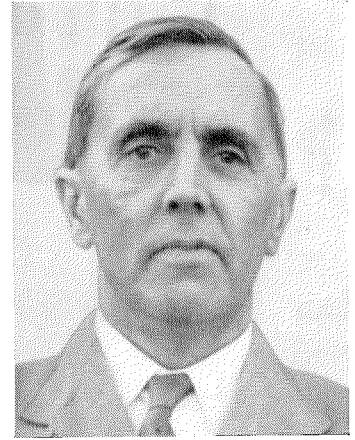
Mr. Howes is a member of the Institution of Radio Engineers, Australia.

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ALEC HURLEY REEVES was born in Redhill, Surrey, England, in 1902. He received the DIC from Imperial College of London University and is also an associate of the City and Guilds Institute. He is a member of the Institution of Electrical Engineers.

In 1923, he joined the International Western Electric Company, which later became Standard Telephones and Cables, Limited. He worked on the original transatlantic radiotelephone system; and later, while in the Paris laboratories, on the Madrid-South America high-frequency link and also on the microwave system across the English Channel. He developed the first single-sideband high-frequency radio system. He pioneered in pulse methods, inventing pulse-width and pulse-time modulation systems, and holds the patent on quantized pulse systems (pulse-code modulation). He also introduced the use of flip-flop circuits for frequency division. The multipoint gas counting tube was another of his inventions.

During the war, he went to the Telecommunication Research Estab-



ALEC HURLEY REEVES

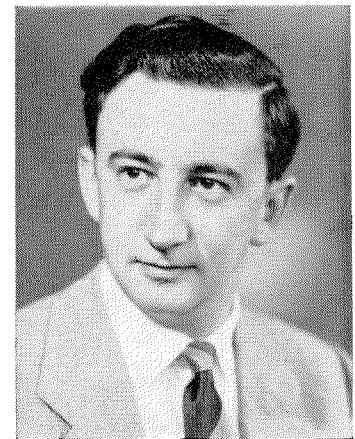
lishment on radar work and was responsible for the development of the Oboe system, for which he received the award of the Order of the British Empire in 1945.

Mr. Reeves is now in charge of a division at Standard Telecommunication Laboratories, where he has worked on electronic switching systems. In the quest for new electronic tools, he developed the special germanium diode that is the subject of a paper in this issue.

. . .

EUGENE VAN DYCK was born in Antwerp in 1924. He received a degree in electrical engineering from the Université de Louvain in 1950.

He then entered the Bell Telephone Manufacturing Company in Antwerp



EUGENE VAN DYCK

After spending a year on a training program in the United States, he returned to Antwerp, where he has been engaged in the application of electronics to switching systems.

He is a coauthor of a paper on an electronic pulse generator for the δA mechano-electronic telephone switching system.

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CAMILLE WEILL was born in Lorient, France, on June 21, 1915. He received the B.S. degree and certificate of mathematics from Paris University and in 1938 the degree of electrical engineer



CAMILLE WEILL

at the Ecole Supérieure d'Electricité in Paris. During the war, he served as an officer in the French Signal Corps.

In 1941, he joined the Laboratoire Central de Télécommunications and in 1951 was named head of the department in charge of electronic switching and industrial applications of electronics.

He was responsible for setting up a laboratory at the French Air Ministry, which took him out of Laboratoire Central de Télécommunications from 1946 to 1949.

Mr. Weill is a coauthor of a paper in this issue on an electronic pulse generator for the δA mechano-electronic telephone switching system.

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Standard Eléctrica, S.A.; Madrid

SWEDEN — Aktiebolaget Standard Radiofabrik; Stockholm

SWITZERLAND — Standard Téléphone et Radio S.A.; Zurich

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CUBA — Cuban American Telephone and Telegraph Company; Havana

CUBA — Cuban Telephone Company; Havana

PERU — Compañía Peruana de Teléfonos Limitada; Lima

PUERTO RICO — Porto Rico Telephone Company; San Juan

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All America Cables and Radio, Inc.; New York, New York
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Mackay Radio and Telegraph Company; New York, New York

ARGENTINA —

Compañía Internacional de Radio; Buenos Aires
Sociedad Anónima Radio Argentina; Buenos Aires (*Subsidiary of American Cable & Radio Corporation*)

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